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Characterization of an Optoelectronic Terahertz Beam System

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Abstract—We describe the performance of an optoelectronic THz beam system. The transmitter operation is based on the repetitive, subpicosecond laser excitation of a Hertzian dipole antenna embedded in a charged coplanar line. With this transmitter electromagnetic beams of 1/2 cycle THz pulses at a repetition rate of 100 MHz are produced. The associated optoelectronic receiver is gated in synchronism with the excitation of the transmitter by subpicosecond pulses from the same laser source. With this receiver, the 10 nW beams of THz pulses were observed with a signal-to-noise ratio greater than 10000:1. Several sources contributing to the noise of the receiver are discussed, together with ways to reduce them. With an integration time of 125 ms, a signal-to-noise ratio of 1 is obtained for a THz beam with an average power of 10^{-16} W. The receiver operates in the sampling mode and has a time resolution of 0.5 ps.

I. INTRODUCTION

IN A SERIES of recent papers, the features and applications of a new high-brightness pulsed THz beam system have been reported [1]–[4]. The highest performance version of the system is based on repetitive, subpicosecond optical excitation of a Hertzian dipole antenna [3]–[5] embedded in a charged coplanar transmission line structure. The burst of radiation emitted by the resulting transient dipole is collimated by a THz optical system into a diffraction-limited beam and focused onto a similar receiver structure, where it induces a transient voltage and is detected. The THz optical system gives exceptionally tight coupling between the transmitter and receiver, while the excellent focusing properties preserve the subpicosecond time dependence of the source.

As mentioned previously [2]–[4], the combination of THz optics with the synchronously gated, optoelectronic detection process has exceptional sensitivity for repetitively pulsed beams of THz radiation. With lenses and mirrors it is possible to direct a large fraction of the radiation, emitted during the excitation of an optoelectronic device, onto a distant device. The transmitted waveforms have been measured with subpicosecond resolution and with signal-to-noise ratios of more than 10000:1. This result was obtained with a repetition rate

of 100 MHz, an integration time of 125 ms, and an average power of 10 nW in the THz beam.

In this paper we present measurements and analysis of the total emitted power, the beam collection efficiency, and the limiting noise properties of the THz receiver. In particular, we show how the combination of THz optics with the optoelectronic, subpicosecond, synchronous gating of the receiver allows for the measurement of incident THz beams with peak powers far less than the incident room-temperature thermal radiation in the same frequency range. Methods to further improve the performance of the receiver are also discussed.

The exceptional sensitivity is due in part to the high performance of the THz optics. Via two stages of collimation a THz beam with a frequency-independent divergence of 25 mrad is obtained from the THz transmitter. The THz receiver with identical optical properties collects essentially all of this beam. The resulting tightly coupled system of the THz transmitter and receiver gives strong reception of the transmitted pulses of THz radiation after many meters of propagation.

Another reason for the exceptional sensitivity is that the optoelectronic gating changes the effective resistance of the receiving antenna from 22 M Ω in the off state to 550 Ω in the on state. The gating window of the on state lasts for approximately 0.6 ps. The duty cycle of 0.6×10^{-4} is given by the ratio of the 0.6 ps gating time to the period of the 100 MHz sampling rate. Consequently, the average resistance seen by a current amplifier directly connected to the receiving antenna is about 7 M Ω . Therefore, with respect to the current amplifier, the receiver has a high impedance together with a subpicosecond response.

A final important feature of the detection method is that it is a coherent process; the electric field of a repetitive pulse of THz radiation is directly measured. Because we synchronously detect a repetitive signal, the total charge (current) from the signal increases linearly with the number of sampling pulses, while the charge (current) from noise increases only as the square root of the number of pulses.

II. EXPERIMENTAL SETUP

The setup used to generate and detect beams of short pulses of THz radiation is presented in Fig. 1. The transmitting and receiving antennas are identical, each consisting of the metallic pattern shown in Fig. 1(a) [3], which

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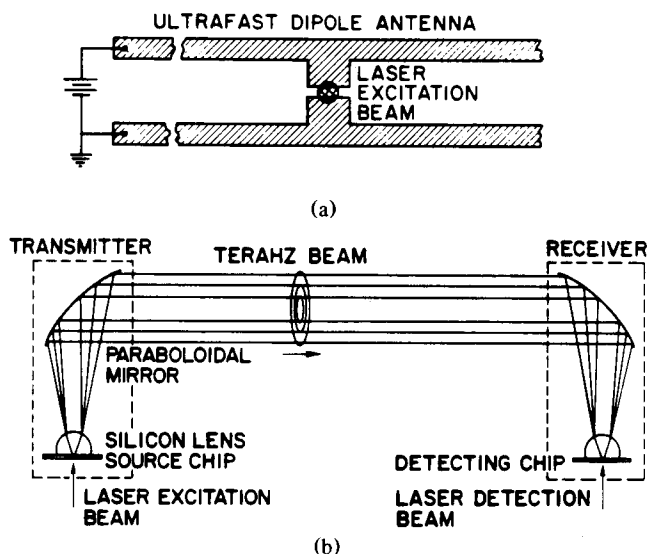


Fig. 1. (a) Ultrafast dipolar antenna. (b) Schematic diagrams of the THz transmitter and receiver.

has been fabricated on an ion-implanted silicon-on-sapphire (SOS) wafer. The $20\text{-}\mu\text{m}$ -wide antenna structure is located in the middle of a 20-mm -long coplanar transmission line consisting of two parallel $10\text{-}\mu\text{m}$ -wide, $1\text{-}\mu\text{m}$ -thick, $5\ \Omega/\text{mm}$ aluminum lines separated from each other by $30\ \mu\text{m}$. A colliding-pulse mode-locked (CPM) dye laser produces $623\ \text{nm}$, $70\ \text{fs}$ pulses which are focused onto the $5\text{-}\mu\text{m}$ -wide photoconductive silicon gap between the two antenna arms. This $100\ \text{MHz}$ periodic excitation causes pulsed, subpicosecond changes in the conductivity of the antenna gap. When a dc bias voltage of typically $10\ \text{V}$ is applied to the transmitting antenna, these changes in conductivity result in pulses of electrical current through the antenna, and consequently bursts of electromagnetic radiation are produced. A large fraction of this radiation is emitted into the sapphire substrate in a cone normal to the interface [2] and is then collected and collimated by a dielectric lens attached to the back side (sapphire side) of the SOS wafer [1]–[4].

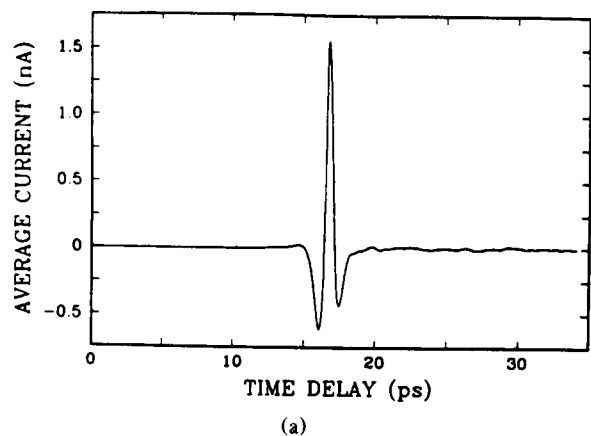
For the work reported here, the dielectric lenses were made of high-resistivity ($10\ \text{k}\Omega\cdot\text{cm}$) crystalline silicon with a measured absorption of less than $0.05\ \text{cm}^{-1}$ in our frequency range. The use of silicon gave significant improvement over the sapphire lenses previously used [1]–[3], although the 10% mismatch in dielectric constant between the silicon lens and the sapphire wafer causes a slight reflection. The center of the truncated 9.5-mm -diameter silicon sphere (lens) is $2.0\ \text{mm}$ above the ultrafast antenna located at the focus of the lens.

After collimation by the silicon lens, the beam propagates and diffracts to a paraboloidal mirror, where the THz radiation is recollimated into a highly directional beam. Although the $70\ \text{mm}$ aperture paraboloidal mirrors have a $12\ \text{cm}$ focal length, a $16\ \text{cm}$ distance was used between the silicon lenses and the mirrors to optimize the response of the system at the peak of the measured spectrum. While the high-frequency components of the

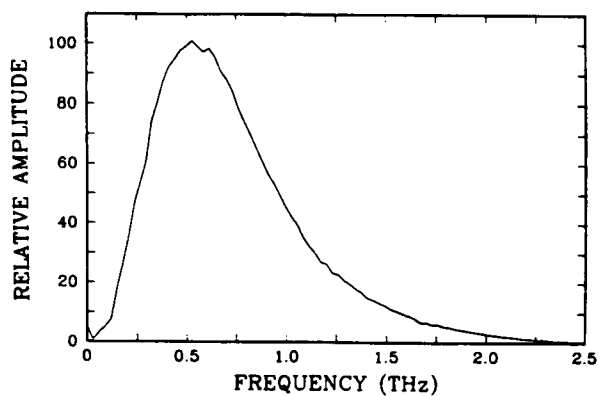
THz pulses remain reasonably well collimated, the very low frequency components quickly diffract and illuminate the entire paraboloidal mirror, which presents a solid angle of 0.15 steradians to collect radiation from the source. After recollimation by the paraboloidal mirror, beam diameters ($10\text{--}70\ \text{mm}$) proportional to the wavelength were obtained; thereafter, all of the frequencies propagated with the same $25\ \text{mrad}$ divergence. The combination of the paraboloidal mirror and silicon lens (THz optics) and the antenna chip constitute the transmitter, the source of a highly directional, freely propagating beam of (sub)picosecond THz pulses. After a $50\ \text{cm}$ propagation distance this THz beam is detected by an identical combination, the THz receiver, where the paraboloidal mirror focuses the beam onto a silicon lens, which focuses it onto a SOS antenna chip, similar to the one used in the emission process. The electric field of the focused THz radiation induces a transient bias voltage across the receiving antenna, directly connected to a low-noise current amplifier. The amplitude and time dependence of this transient voltage are obtained by measuring the collected charge (current) versus the time delay between the THz pulses and the CPM laser pulses that periodically gate the sensitivity of the receiver. The detection process with gated integration can be interpreted as a subpicosecond boxcar integrator.

III. MEASUREMENTS

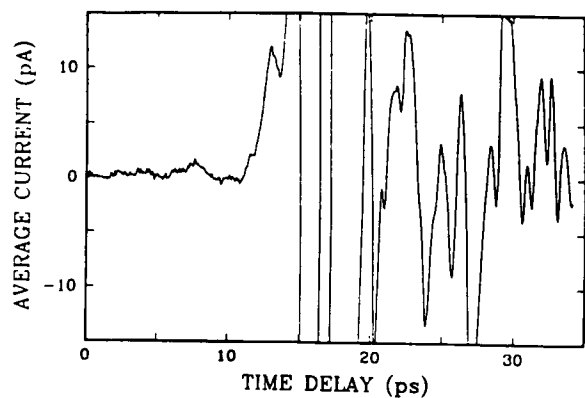
A typical time-resolved measurement is shown in Fig. 2(a). The clean pulse shape is a result of the fast action of the photoconductive switch at the antenna gap, the broad-band response of the ultrafast antennas, the broad-band THz optical transfer function of the lenses and paraboloidal mirrors, and the very low absorption and dispersion of the silicon lenses. The measured pulse width of only $0.5\ \text{ps}$ (FWHM) is only an upper limit to the true pulse width, because no deconvolution has been applied to the measurement to take out the response time of the antenna gap. The value of $14\ \text{mV}$ was established as a lower limit for the peak transient voltage across the gap; this was done by determining the required dc voltage for a photocurrent in the receiving antenna equal to the measured average current at optimum delay. The words *lower limit* indicate that the measured current results from a convolution of the transient voltage with the transient conductivity of the gate. In this regard, notice that the FWHM of the central spike is only $0.5\ \text{ps}$ wide, compared with the $0.6\ \text{ps}$ pulse width of the integrating gate. A deconvolution results in a peak transient voltage of $35\ \text{mV}$, which is about 2.5 times larger than the simple estimate mentioned above. The associated electrical pulse generated on the coplanar line by the excitation of the transmitting antenna is typically measured to be $1.0\ \text{ps}$ in duration and close to $1\ \text{V}$ in amplitude. Taking into account the response time of the measurement, the pulse on the line is considered to have a pulse width of about $0.7\ \text{ps}$ with a rise time of the order of $0.3\ \text{ps}$ and a longer



(a)



(b)



(c)

Fig. 2. (a) THz pulse measured by scanning the time delay between the optical gating pulses and the incident THz pulses, while monitoring the current induced in the THz receiver. (b) Amplitude spectrum of the measured pulse shape. (c) THz pulse on a 100 times expanded vertical scale.

fall time determined by the 0.6 ps carrier lifetime in implanted silicon [7].

In Fig. 2(b), the Fourier transform of the measured signal (Fig. 2(a)) is shown to stretch from about 0.1 to 2.0 THz. This represents only a lower limit to the true extent of the emitted radiation as it contains the frequency response of the receiver. At the low-frequency end, the efficiency of both emitter and receiver has been shown to be proportional to the length of the antenna, i.e., proportional to the separation between the two lines of the coplanar transmission line. For extremely low frequencies

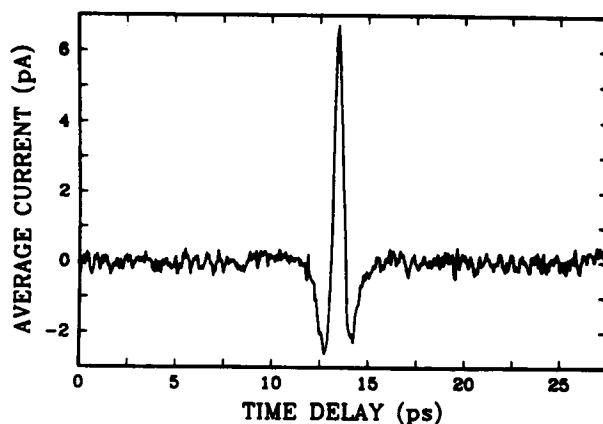


Fig. 3. Measured THz pulse with a 10^5 times reduction (compared with Fig. 2(a)) of the THz beam power.

the size of the paraboloidal mirrors will also limit the efficiency. For the high-frequency limit the efficiency of the antenna is strongly reduced when $1/2$ the wavelength (in the dielectric) of the emitted radiation is no longer large compared with the antenna length. This frequency for our $30 \mu\text{m}$ antenna is 1.5 THz, so that the observed signal and corresponding spectrum are somewhat limited by the antenna response, which has dropped by 50% at this frequency. The high-frequency part of the spectrum is also limited by the finite rise time of the current transient and the nonideal imaging properties of the THz optics.

In Fig. 2(c), the time-resolved signal is shown on a vertical scale that has been expanded by a factor of 100. The structure observable after the main pulse is reproducible and is due to reflections of the electrical pulse on the transmission line, reflections of the THz pulse from the various dielectric interfaces, and absorption and dispersion of water vapor [4] in the 1 cm path outside the vaportight box placed around most of the setup. The observed noise in front of the main pulse is about 1.3×10^{-13} A rms for an integration time of 125 ms, corresponding to an integration bandwidth of 1 Hz determined by a 12 dB/octave filter. An identical noise value is obtained when the far-infrared beam is completely blocked. The signal-to-noise ratio in this 4 minute scan is more than 10000:1. Another 4 minute scan is shown in Fig. 3, for which the intensity of the pump laser beam was reduced from the 6 mW normally used to only $15 \mu\text{W}$. This 400-fold reduction in laser power led to a reduction in the transient photocurrent of 320, instead of the expected 400. The discrepancy indicates a slight nonlinearity due to the onset of saturation, related to the fact that the electrical pulses generated on the transmission line are quite strong (almost 1 V in either direction). This 320-fold reduction in photocurrent led to a reduction in the power of the THz beam by the factor 1.0×10^{-5} . However, despite this enormous reduction in power, the peak amplitude is still more than 30 times larger than the rms noise. Based on calculations presented in the next section we believe that the average power in the THz beam during this measurement was about 10^{-13} W. If the

power of the THz beam were even further reduced, the detection limit of the THz receiver would be reached at 1×10^{-16} W, for a signal-to-noise ratio of unity and a 125 ms integration time.

IV. CALCULATED POWER IN TERAHERTZ BEAM

Before we start a detailed analysis of the noise characteristics of the THz receiver, the emission and detection of the THz beam will be discussed in more detail. Below we will give an estimate of the power emitted by the transmitter, starting with the well-known equation for the power emitted by a harmonically oscillating dipole in free space. Several corrections have to be applied to this simple formula.

First of all, the current, generated in the transmitting antenna through the creation of photocarriers, is nonharmonic. To calculate the emitted power we decompose the transient current into its Fourier components, use the standard formula for each of these components, and finally perform the frequency integration. Most of the radiation is emitted in the rising edge of the transient current and not in the 0.6 ps exponential decay time of the conductivity, which is equal to the carrier lifetime for the radiation-damaged silicon [7]. Thus, the rise time is of vital importance and it proved to be slower than expected from the 70 fs laser pulse width. Using an exponential rise time as adjustable parameter, a good fit to the measured far-infrared spectrum was obtained with a rise time of slightly less than 200 fs. The reasons for the slow rise time are not presently understood. Tentative explanations involve carrier cooling effects, the frequency-dependent conductivity, radiation damping, and the circuit response.

As a second feature, it is important to note that the antenna is not an infinitely short dipole. The effective length of the antenna is probably more than the $30 \mu\text{m}$ separation between the lines and depends on the current flow from the antenna into the transmission line. In the calculations an effective length of $35 \mu\text{m}$ is used. This length is roughly equal to a half wavelength of 1.5 THz radiation in the dielectric, resulting in a decrease in emission efficiency of about 50% and limiting the high-frequency end of the integration.

As a last correction we mention that the antenna is situated at an interface between air and a dielectric medium with an index of refraction in the far infrared of n . It has been shown [8], that this situation results in an increase in emitted power by roughly a factor n when n is much bigger than 1. Furthermore, the emitted radiation is strongly directed into the dielectric and concentrated along the normal to the interface. The exact angular distribution of the radiation can be very complicated [2], [8], [9], making it difficult to estimate the fraction of emitted radiation that is collimated into the THz beam. A first consideration is the reflection loss that occurs when the radiation is coupled through the silicon lens into the air. This reflection loss is about 30% at normal incidence. However, the angle for total internal reflection for silicon

is only 17° . Calculations show that with a spherical lens it is therefore impossible to capture more than 40% of the THz radiation. To estimate the actual fraction that ends up in the THz beam, a 5-mm-diameter diaphragm was placed in front of the silicon lens. As a result the power in the THz beam was reduced by roughly a factor 2. The diaphragm restricts the captured radiation to an emission cone of 19° (half angle). Calculations show the fraction of radiation within this cone to be only 7%, reflection losses included. Consequently, we estimate that without the diaphragm only 15% of the total emitted radiation ends up in the THz beam, the rest being lost by (total internal) reflection and lens aberration. The complicated field profile at the dielectric lens diffracts out upon propagation to the paraboloidal mirror, and the THz beam profile between the paraboloidal mirrors roughly resembles a (Gaussian) TEM_{00} mode for each of the frequency components, with the diameter increasing as a function of wavelength.

At a dc bias of 10 V and a laser power of 6 mW we find an average photocurrent of 1.1×10^{-6} A, corresponding to a peak current of about 1.8×10^{-2} A. Using this value and the method just described we calculate the total frequency-integrated power emitted by the transmitting antenna to be 75 nW. Using our previously derived value of 15% to describe the coupling efficiency, we obtain 11 nW in the collimated, freely propagating THz beam emitted from the transmitter.

An alternative estimate of the transmitted power can be obtained at the receiving antenna, by combining the measured amplitude of the incident far-infrared field with the measured size of the focal spot. The corrections that have to be applied are similar to those mentioned above for the transmitter.

First of all the response time of the receiving antenna is determined by its gating time and the rise time of the transient conductivity. The Lorentz reciprocity theorem shows that the frequency dependence of the receiving antenna should equal that of the transmitting antenna [10]. Second, the boundary condition results in a complicated angular sensitivity of the receiving antenna and an increase in electric field by the factor $2n/(n+1)$ for radiation incident perpendicular to the interface. As a third and a fourth correction we have to mention the loss of high-frequency components due to the size of the antenna and the reflection losses at the front surface of the dielectric lens.

Measurements show that the size of the focal spot on the receiving antenna is wavelength dependent, being larger for longer wavelengths. The measured (FWHM) diameters of the focal spots were about $180 \mu\text{m}$ at 1.0 THz, $280 \mu\text{m}$ at 0.5 THz, and $460 \mu\text{m}$ at 0.25 THz. The (deconvoluted) induced peak voltage of about 35 mV over the $35 \mu\text{m}$ antenna structure corresponds to a focused field of about 6 V/cm in the dielectric, if we correct for surface enhancement. These figures yield an average power of about 7 nW in the air incident on the dielectric focusing lens of the receiver. This value is in reasonable

agreement with the previously derived value of 11 nW from the point of view of the transmitter.

An alternative analysis makes use of the concept of antenna impedance, defined as the ratio of the emitted power over the mean square of the current. In this respect it is important to notice that the current is constant along the antenna structure, which makes it a more efficient emitter than the related center-fed antenna [11]. The interface increases the antenna impedance and we find a value of 12Ω at 0.6 THz for our situation. From the transmitter point of view, a peak current of about 1.8×10^{-2} A is pulsed through the antenna. If we assume that the current and emitted power were concentrated in a narrow band around 0.6 THz, then the emitted peak power would simply be I^2R , and the average emitted power would be about 17 nW, a number that includes the aforementioned 15% coupling efficiency into the THz beam and the 0.6×10^{-4} duty cycle. From the 35 mV induced voltage at the receiving antenna, one can derive an average incident power of 4.5 nW in a similar way. It was only a rough approximation to assume that all radiation was concentrated around the 0.6 THz peak of the spectrum. When we take the proper frequency integrations into account together with the corrections mentioned above, we arrive at 11 nW for the emitted power and 3.5 nW for the received power at the receiving antenna. Taking into account the 30% reflection at the silicon lens, the value obtained from calculations at the receiving antenna gives 5 nW incident on the silicon lens.

Another useful parameter is the ratio of the voltage induced in the receiving antenna to the current in the transmitting antenna. This frequency-dependent ratio is called the mutual impedance [11], [12]. For perfect coupling the mutual impedance should equal the impedance of the (identical) antennas [12]. From our measurements we extract a mutual impedance of about 3Ω at 0.6 THz, corresponding to 25% amplitude coupling between the transmitter and the receiver for this frequency.

By reasoning from both the transmitter and the receiver point of view, we calculated the average power in the far-infrared beam to be about 11 nW and 5 nW or 7 nW, respectively. For the purposes of the discussion to follow, we consider the average transmitted power in the THz beam to be 10 nW, and we believe that this value is within a factor 2 of the actual value. This average power is comparable to that of conventional thermal sources [13], but the peak power of our source is orders of magnitude larger.

V. NOISE IN THE RECEIVER

In this section we will present an analysis of the noise of the THz receiver. The following noise analysis is generally applicable to any type of photoconductive switch, although the actual numbers will depend on the type of device. If we were to destroy the directionality of the receiver by removing the paraboloidal mirror and the add-on silicon lens, the noise characteristics of the antenna/switch would in fact be unchanged. However, the

sensitivity of the receiver for the THz beam would be drastically reduced.

Experimentally, we measured the receiver noise with a frequency analyzer (Hewlett Packard 3585B) connected to the output of a low-noise current amplifier (Ithaco 1211). The amplifier was in turn connected to the receiving antenna via the terminating coplanar transmission line. This arrangement allowed the investigated noise to be separated from the small pickup of harmonics of the line current. An alternative approach was to measure the fluctuations on the output of the lock-in amplifier that was also connected to the current amplifier (CA). These fluctuations were then converted to an rms current at the input terminals of the current amplifier and could thus be compared with the noise floor (in dBm) measured with the spectrum analyzer.

In the absence of the CPM laser beam, the noise of the first receiving antenna chip investigated was 1.5×10^{-13} A rms for a 125 ms integration time on the lock-in (-73 dBm, with the CA at a setting of 10^{-8} V/A and a bandwidth of 10 Hz). That this value was due to the thermal Johnson noise, also called Nyquist noise, could be demonstrated by replacing the antenna chip by an ordinary resistor of identical value ($0.7 \text{ M}\Omega$). The easiest way to reduce this noise is to strip the silicon from the areas of the chip that are not illuminated by the laser during normal operation. Such an antenna chip with stripped silicon showed a much increased resistance of $22 \text{ M}\Omega$ in the off state. For this antenna the noise measurement gave about 3.9×10^{-14} A rms (125 ms), being only slightly affected by the much lower, open-ended current noise of the CA of 1.5×10^{-14} A rms.

When a 6 mW laser beam was focused on the antenna gap, the noise increased to 2.2×10^{-13} A rms (125 ms) for the first antenna chip and to 1.3×10^{-13} A rms for the second stripped antenna chip. For the second chip we could show that the amplitude of the additional noise was proportional to the square root of the laser power. A likely candidate for this additional noise is the Johnson noise arising from the thermally driven random motion of electrons. Each laser pulse temporarily reduces the resistance between the two lines of the terminating coplanar transmission line. During the 0.6 ps switching time the resistance is typically about 550Ω . The integrated thermal noise power scales with the average conductivity, which in our situation is reduced from $22 \text{ M}\Omega$ to about $7 \text{ M}\Omega$. Thus, we would expect this noise to be 4.8×10^{-14} A rms (125 ms, 1 Hz). However, at 6 mW the measured noise power was equal to the Johnson noise of a $1 \text{ M}\Omega$ resistor and not that of the measured average resistance of $7 \text{ M}\Omega$. Consequently there must be additional noise sources, some of which are described below. The measured antenna noise can also be interpreted as a random effective voltage and charge transfer at each shot of the laser. The 1.3×10^{-13} A rms (125 ms) noise level corresponds to a shot-to-shot effective voltage of 4.2 mV. This value seems to be very large, but results only in a charge transfer of 4.6×10^{-18} C or roughly 30 electrons per shot,

while the incident laser beam creates about 10^8 electrons per shot.

In the experiment it was noticed that, even though no bias voltage was applied to the receiving antenna, the focused laser beam still gave rise to a constant photocurrent of about 3 nA, as if the antenna were biased with about 30 mV. We think this might be due to a nonohmic contact between the silicon and the metallic antenna and terminating transmission line. Although this stray photocurrent does not show up in the signal it can lead to additional noise. First of all the laser power is not perfectly stable, but experiences fluctuations of about $\Delta I/I = 1.0 \times 10^{-5}/\sqrt{\text{Hz}}$ at the 1 kHz modulation frequency. The noise arising from these fluctuations will definitely contribute but cannot be the dominant noise source, because if it were the noise amplitude should be directly proportional to the laser power. Second, the 3 nA stray current will lead to quantum noise of about 3×10^{-14} A ($RC = 125$ ms). We estimate both noise sources arising from this stray current to be of the same order of magnitude as the expected Johnson noise.

VI. THERMAL BACKGROUND RADIATION

The classic broad-band detector for the (far) infrared is the bolometer. This device is strongly affected by thermal noise, which sets serious limitations on its sensitivity [6], [14]. It has been shown [14] that the noise equivalent power (NEP) of temperature sensors in general can never be less than 8×10^{-12} W (125 ms) at room temperature. Therefore bolometers are usually operated at 4.2 K or lower. At 4.2 K their NEP¹ is typically 1×10^{-13} W/ $\sqrt{\text{Hz}}$ [6].

The gated and coherent nature of our THz receiver dramatically reduces its sensitivity to thermal background radiation. We will now estimate the noise contribution on our antenna arising from this thermal background. The combined action of thermal far-infrared radiation, with an energy of about kT per mode, from low frequency to the high-frequency response limit of the receiver induces a random voltage across the receiving antenna which is converted into a current and detected. This random voltage is $\sqrt{(4kTR)}$ in $V/\sqrt{\text{Hz}}$, where R is the characteristic impedance of the antenna [15], [16]. We estimated the charge transfer due to the thermal background at room temperature to be about 2.5×10^{-19} coulomb (C) at each shot of the laser. This corresponds to an effective pickup voltage of about 0.23 mV, during the 0.6 ps gate. Because the thermal radiation is incoherent and the pickup voltage for successive shots is unrelated, the net charge transfer across the antenna gap can be viewed as a random walk. The noise in the average current created by the thermal background should thus be about 7×10^{-15} A (125 ms, 100 MHz repetition rate), or roughly a factor 7

¹As explained in [15], a coherent detector measures amplitude; therefore, the NEP is proportional to the measurement bandwidth. In contrast to this, an incoherent detector measures power; therefore, the NEP is proportional to the square root of the measurement bandwidth.

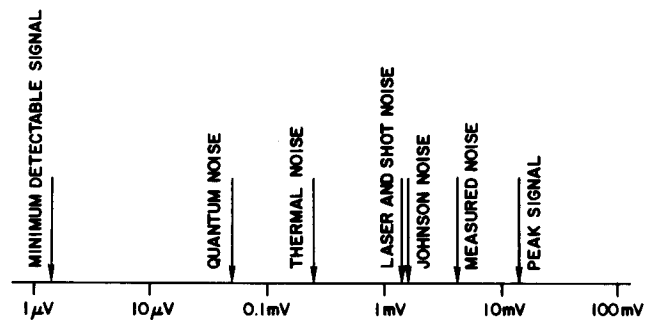


Fig. 4. Logarithmic presentation of the measured peak signal, the minimum detectable signal, and the discussed noise sources, in terms of the induced voltage across the receiving antenna.

lower than the expected Johnson noise. Via a similar calculation we obtain the random charge transfer due to the quantum fluctuations, having a power density of $h\nu$ per mode, to be roughly 5×10^{-20} C per shot of the laser, corresponding to an effective voltage of 0.05 mV and a current noise of 1.4×10^{-15} A (125 ms). Both values depend strongly on the frequency response of the gate and thus on the rise time of the transient conductivity. This dependence occurs because the noise power at high frequencies is much larger than that at low frequencies, due to the quadratic increase of the density of electromagnetic modes with frequency. Consequently the above-mentioned noise values are only rough estimates, but they do provide important points of reference for the sensitivity of the THz receiver. The noise arising from the thermal background and the quantum fluctuation would remain unchanged if the paraboloidal mirror and add-on lens were removed from the receiver, as this only changes its directionality and the receiving antenna "sees" the same temperature in all directions.

VII. DISCUSSION AND SUMMARY

At this point we want to review and compare some relevant figures. The THz receiver detects, with signal-to-noise ratios of approximately 10000:1, subpicosecond, 14 mV pulses coming at a 100 MHz repetition rate in a highly directional beam of THz radiation with an average power of 10 nW. Consequently, the detection limit for these repetitive pulses is about 1.4 μ V. However, it has just been shown that the sampled voltage, during a single 0.6 ps gating pulse, on the receiving antenna due to the thermal background is about 0.23 mV and due to the vacuum fluctuations is 0.05 mV. Thus, in terms of instantaneous voltages, the receiver can detect 1/160 of the thermal background and 1/35 of the vacuum fluctuations. Beams of THz radiation can be detected with peak powers of only 4×10^{-5} that of the incident thermal radiation. This impressive performance is due to the high directionality of the THz receiver and to the fact that the thermal noise is incoherent and adds randomly for successive gating pulses, while the signal propagating in the THz beam is coherent and scales linearly with the number of gating pulses. Fig. 4 graphically presents the various noise

sources and the signal amplitude in terms of the effective induced voltage at each gating pulse.

The generation and detection of the terahertz (far-infrared) radiation is coherent; therefore the THz receiver is intrinsically much more sensitive than the incoherent bolometer. With respect to this comparison it should be clear that the bolometer can measure the average power in the repetitively pulsed THz beam. However, our synchronously gated and coherent detection method cannot directly measure incoherent THz beams. The only detectable effect of an incoherent beam on our detector would be an increase in the noise on the receiving antenna. Another consideration is whether the gated detector can detect a single frequency wave. In the absence of synchronization between monochromatic radiation and a high harmonic of the 100 MHz gating rate, the gated detection will yield a beat signal which averages to zero. Even if the CW wave is synchronized, a large reduction in sensitivity will occur, because of the limited duty factor of 0.6×10^{-4} . With respect to the above comment, it is still appropriate to compare the detector with a radio receiver, which in fact it is, although the antenna is only 35 μm long and the device is operated in this nonconventional gated mode. For a good radio the NEP can be as low as 10^{-19} W/Hz (see [17]), 1000 times lower than for our receiver. To compare our detector with a radio it is instructive to interpret the gated operation of the receiver in the frequency domain. Here, the 10 ns periodic gating of the receiver results in the simultaneous detection of thousands of narrow frequency bands separated by the 100 MHz rate and with a THz roll-off determined by the 0.6 ps gating time. This results in broad-band performance together with the sensitivity increase due to coherent detection. This picture also shows why the influence of thermal (incoherent) radiation is so small. The receiver is only sensitive to radiation close (within 1 Hz for a 125 ms integration time) to the 100 MHz multiples of the comb.

The performance of a radio deteriorates when the impedance of the lead from the antenna to the radio is not matched with that of the antenna itself or when the lead is lossy. The laser gating pulses reduce the resistance between the lines to 550 Ω , which definitely is not matched with the antenna impedance, which we calculated to be about 12 Ω at 0.6 THz. A better match is obtained for longer antennas. Changing both transmitting and receiving antenna to a structure with lines separated by 50 μm did indeed increase the peak signal by more than a factor of 2, while the noise was not affected. However, increasing the size of the antenna leads to a reduction in the speed of the device. Furthermore, the way the THz pulses are detected, by making the area between the antenna stubs conductive, is based on real resistance and thus by definition is lossy. Only the creation of a complete short could result in loss-free coupling to the transmission line, a coupling that would be optimum if the antenna impedance were matched to that of the transmission line. At room temperature the ultimate

performance of the THz receiver would then be determined by the fluctuations in thermal background.

In this paper we have described the performance of an optoelectronic system for the generation and detection of beams of 1/2 cycle pulses of THz (far-infrared) radiation. The THz transmitter operation is based on repetitive, subpicosecond laser excitation of a Hertzian dipole antenna embedded in a charged coplanar line. The associated optoelectronic receiver is gated in synchronism with the excitation of the transmitter by subpicosecond pulses from the same laser source. With this transmitter 10 nW (average power) highly directional electromagnetic beams, consisting of 0.5 ps pulses of THz radiation with a repetition rate of 100 MHz, are generated. After freely propagating to the receiver, these beams can be detected with a signal-to-noise ratio of 10000:1 and with a sampling time resolution of 0.5 ps. We have discussed several different sources of noise in the present receiver design, which for an integration time of 125 ms can detect repetitive subpicosecond pulses in a beam with an average power as low as 10^{-16} W. We have shown how the thermal noise will eventually limit the performance of an ultimate receiver. Our receiver is more noisy in amplitude than this ultimate device by a factor of approximately 20. The reasons for that are reasonably well understood, and efforts are being made to improve the system performance.

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