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Chapter 5

Principles of THz Direct Detection¹

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This chapter is a comprehensive review of the physical principles and engineering techniques associated with room-temperature THz direct detectors. It starts with the basic detection mechanisms, both rectifying and thermal. It then addresses the noise mechanisms using both classical and quantum principles, and the THz coupling using impedance-matching and antenna-feed considerations. Fundamental analyses are provided of the noise mechanisms because of lacking coverage in the popular literature. All THz detectors can then be described with a common performance formalism based on two metrics: noise-equivalent power (NEP) and noise-equivalent temperature difference (NETD). A summary is then provided of the typical, and best, detector performance reported to date, including the various CMOS-based detectors. The chapter concludes with a comparison of the best THz detector results at room temperature, and demonstrates that none of the existing detector types are operating anywhere near fundamental theoretical limits, so there is room for significant performance advances.

5.1. Detection Mechanisms

The THz region is often considered as the transition between the radio frequency (RF) bands below 100 GHz and the popular infrared (IR) bands at wavelengths beyond ~ 12 μm . As such, it has a tendency to exhibit passive components and devices having both RF and IR characteristics. A good example is so-called “quasi-optical” components for which free-space radiation is coupled to passive components using mirrors, lenses, and other traditional optical components. Another good example is THz “direct” detectors which by themselves convert THz incoming radiation *directly* to a low-frequency baseband. In the THz region the direct detector is almost always a power-to-voltage or power-to-current converting device. That is, it is a device that puts out a voltage or current in proportion to the incoming power. There are many examples of such devices, but the most popular for room-temperature operation are field detectors and thermal detectors.

Field detectors, such as Schottky diodes, respond to the THz electric field and usually generate an output current or voltage through a quadratic term in their current-voltage characteristic. Thermal detectors are often composite devices consisting of a THz absorbing element coupled to some temperature-sensitive transducer. The THz absorber is generally isolated thermally from the environment so that the absorbed THz power raises the temperature both of the absorbing layer and the attached transducer. If the transducer is a thermistor, the detector is called a bolometer. If the

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transducer is a temperature-dependent capacitor based on pyroelectricity, it is called a pyroelectric detector. These are the two types that we will focus on in this review article.

The thermistor in a bolometer is generally a device that displays a large change of resistance to a small change of temperature. Such behavior is displayed by certain transition metals and semi-metals. Furthermore, such materials can also be fabricated as THz resistors which effectively absorb THz radiation via the AC form of Ohm's law. Therefore, it is possible to construct bolometers such that the absorbing function and the thermistor function are carried out by the same element. Similarly, if the temperature-dependent material in a pyroelectric capacitor also absorbs THz radiation sufficiently, then the capacitor serves both functions as well. This type of bolometer and pyroelectric are referred to, henceforth, as "integrated" direct detectors.

A more recent THz detection mechanism is based on plasma-wave propagation and rectification in the 2D channel of high-electron mobility field-effect transistors (HEMTs). The basic idea is that the 2DEG of HEMTs or cryogenically cooled MOSFETs can support plasma wave propagation even at frequencies well above the unity-current cutoff frequency (f_τ), which is generally limited by the electron transit time between source and drain. For the typical sheet concentrations in these devices ($\sim 10^{12}/\text{cm}^2$), the plasma-wave group velocity is roughly one order of magnitude higher than that of the free electrons. So a change of gate-source voltage V_{GS} affects the drain-source voltage roughly 10-times faster than in conventional FETs. And as described below the drain-source voltage is a nonlinear function of V_{GS} , so an ac gate-source voltage yields a DC term in the drain-source potential, which is the basis for THz detection.

5.1.1. E-Field Rectification

Simply stated, a square-law field rectifier detector is a device or circuit that takes an input signal and produces an output that is proportional to its square,

$$X_{OUT} = AX_{IN}^2, \quad (5.1)$$

where X_{OUT} could be a current or voltage and A is a proportionality constant. The utility of such a device in detection is two-fold: (1) rectification, or producing a DC output from a purely AC input, or (2) frequency down-conversion (i.e., mixing). Rectification is most easily understood in the special case of a coherent signal

$$X_{IN}(t) = B\cos(\omega t + \phi) \quad (5.2)$$

If put through a square-law detector, the output becomes

$$X_{OUT}(t) = \frac{1}{2} AB^2 \{1 + \cos[2(\omega t + \phi)]\}. \quad (5.3)$$

So if the square-law detector is followed up by a low pass filter (i.e., a time-domain integrator) with integration time $\tau \gg 2\pi/\omega$, then the second term will not contribute and the output of the filter will be

$$X_{OUT} = \frac{1}{2} AB^2 \equiv \mathcal{R}P_s \quad (5.4)$$

where $\mathcal{R}$ is a constant and P_s is the average absorbed input signal power. In the language of sensor theory, $\mathcal{R}$ is usually called the *responsivity* and is measured either in A/W or V/W. Presumably, A is

known, so that B can be determined from the X_{OUT} DC term. Square-law detectors are preferred in RF systems over cubic and other possible detectors partially for this reason. The proportionality constant A is dependent only on the detector characteristics and not on the power level, at least up to a point where saturation and higher-order effects begin to occur. A single calibration of A in the “small-signal” regime is all that is required to use the square-law detector over a wide range of input power.

The paradigm THz rectifier is the metal-semiconductor barrier (Schottky) diode. Originally this was fabricated as a sharpened metal wire, or “whisker”, contacting a bare semiconductor such as galena (PbS). Later on with the advent of microelectronic fabrication, the whisker was replaced by a lithographically defined metal contact deposited on the virgin semiconductor surface and the resulting current-voltage characteristic was found to be well described by the highly asymmetric (with respect to bias voltage V), generic expression

$$I = I_S [\exp(eV/\alpha k_B T) - 1] \quad (5.5)$$

where I_S is the reverse-bias saturation current, k_B is the Boltzmann constant, T is the contact temperature, and α (typically $1.0 < \alpha < 2.0$) is a physically-rich parameter called the ideality factor. With the importance of rectifying detectors to radar receivers during World War II, detailed studies were made of the small-signal responsivity of such devices and the following general expression was found for the short-circuit current responsivity [1]

$$\mathcal{R}_I = \frac{1}{2} \frac{d^2 I / dV^2}{dI / dV} \quad (5.6)$$

When applied to the ideal metal-semiconductor diode Eq. (5.5), (5.6) becomes

$$\mathcal{R}_I = e / (2 \alpha k_B T) \quad (5.7)$$

At 300 K and assuming $\alpha = 1$, this calculates to $\mathcal{R}_I = 19.3$ A/W, which represents a maximal short-circuit responsivity. Notice that this result and Eq. (5.7) in general is independent of bias voltage and the area of diode too to the extent that α is independent of these operating parameters. If the rectifier is terminated in a load impedance (e.g., high input-impedance amplifier) much greater than dV/dI , then the diode is self-loading and Eq. (5.7) can be written as an open circuit voltage responsivity. For the ideal metal-semiconductor diode, this becomes

$$\mathcal{R}_V \approx \mathcal{R}_I \cdot dV/dI = \frac{1}{2} \frac{d^2 I / dV^2}{(dI/dV)^2} = e R_j / (2 \alpha k_B T) \quad (5.8)$$

where $R_j \equiv dV/dI$ is the differential resistance. Since Schottky rectifiers are usually biased with constant voltage, Eq. (5.8) is re-written as:

$$\mathcal{R}_V = [\exp(-eV/\alpha k_B T)] / (2 I_S) \quad (5.9)$$

This clearly shows that the voltage responsivity does depend on bias voltage and device area (through I_S).

Fig. 5.1 shows a family of I-V curves (a) and the corresponding $\Re_V$ -V curves (b) for a relatively recent type of Schottky diode consisting of a single-crystal ErAs semimetallic contact on a quaternary semiconductor (InGaAlAs) having a tunable bandgap via the Al fraction [2]. This exhibits a number of interesting effects. For zero percent Al, we have a metal contact on $\text{In}_{0.53}\text{Ga}_{0.47}\text{As}$, a semiconductor having a band gap of approximately 0.75 eV at room temperature. The I-V curve looks very linear (i.e., “ohmic”). As the Al fraction is increased, we begin to see more unilateral current flow and an increase in the peak electrical responsivity in Fig. 5.1(b). For an Al fraction of 0.6, the peak responsivity approaches the maximum theoretical value of 19.3 A/W at approximately 0.05 V bias. For a further increase in Al fraction, the maximum remains about the same, but the voltage at which the maximum occurs increases to approximately 0.1 V. Both maximum voltages are way lower than what one would measure in the more common metal-GaAs or metal-Si Schottky diodes largely because the band-gap is much less. In fact, the material combinations in Fig. 5.1 were developed most for fabricating THz Schottky rectifiers operating with zero bias. This is easier to implement in practice than biased rectifiers, and because there is no bias current they display no $1/f$ noise of any type.

It would appear that Schottky rectifiers are relatively simple, and this is why historically they have usually been the first detector type to be applied in practically all emerging frequency bands starting with microwaves during World War II. As interest grew in THz frequency-operation during the 1960s and beyond, it was realized that reliable, planar-fabricated Schottkys could be realized simply by reducing their area to sub-micron dimensions to reduce the junction capacitance commensurately.

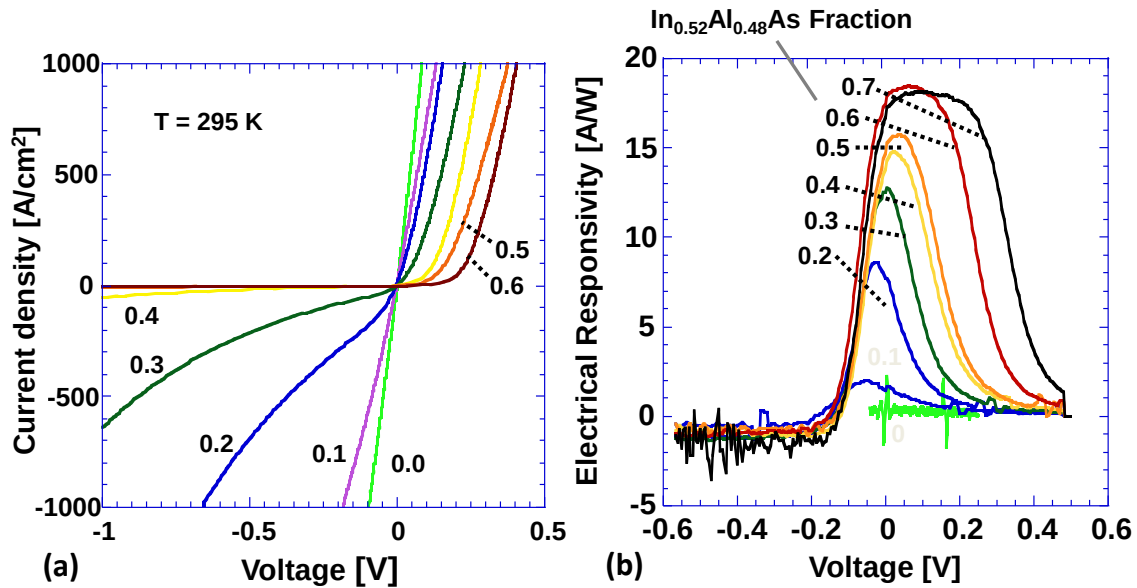


Fig. 5.1: (a) Room-temperature I-V curves and (b) corresponding $\Re_V$ -V curves for a family of Schottky diodes fabricated from ErAs (semimetal)-InGaAlAs (semiconductor) junctions and with various In, Ga and Al fractions.

However by so doing, the resistive part of the impedance also increases such that the $R_D C$ time constant remains roughly the same, at least near zero bias. Hence, the Schottky THz impedance is still reactive, making it difficult to conjugately match to any realistic THz coupling element, such as a waveguide or planar antenna. This topic will be addressed later in the chapter.

5.1.2. Thermal Detection

Two of the three detection technologies explored in this chapter belong to the class of detectors known as thermal detectors. The basic idea behind all such detectors is that the incident radiation causes a temperature change in some detection element, similar to a thermometer, which can then be measured electronically. While this sounds like a relatively simple concept, its realization can range from relatively simple in the case of the bolometric detectors discussed below, to intricate like the Golay cell.

5.1.2.1. Bolometers

A bolometer is a type of thermal detector that acts like a thermistor, i.e., a temperature-dependent resistor, for which the temperature change is created by the incident THz radiation. A schematic diagram of bolometer functionality is shown in Fig. 5.2. Bolometers typically include two functions: (1) a radiation coupling component, such as an integrating cavity or planar antenna on a substrate; and (2) a thermistor element to transduce the change of temperature to a change of electrical resistance, which is then measured in a read-out circuit via an applied bias. For room temperature operation, the thermometer is often made of bismuth or some semimetal having relatively low electrical conductivity compared to the transition or noble metals, but a similar thermal coefficient of resistance, α , defined by

$$R(T) = R_0(1 + \alpha \Delta T) \quad (5.10)$$

R_0 is the resistance at some reference temperature T_0 (usually 300 K), and $\Delta T = T - T_0$ with ΔT generally $\ll T_0$. If a two-terminal resistor is made of such material and then biased with a constant-current of I_B , a temperature change of the resistor will lead to a change of terminal voltage given by

$$\Delta V = V(R) - V(R_0) = I_B(R - R_0) = I_B R_0 \alpha \Delta T \quad (5.11)$$

Since ΔT is assumed much less than T_0 , then linear thermal transport theory can be applied by which $\Delta T \approx \Delta P / G_0$, where ΔP is the change in average power dissipated in R that creates the temperature rise, and G_0 is the thermal conductance at T_0 that links the resistor to the environment (e.g., heat sink). Hence we can write

$$\Delta V \approx I_B R_0 \alpha \Delta P / G \quad (5.12)$$

And the electrical responsivity is given by

$$\mathcal{R}_V \equiv \Delta V / \Delta P \approx I_B R_0 \alpha / G_0 \quad (5.13)$$

This is a much more complicated responsivity than for Schottky rectifiers in that all four parameters depend on the fundamental bolometer material characteristics, and three of them (I_B , R_0 , and G_0) also depend on the size, and operational conditions. This gives bolometers a very broad design space, but also introduces trade-offs that limit the room-temperature sensitivity to values well below those of high-quality rectifiers once the requirements of efficient coupling of radiation and limited current bias are factored in.

According to Eq. (5.13) the current bias wants to be as high as possible to maximize the responsivity. But in practice it is limited by heating of the bolometer element to the point of thermal runaway. To

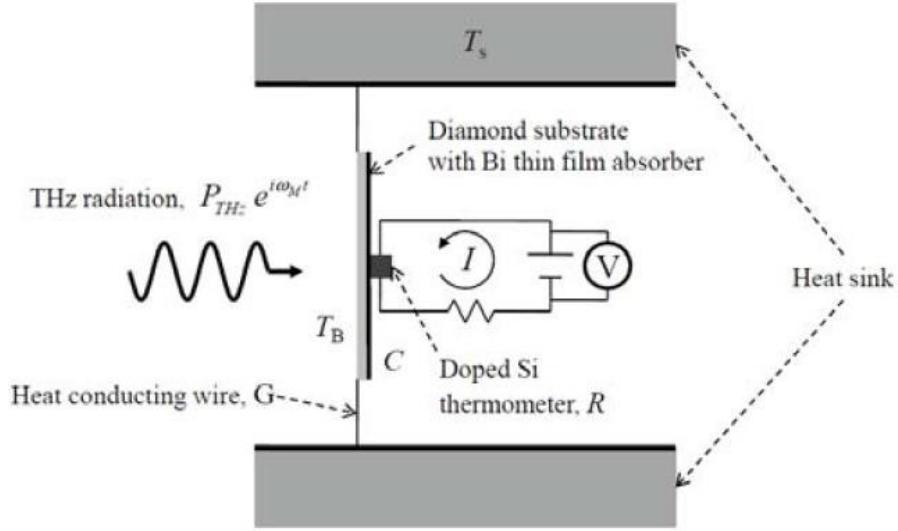


Fig. 5.2: Schematic of a bolometer detector (Ref. [3]).

see this we can re-write Eq. (5.10) with R defined in terms of the bias conditions ($R = V/I$) and ΔT determined entirely by the electrical bias power, $P = V \cdot I$, such that

$$R(T) = V/I = R_0(1 + \alpha V \cdot I / G_0) \quad (5.14)$$

Solving Eq. (5.14) for V , we get

$$V = R_0 I / [1 - R_0 \alpha I^2 / G_0] \quad (5.15)$$

Notice that the denominator can display a zero, meaning that if α is positive the voltage diverges at a current given by

$$I_{max} = [G_0 / R_0 \alpha]^{1/2} \quad (5.16)$$

Such a singular expression is reminiscent of regenerative feedback in circuit theory. However, if α is negative, the voltage is stable, reminiscent of degenerative feedback. In either case, the phenomenon leading to Eq. (5.15) is called *electrothermal feedback*, and is evident in all types of bolometers. It is important to realize, however, that even in the case of degenerative electrothermal feedback, there is still a maximum bias current dictated by the rise in temperature of the bolometer with respect to the ambient. Just as in an electrical fuse, Joule heating will eventually overwhelm the thermal dissipation and cause the bolometer element to be damaged or even melt.

Thermal transport also dictates another important aspect of bolometer behavior – the temporal response. Assuming the bolometer has a heat capacity C that is much greater than the heat capacity of the thermal link to the environment, then a step increase in temperature of the bolometer will create a temperature change having a characteristic (natural exponential) time constant of

$$\tau \approx C/G. \quad (5.17)$$

For macroscopically large bolometers (> 1 mm in all dimensions), this time constant tends to be prohibitively large at room temperature, often approaching 1 s. But fortunately, C scales with volume

while G tends to scale with peripheral area, so an isotropic reduction in volume tends to reduce the time constant accordingly to the millisecond level or even less.

Bolometer size is also important in the THz region because of radiative coupling. In the IR region the radiation coupling is relative simple because the thermistor function and radiation absorption can occur in the same detector element. But in the THz region, this is practically impossible because of the much longer wavelengths. A much better approach is to separate the thermistor and radiation coupling functions by coupling the bolometer element to some type of waveguide or antenna, be it planar, horn, or otherwise. To obtain a good THz impedance match, this then constrains R_0 in Eq. (5.11) to typical antenna and waveguide source impedances of order $100\ \Omega$. This constraint can be mitigated by coupling the bolometer to an integrating cavity, for example, but such an approach is not conducive to proliferating the bolometers into 1D or 2D arrays – a very useful exercise if the detectors are to be used for spatial imaging.

A fundamental issue with bolometers is that the room-temperature values of α are rather small, tend to have the same absolute value for all metallic solids including semimetals, but can differ in sign. The common absolute value turns out to be $\alpha \sim 0.003$. This follows from the fact that all metals tend to have a bulk resistivity that depends linearly on T around 300 K, so that $(1/R)dR/dT = (1/\rho) d\rho/dT \approx 1/T$. Naturally, one is then led to consider other possible bolometer materials, especially semiconductors. And indeed, they can be engineered to have values of α ten or more times higher than $1/T$ around 300K. The problem with semiconductors, as will be emphasized later, is that their resistivity tends to be too high to couple efficiently to THz antennas of any type. To achieve a good impedance match, the semiconductor has to be physically large compared to the antenna feed dimension. Again, this is not an issue if the semiconductor bolometer is coupled through an integrating cavity or similar coupling structure, which becomes particularly advantageous in cryogenic operation ($\ll 100$ K) where the α values in semiconductors get much higher than at room temperature. The resulting responsivity of a typical Si or Ge bolometer can be of order 10^7 V/W at liquid helium temperatures (4.2 K). And the external coupling can be maintained usefully high because of the integrating cavity. But suffice it to say that at room-temperature, semiconductor-based bolometers have not been competitive with metallic ones, especially semimetallic ones, as will be addressed later in the chapter.

5.1.2.2. Pyroelectric Detectors

Pyroelectric detectors are also thermal detectors but they rely on a completely different physical phenomenon. Fig. 5.3 shows a schematic of how a typical pyroelectric detector works and is implemented [1]. The detector is based on a crystal in which each unit cell has a built-in dipole moment pointing along a particular axis of the crystal. As shown in Fig. 5.3(a) this creates a net electric polarization along with surface charge densities on opposite ends of the crystal that are normally cancelled by internal charges. The polarization, however, is temperature dependent so a change in the temperature of the crystal will create a transient macroscopic polarization and surface charge while the internal bound charges are moving to a new equilibrium. If the crystal is cut precisely across the axis of polarization and inserted into a circuit, a polarization current will flow when this transient surface charge is created. One practical effect of this type of detection mechanism is that the

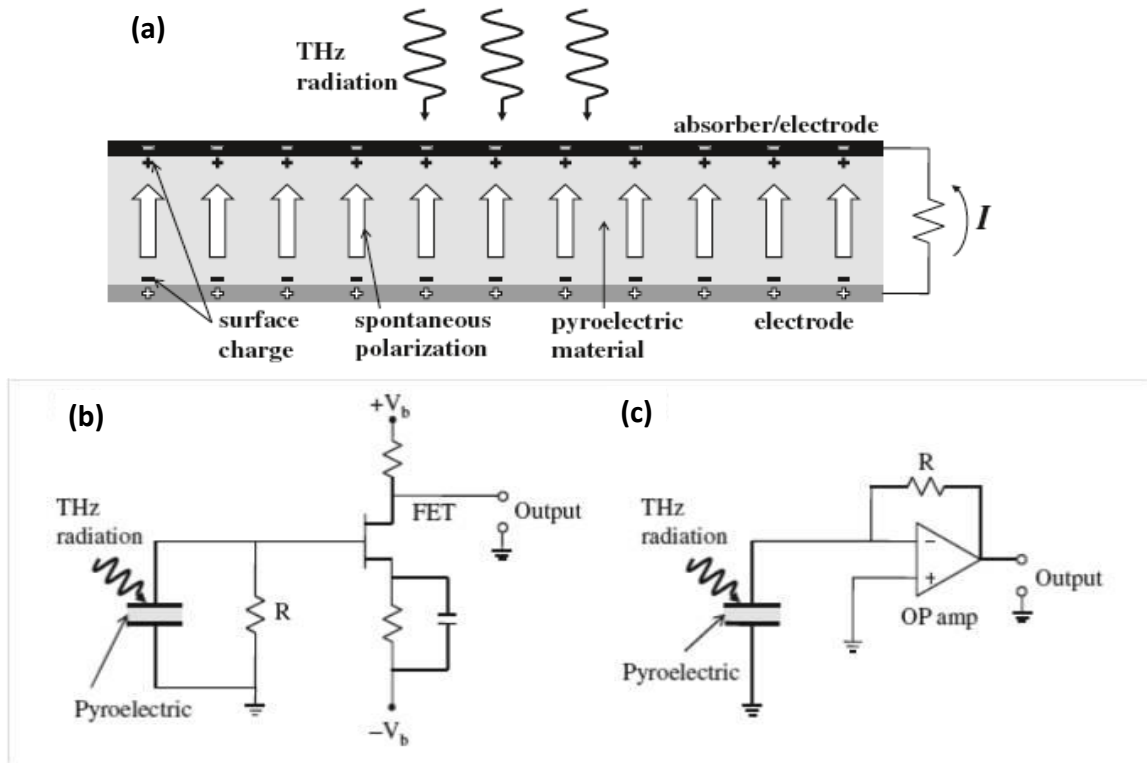


Fig. 5.3: (a) Schematic diagram of a typical pyroelectric detector. Exemplary detection circuits for a pyroelectric detector coupled to (b) voltage and (c) current amplifiers (Ref. [3]).

detector is inherently incapable of responding to a steady radiation source, so the THz radiation must be modulated [4].

A plot of the electric polarization of the crystal (P) vs. temperature (T) shows that the polarization curves downward with increasing temperature, reaching zero at a point known as the Curie temperature. The slope of this graph at any given temperature, $p = dP/dT$, is known as the pyroelectric coefficient. The magnitude of the current created by a particular change in temperature is then

$$I = pA \frac{d(\Delta T)}{dt} \quad (5.18)$$

where ΔT is the change in temperature of the crystal and A is the area of the pyroelectric crystal.

In practice, the current generated is too small to be of much use by itself so some sort of amplification must be included as part of the detector circuitry. This circuitry is of two kinds depending on whether the pyroelectric crystal and the electrodes attached to it are coupled to a current amplifier or a voltage amplifier. Fig. 5.3(b) and (c) shows exemplary circuits for both.

5.1.2.3. Golay (Pneumatic) Detectors

The Golay cell, invented as an infrared detector [5], is sometimes known as a pneumatic detector because the radiation absorbing element heats a small amount of gas that expands as the temperature increases. This gas is contained in a small chamber that has an absorbing film on the front to efficiently transfer the thermal energy from the radiation to the gas and a flexible mirror on the back that moves as the gas expands and contracts. In order to detect the change in volume of the chamber,

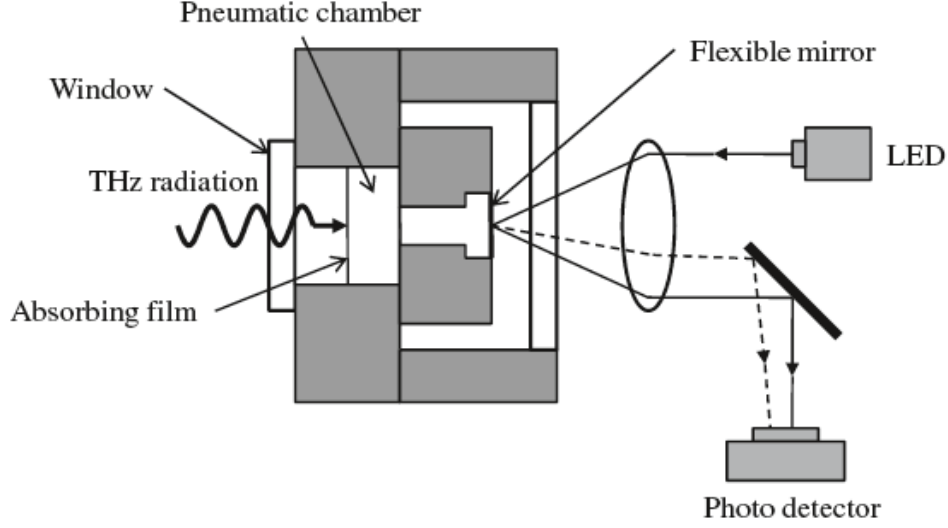


Fig. 5.4: A schematic diagram of a Golay cell (Ref. [3])

an optical system consisting, in essence, of a light source, a line screen, a lens, and a photocell arranged in such a way that changes in the position of the mirror change the amount of light that is incident on the photocell and thus change the output voltage of the photocell. Like some other thermal detectors, the incident radiation must be modulated because of the AC readout circuit of the photocell. Fig. 5.4 shows the block diagram of a Golay cell. Since the Golay cell is broadband and operates by absorbing the incident radiation, it is very sensitive to thermal IR that peaks near $10\text{ }\mu\text{m}$ at room temperature. In order to detect THz, a window is usually used to block IR and pass THz. These windows can be made out of a variety of materials, but they are usually composed of high density polyethylene (HDPE), or diamond [6] with some IR blocking material such as black polyethylene. Of course, the complexity of the Golay cell entails many sources of error and non-uniformity in the manufacturing process, Golay cells are fragile and their performance can be quickly damaged by one careless action, such as exposure to too much radiative power.

5.1.3. Plasma-Wave, HEMT, and MOS-Based Detection

As described first by Dyakanov and Shur [7], the high sheet concentration of free electrons in HEMTs and cryogenic MOSFETs can support plasma wave propagation between the source and drain even at frequencies well above the free-electron transit-time cutoff the device. This is because plasma waves are a collective excitation of the entire electron population, considered as a Fermi fluid. Such excitations generally propagate much faster than the free-electron velocity and do transfer energy at the appropriate group velocity. The remarkable prediction of Dyakonov and Shur was that the 2DEG (2 dimensional-electron-gas) in HEMTs and MOSFETs would create a superlinear response, and therefore a dc term, in the drain-source voltage in response to a purely AC voltage applied between the gate and source. This can be explained more precisely in the following way. The change in gate-source potential of a HEMT affects two parameters of the plasma-wave propagation: (1) the sheet charge density, through the relation $\rho = C_G \cdot V_{GS}$, where C_G is the gate capacitance, and (2) the plasma-wave velocity through the relation, $v = (eV_{GS}/m^*)^{1/2}$ (where e is the electron charge, m^* is the electron effective mass and V_{GS} is the gate-source voltage relative to threshold). Hence, there is a superlinear

dependence of the current density $J = \rho e \cdot v$ on V_{GS} . So if an incident THz signal is applied across the gate-source contacts, a non-zero DC term will occur in the drain-source current, which will manifest as a non-zero DC voltage if the drain-source termination has high-impedance. Experimental realization of this new mode of THz detection will be addressed below in Sec. 5.7.3.

The work from Dyakonov and Schur opened the possibility to use discrete FET transistors as terahertz detectors due to a non-resonant response of the 2DEG. The previous paragraph explained that the conventional low-frequency operation of a cold (non-biased source-drain junction) FET detection could be extended to frequencies well above the FET cut-off frequencies in the THz regime. The real large scale interest on FET as THz detectors comes from the fact of using the well established Si technology of CMOS devices that can integrate in the same wafer the Si-CMOS devices for detection with the read-out electronics allowing room-temperature THz detectors absolutely competitive with the Schottky barrier diodes. The main advantages of Si-CMOS technology are their room-temperature operation, fast response times and easy on-chip integration with read-out electronics leading to forward array fabrication. The detection mechanism can be explained either by plasma wave or by an alternative model of distributed resistive self-mixing that allows a more direct detector design.

Fig. 5.5 shows two possible detection mechanisms for THz radiation. The first one considers the antenna integration while the second one only takes the FET resistive mixer consideration into account. It can be noted that both of them present an asymmetry between the source and drain needed to induce the photoresponse. For the left one the asymmetry is achieved by special antenna connections in balanced antennas such as the bow-tie where the drain reads-out the dc current once the balanced ports of the antenna are connected between the gate and source of the device. For the right one the asymmetry is achieved by incorporating an external C_{gd} capacitor between the gate and drain that allows to take the dc current at the drain once the incident signal is applied between the gate and source. This model can be considered as a quasi-static approach since it is a simplified version of the detection mechanism that will be only valid for GHz frequencies. Unfortunately in "garden-variety" FETs having micron-scale gate length, the intrinsic gate-drain (and gate-source) capacitances are large enough to symmetrize the effect.

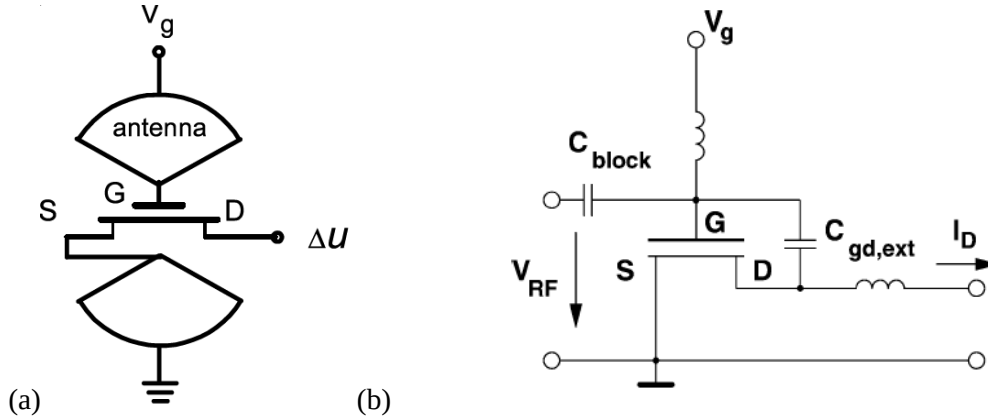


Fig. 5.5: Configurations for plasma-wave detection in FETs. (a) THz configuration directly integrated with a bow-tie antenna; (b) lower-frequency equivalent circuit showing MOSFET detector coupled through an external gate-drain (C_{gd}) capacitor.

The simplified model presented in Fig. 5.5.b tries to explain the detection mechanism but is only valid at GHz frequencies. At THz frequencies, however, the capacitor C_{gd} does not any longer couple the signal between the gate and drain. Then, the approximation shown breaks down and has to be replaced

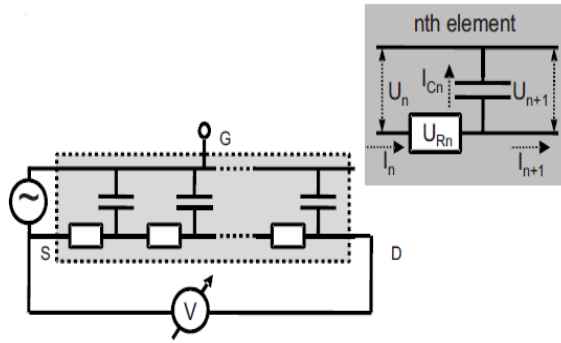
by a non-quasi-static one, according to the theory presented by Dyakonov and Schur. This theory is based on the electron dynamics in the channel and on the continuity equation. For THz frequencies those equations can be simplified as follows

$$\frac{\partial n}{\partial t} + \frac{\partial(nv)}{\partial x} = 0 \quad (5.19)$$

$$v = -\mu \frac{\partial U}{\partial x} \quad (5.20)$$

Where $n(x,t)$ is the electron density, μ is the electron mobility, $v(x,t)$ is the average drift velocity and $U(x,t)$ is the electrical potential along the channel. This reduced pair of differential equations (you can see [8] for the complete set) can be represented in the form of a RC transmission line that describes the non-resonant self-mixing. This can be seen in figure 5.6.a (taken from [8]) where the grey area of the waveguide indicates the channel of the transistor. For this non-resonant case, plasma oscillations are overdamped as shown in Fig. 5.6.b [9].

(a)



(b)

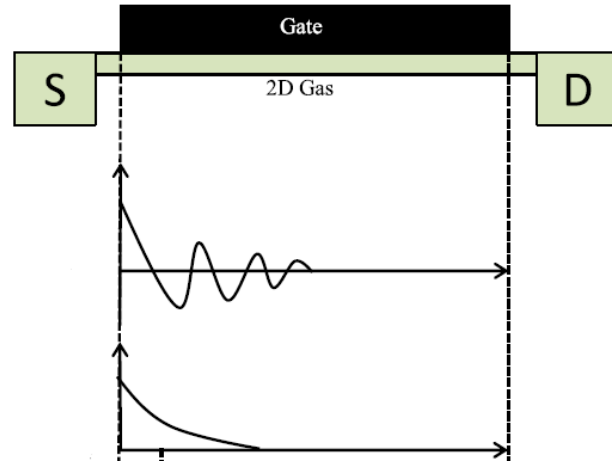


Fig. 5.6: (a) Equivalent distributed circuit for the THz non-resonant self-mixing. (b) Qualitative space dependence of electron oscillations excited by THz signal: upper part for $\omega\tau \gg 1$, bottom part for $\omega\tau \ll 1$

Previous paragraphs state that THz radiation can be coupled between the gate and source of the FET; this THz wave simultaneously modulates the carrier density and its drift velocity leading to nonlinearity and inducing a plasma wave that propagates in the channel. At cryogenic temperatures resonant plasma modes can be excited leading to narrowband tunable detection (for this case the product $\omega_0\tau \gg 1$ where ω_0 is the fundamental plasma harmonic frequency and τ is the momentum relaxation time depending on the carrier mobility). At room temperature plasma waves are over damped and the carrier density decays with a distance of the order of tens of nanometers. For this case the product $\omega_0\tau \ll 1$ and the FET response is a smooth function of frequency and gate voltage and leads to non-resonant detection.

The applied voltage propagates through the channel from the source (left) to the drain (right) showing an exponential damping. It can be said that efficient resistive mixing takes place close to the source while the rest of the device acts as a distributed capacitance and parasitic series resistance. This makes this long-channel device be used for direct power detection at THz frequencies. Then, the response signal can be given as

$$\Delta u = \frac{e \cdot u_a^2}{4m^* s^2} \left[1 - \frac{1}{\sinh^2 Q + \cos^2 Q} \right] \quad (5.21)$$

Where e is the electron charge, u_a is the THz voltage on the gate relative to the source, m^* is the effective mass of the electron, s is the plasma wave velocity and Q is the ratio of the gate length to the characteristic length of the voltage decay from source to drain. After some algebraic operations, in [10], the maximum value of the response signal is given as

$$\Delta u = \frac{e \cdot u_a^2}{4\eta k_B T} \quad (5.22)$$

Where η is a parameter depending on the sub-threshold slope of the channel conductivity and k_B is the Boltzmann constant. From the previous expression the responsivity that relates the detected signal to the THz power can be obtained. It can be seen that, ideally, the responsivity may become exponentially large below threshold. For actual conditions this is not really so since it depends on the antenna coupling and on the loading effects. Thus, as the input impedance of the read-out circuit becomes smaller, due to a simple dividing effect, the detected signal will decrease. A similar effect can be extracted from the imaginary part of the input impedance since a small capacitance can lead to large RC constants. In this way an increase in the modulation frequency can also lead to a decrease in the detected signal.

An alternate model for explaining the detection mechanism at lower frequency in CMOS technology can be given from Fig. 5.5(b) [11]. This model has been chosen for simplicity since the results in the THz band will be qualitatively valid. Then, it can be seen that there is an extra gate to drain capacitor which serves to facilitate the self-mixing [12] [12]. This capacitor allows simultaneous coupling of the input signal to the drain of the FET (acting as the "local oscillator" port) and to the gate (acting as the "RF port"). It can also be seen that a dc gate-bias V_g may be provided through an RF-choke. Thus, the voltages at each port can be written as

$$v_{gs}(t) = v_{RF}(t) + V_g \quad (5.23)$$

$$v_{ds}(t) = v_{RF}(t) = V_{RF} \sin \omega t \quad (5.24)$$

As the device is operated in the linear region, the drain current is obtained through the multiplication between the voltage and the transconductance as

$$i_{ds}(t) = v_{RF}(t) \cdot g_{ds}(t) \quad (5.25)$$

The transconductance, for the quasi-static case, can be written as

$$g_{ds}(t) = \frac{W}{L} \cdot \mu \cdot C_{OX} \cdot \left[\frac{v_{RF}(t)}{2} + (V_g - V_{thres}) \right] \quad (5.26)$$

For the non-quasi-static case, the transconductance per unit length is written as

$$G_{ds}(v(x, t)) = \mu C_{OX} W (v(x, t) - V_{th}) \quad (5.27)$$

By taking the DC term in (5.26) it can be seen that the read-out current is given as

$$I_{ds} = \frac{W}{L} \cdot \mu \cdot C_{OX} \frac{V_{RF}^2}{4} \quad (5.28)$$

The read-out voltage of the detector can be obtained by multiplying the DC current by the channel resistance. This can be written, whenever the drain-to-source voltage is in the linear region, as

$$V_{ds} = \frac{I_{ds}}{G_{ds}} = \frac{V_{RF}^2}{4(V_g - V_{thres})} \quad (5.29)$$

From this expression the electrical (voltage) responsivity can be obtained as

$$\mathcal{R}_V = \frac{V_{ds}}{P_{in}} = \frac{\frac{V_{RF}^2}{4(V_g - V_{thres})}}{\frac{V_{RF}^2}{R_{in}}} = \frac{R_{in}}{4(V_g - V_{thres})} \quad (5.30)$$

The maximum responsivity is obtained in the subthreshold region since the detection current is generated across a larger internal DC resistance. In Si CMOS technology, on-chip amplification close to the detector prevents external capacitive loading avoiding both the reduction of the responsivity and the reduction of the bandwidth.

5.2. Noise Mechanisms

In statistical mechanics and then again in quantum mechanics, we learn that upon measurement every physical quantity, including radiation, is subject to fluctuations. The received power in sensor systems is generally very weak, typically orders-of-magnitude weaker than it is in communications systems. So an important issue is the “masking” of the signal by fluctuations in the power (i.e., the “noise”) in the receiver. The “noise” is the totality of all the electronic and electromagnetic mechanisms, especially those in the detector itself. This section examines these noise mechanisms in the context of THz direct detectors operating at room temperature and in the ambient environment. A fundamental, physical viewpoint is adopted primarily because noise is one of the least understood topics in THz detectors and sensors as a whole. Therefore, this section could also serve for pedagogical purposes in academic courses or professional workshops.

5.2.1. Noise from Electronic Devices

Within every sensor system, particularly at the front end, are components that contribute significant noise to the detection process and therefore degrade the ultimate detectability of the signal. The majority of this noise usually comes from electronics, particularly the first device, which is often a mixer or direct detector. After this there is generally a low-level amplifier that contributes separate noise. The majority of noise from such devices falls in two classes: (1) thermal noise, and (2) shot noise. Thermal noise in semiconductors is caused by the inevitable fluctuations in voltage or current associated with the resistance in and around the active region of the device. This causes fluctuations in the voltage or current in the device by the same mechanism that causes resistance – the Joule dissipation that couples energy between charge carriers and electromagnetic fields. The form of the thermal noise is very similar to that for free-space blackbody radiation. And the Rayleigh-Jeans approximation is generally valid for THz radiation and room temperature operation, so that the Johnson-Nyquist theorem applies. However, one must account for the fact that the device is coupled to a transmission-line circuit or similar, not to a free space mode, and the device may not be in

equilibrium with the radiation as assumed by the blackbody model. These issues are addressed by Nyquist's generalized theorem [13]

$$V_{rms} = [\langle \Delta V^2 \rangle]^{1/2} = [4k_B T_D \text{Re}\{Z_D\} \Delta f]^{1/2} \quad (5.31)$$

where T_D , Z_D , and Δf are the temperature, differential impedance, and bandwidth of the device. Even this generalized form has limitations since it is often difficult to define the temperature of the device if it is operating well away from thermal equilibrium.

Shot noise is a ramification of the device being well out of equilibrium, usually by virtue of electrical bias and some barrier blocking the flow of charge carriers. It is generally described as fluctuations in the current transmitted through the barrier caused by random changes in the rate of carriers incident on the opposite side. Because of the general importance of this phenomenon to THz detectors, and all radiative detectors for that matter, we will examine it in more detail than thermal noise with a similar physical basis as presented in Ref. [14]. From statistical kinetic theory, which generally applies well to charge carriers in non-degenerate semiconductors, we know that there is a mean time τ between collisions, and a mean rate of collisions $1/\tau$ for each carrier. We then assume that the probability of *having* a single collision between time t and $t + \delta t$ is exponential, $P_1(t) = \exp(-t/\tau) \cdot \delta t / \tau$ [15], which follows from the "relaxation time" approximation. This can be generalized to an ensemble of uncorrelated collisions to predict the probability of N collisions between 0 and time t ,

$$P_N(t) = [(t/\tau)^N / N!] \cdot \exp(-t/\tau), \quad (5.32)$$

which is a form of the famous Poisson probability density function. To apply Eq. (5.32) to the shot-noise phenomenon, we think of $1/\tau$ – the collision rate R – also as the the impinging rate on the barrier,

$$P_N(t) = [(Rt)^N / N!] \cdot \exp(-Rt). \quad (5.33)$$

Hence, $P_N(T)$ is to be thought of as the probability of N carriers incident on the barrier in time t . If we limit the ensemble to those carriers with enough kinetic energy to cross the barrier, Eq. (5.33) still applies with N constrained to be consistent with the ensemble-averaged current i through the barrier during measurement sampling time t_s :

$$\langle i \rangle = q \langle N \rangle / t_s \equiv I, \quad (5.34)$$

where q is the signed charge. And consistent with the ergodic hypothesis, this ensemble averaged current should equal the time-averaged current I over the same interval t_s .

The Poisson density function has the remarkable property that the variance is equal to the mean, i.e.,

$$\langle (N - \langle N \rangle)^2 \rangle \equiv \langle (\Delta N)^2 \rangle = \langle N \rangle \quad (5.35)$$

Applying this to Eq. (5.34) and given that t_s is deterministic, we can calculate the variance of the electrical current

$$\langle (i - \langle i \rangle)^2 \rangle \equiv \langle (\Delta i)^2 \rangle = q^2 \langle (\Delta N)^2 \rangle / t_s^2 = (q^2 / t_s^2) \langle N \rangle = q \langle i \rangle / t_s, \quad (5.36)$$

which can also be thought of as the mean-square fluctuation of i during interval t_s . According to sampling theory, $1/t_s$ – the sampling rate – determines the effective bandwidth over which power can

be measured unambiguously according to the Nyquist sampling relation, $B_{max} = 1/(2t_s)$, so that we can write

$$\langle (\Delta i)^2 \rangle = 2q \langle i \rangle \cdot B_{max} \equiv 2qI \cdot B_{max} \quad (5.37)$$

In real devices, t_s is usually dictated by an intrinsic time constant, such as the characteristic transit time of carriers across the barrier, or the device internal RC time in devices for which the transit time is relatively short. In this case, B is replaced by an effective bandwidth B_{eff} as determined by the longest of the internal characteristic times

$$\langle (\Delta i)^2 \rangle = 2qI \cdot B_{eff} \quad (5.38)$$

Eq. (5.38) is the famous Schottky relation [16], discovered even before the Johnson-Nyquist form of thermal noise given above. While an approximation, it has withstood the test of time and found to be accurate in many THz-detector devices, such as Schottky diodes. However, some quantum-effect devices, or devices that can store space charge in the barrier region such as resonant tunneling diodes, have displayed a modified form of Eq. (5.38) given by [17]

$$\langle (\Delta i)^2 \rangle = 2qI \cdot \Gamma B_{eff} \quad (5.39)$$

where Γ is a numerical factor representing the degree to which the random Poissonian fluctuations of transmission times are modified by the transport or charge storage between the cathode (or emitter) and anode (or collector). If $\Gamma = 1$, there is no effect and the anode current has the "full" shot noise. When $\Gamma < 1$, the transport or charge storage reduces the fluctuations, often through some form of degenerative feedback mechanism, and the shot noise is said to be suppressed. When $\Gamma > 1$, the transport or charge storage increases the fluctuations, often through some form of regenerative feedback mechanism, and the shot noise is said to be enhanced.

5.2.2. Phonon Noise

By their physical nature, THz thermal detectors are inherently sensitive to fluctuations in the incident power, be it by radiative coupling (photons) or thermal conductance (phonons) to the environment. Such fluctuations get transduced into electrical noise in the same way that the average absorbed electromagnetic radiation gets transformed into a steady electrical output signal. It is therefore a fundamental matter to estimate what the power fluctuations are for a given thermal detector type in its environment. For THz thermal detectors, especially those operating at room temperature, the most important effect is power fluctuation via phonons.

To estimate the power spectrum of phonon fluctuations, we start with the same kinetic analysis as for carrier shot noise, but this time applied to a volume of phonons incident on a thermal link between the detector at temperature T and the bath. As with the charge carriers, we assume that the incident rate of phonons on the link fluctuates with Poissonian statistics. From fundamental solid-state physics, the mean energy of a phonon is $\langle U(\nu) \rangle = h\nu [\exp(h\nu/k_B T) - 1]$, where ν is the corresponding (atomic lattice) vibrational frequency and h and k_B are Planck's and Boltzmann's constants, respectively. A key part of the analysis is to assume the "classical limit" whereby $h\nu \ll k_B T$, such that $\langle U(\nu) \rangle \approx k_B T$. While arguably valid at room temperature for solids having Debye temperatures < 300 K, this assumption must certainly be corrected for thermal detectors operating at cryogenic temperatures (see

below). Another part of the analysis is that the phonons that arrive at the link from the detector flow ballistically across the link just like the charge carriers in the shot noise analysis. Even in the classical limit, we expect this number to be a function of ν and also of the detector volume V , both of which factor into the 3D acoustical-phonon density-of-states [18].

Given these assumptions, the average power in frequency interval $d\nu$ and volumetric element dV is given as mean energy per phonon times the total number of phonons per sampling time t_s

$$\langle P(\nu, V) \rangle = k_B T_D \langle N(\nu, V) \rangle / t_s \quad (5.40)$$

Given the Poissonion assumption about phonon transport rate, the mean-square fluctuation (i.e., variance) of power is given by

$$\langle [\Delta P(\nu, V)]^2 \rangle = (k_B T_D)^2 \langle [\Delta N(\nu, V)]^2 \rangle / t_s^2 = (k_B T_D)^2 \langle N(\nu, V) \rangle / t_s^2 \quad (5.41)$$

The total thermal power will be the integral over all frequencies and the specific detector volume:

$$\langle P \rangle = \int_0^{\nu_{\max}} \int_0^V \langle P(\nu, V) \rangle d\nu dV = (k_B T_D / t_s) \int_0^{\nu_{\max}} \int_0^V \langle N(\nu, V) \rangle d\nu dV \equiv G T_D \quad (5.42)$$

where G is the macroscopic thermal conductance between the detector and its environment. The corresponding total power fluctuations is,

$$\langle (\Delta P)^2 \rangle = \int_0^{\nu_{\max}} \int_0^V \langle [\Delta P(\nu, V)]^2 \rangle d\nu dV = (k_B T_D / t_s)^2 \int_0^{\nu_{\max}} \int_0^V \langle N(\nu, V) \rangle d\nu dV \quad (5.43)$$

Substitution of the last step of Eq. (5.42) into the last step of Eq. (5.43) yields

$$\langle (\Delta P)^2 \rangle = (k_B T_D / t_s)^2 G \cdot T_D (t_s / k_B T_D) = k_B T_D^2 G / t_s \quad (5.44)$$

Application of the same (Nyquist) sampling argument as for shot noise, namely $B_{\max} = 1/(2t_s)$, yields the result

$$\langle (\Delta P)^2 \rangle = 2k_B T_D^2 G B_{\max} \quad (5.45)$$

Eq. (5.45) requires some clarification and generalization to be practical. Firstly, all thermal detectors have a characteristic thermal time constant, $\tau_T \approx C/G$, where C is the heat capacity. This generally limits the bandwidth to an effective value B_{eff} much lower value than in rectifying devices. Secondly, unlike the analysis of shot noise where the electrical current flow across the barrier was considered unilateral, thermal detectors often have a second significant source of power fluctuations associated with heat flowing from the bath to the detector. Hence, a second term must be added to Eq. (5.45) that is ostensibly uncorrelated to that from the detector to the bath. Hence, we can just add the variances of the two components:

$$\langle (\Delta P)^2 \rangle = 2k_B G B_{\text{eff}} (T_D^2 + T_B^2) \approx 4k_B G B_{\text{eff}} T_D^2 \quad (5.46)$$

where the last step follows when $T_D \approx T_B$, which is often true in room-temperature thermal detectors. Eq. (5.46) is a famous result in thermal-detector theory, often cited in the literature as a general result from thermodynamics. However, from the above simple-minded derivation, it requires statistical transport theory, which is usually beyond the scope of thermodynamics [19].

Even more remarkably, Eq. (5.46) can be generalized to handle obvious shortcomings of the above analysis, such as the cryogenic operating regime where $k_B T < h\nu$ and $T_D > T_B$. Both of these cases were covered by the seminal analysis of Boyle and Rodgers using Planck (massless-Boson) statistics for the phonon population [20]. They showed that

$$\langle (\Delta P)^2 \rangle = 2k_B G B_{eff}(T_D^5 + T_B^5)/T^2 \quad (5.47)$$

where T is the arithmetic mean between T_D and T_B . Note that it reduces to the classic formula when $T_D \approx T_B$, meaning that the physical effect of the Planck statistics is embedded in G , not the temperature dependence. To further see how much physical information is contained in G , we note that Eq.s (5.46) and (5.47) have also been applied successfully for calculating the power fluctuation of thermal detectors exchanging heat by thermal mechanisms other than phonons. Two good examples are thermal conduction by electromagnetic radiation (photons), and thermal convection by fluids, usually air. The challenge, then, is to calculate the physics-rich quantity, G .

5.2.3. Photon Noise with Direct Detection

No analysis of detector behavior anywhere in the electromagnetic spectrum is complete without a discussion of radiation fluctuations. Quantum mechanics teaches us that the act of measuring any physical observable is inherently probabilistic with a measure of uncertainty given by Planck's constant h . Therefore, we expect all radiative sources to fluctuate in power when measured by any detector, possibly at a very low level compared to the other noise sources. The theory of radiative power fluctuations is relatively simple in two extremes: (1) coherent radiation, and (2) thermal radiation. In the THz region, the latter is more important in the case of direct detectors having wide spectral bandwidth. To carry out the analysis, we start by considering n , the number of photons in spatial mode m , as a random variable to calculate the fluctuation in the incident power:

$$\Delta P_{inc} \equiv P_{inc} - \langle P_{inc} \rangle = \sum_m^M \int_{\nu_0}^{\nu_0 + \Delta\nu} h\nu \cdot \{n_m(\nu) - \langle n_m(\nu) \rangle\} d\nu \equiv \sum_m^M \int_{\nu_0}^{\nu_0 + \Delta\nu} h\nu \cdot \Delta n_m(\nu) \cdot d\nu \quad (5.48)$$

An important measure of noise is the variance, or mean-square fluctuation

$$\langle (\Delta P_{inc})^2 \rangle = \langle \left(\sum_m^M \int_{\nu_0}^{\nu_0 + \Delta\nu} h\nu \cdot \Delta n_m(\nu) \cdot d\nu \right) \cdot \left(\sum_{m'}^{M'} \int_{\nu_0}^{\nu_0 + \Delta\nu} h\nu \cdot \Delta n_{m'}(\nu) \cdot d\nu \right) \rangle \quad (5.49)$$

where m' is a dummy summation index. In evaluating this expression we utilize the fact that perfectly random fluctuations in different orthogonal modes are uncorrelated, so that only cross-products having the same mode index will survive the ensemble average

$$\langle (\Delta P_{inc})^2 \rangle = \langle \sum_m^M \left(\int_{\nu_0}^{\nu_0 + \Delta\nu} h\nu \cdot \Delta n_m(\nu) \cdot d\nu \right)^2 \rangle \quad (5.50)$$

This is further simplified by the “band-limited” assumption: $\Delta\nu$ being narrow enough that $h\nu$ and Δn can be considered constant over the range of the integration:

$$\langle (\Delta P_{inc})^2 \rangle \approx \langle \sum_m^M (h\nu_0 \Delta n_m(\nu_0) \cdot \Delta\nu)^2 \rangle = \sum_m^M (h\nu_0 \Delta\nu)^2 \langle [\Delta n_m(\nu_0)]^2 \rangle \quad (5.51)$$

In the special but common case in sensitive THz detectors of only one spatial mode, such as typically occurs with antenna or waveguide coupling, we find

$$\langle (\Delta P_{inc})^2 \rangle = (h \nu_0 \Delta \nu)^2 \langle [\Delta n(\nu_0)]^2 \rangle \quad (5.52)$$

For direct detectors in the THz region, one should consider first the fluctuations of background thermal radiation, given how important these are in the higher-frequency infrared bands. It is a basic exercise of quantum statistical mechanics to show:

$$\langle (\Delta n)^2 \rangle \equiv \langle (n - \langle n \rangle)^2 \rangle = \langle n^2 \rangle - \langle n \rangle^2 = \langle n \rangle (\langle n \rangle + 1) = f_p (1 + f_p) \quad (5.53)$$

where f_p is the Planck function. This expression has an interesting low-frequency, or Rayleigh-Jeans limit, where $h \nu \ll k_B T$, $f_p \gg 1$ and $\langle (\Delta n)^2 \rangle$ goes to f_p^2 , which is approximately $(k_B T / h \nu_0)^2$. This limit is quite accurate in the THz region at terrestrial temperatures. Substitution into Eq. (5.52) then yields:

$$\langle (\Delta P_{inc})^2 \rangle = (k_B T \Delta \nu)^2 \quad (5.54)$$

And we can write

$$P_{RMS} = [\langle (\Delta P_{inc})^2 \rangle]^{1/2} = k_B T \Delta \nu \equiv \langle P_{abs} \rangle \quad (5.55)$$

We observe the remarkable result that for perfect coupling ($\eta = 1$), the rms absorbed power fluctuation is exactly equal to the average power – a sign of exponential (Boltzmann) statistics! Since the power spectral density is uniform in frequency (i.e., "white"), it is then given by:

$$S_P(\nu) = P_{RMS} / \Delta \nu = k_B T, \quad (5.56)$$

a form of the famous Johnson-Nyquist theorem for one mode in free space.

Although fundamental and interesting, the above derivation serves just one purpose for the present chapter, which is predicting the maximum possible *output* signal fluctuations from a THz direct detector in the presence of a 300 K background having power spectrum in Eq. (5.56). To do this, we take advantage that all of the common THz detectors are essentially "square-law" devices, meaning that they convert incident power into output electrical signal according to Eq. (5.1):

$$y = A x^2 = \mathcal{R} P_{in} \quad (5.57)$$

where y is output voltage or current. This is true even for thermal detectors, for which x is related to the incident electric field magnitude, so that x^2 is proportional to the incident power. The key to understanding the resulting fluctuations in y is to realize that different frequency components of Gaussian incident thermal noise, say at ν and ν' , are statistically uncorrelated and therefore produce a zero DC output from the detector after averaging. However, they can still contribute a fluctuation and therefore, output noise, if their difference frequency $\Delta \nu$ is within the detector electrical bandwidth. So we need to calculate the output power spectral density.

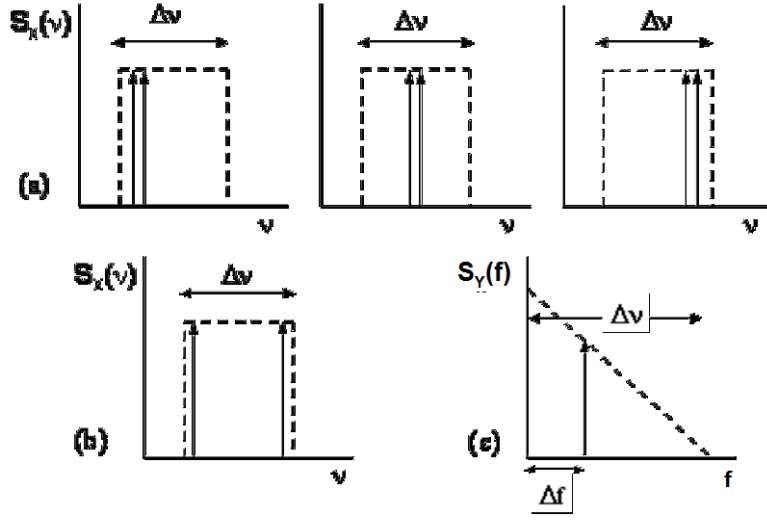


Fig. 5.7: (a) Graphical representation of input power spectrum to a square-law detector created by thermal incident radiation over bandwidth $\Delta\nu$, and showing separate frequency components. (b) Same as (a) but with separate frequency components at edges of input band. (c) Output power spectrum from square-law detector as a function of difference frequency $\Delta f = |\nu_2 - \nu_1|$, and highlighting the specific difference frequency Δf .

This is a classic calculation of nonlinear signal-processing and probability theory carried out elegantly, for example, by Davenport and Root [21]. They prove that a square-law detector changes the noise statistics from Gaussian at the input to "chi-squared" at the output. For our purposes, a more qualitative method can be applied that misses the statistics but approximates the output-signal mean-square fluctuations, which suffices to get the signal-to-noise ratio and noise-equivalent power [22]. We draw the input power spectrum as a THz-band-limited white spectrum as shown in Fig. 5.7(a). If we look at small difference frequencies, there are a maximal number of input frequency components available, proportional to the input THz bandwidth $\Delta\nu$. But with increasing frequency, the number of available input components drops until we reach the minimal condition depicted in Fig. 5.7(b) where just the extremes of the input spectrum are effective. The resulting output spectrum is then given approximately by

$$S_y(f) \approx 2\Re^2(S_x)^2 \Delta\nu (1 - f/\Delta\nu) \quad (5.58)$$

where S_x is the input power spectral density at any frequency in the passband, f is the post-detection frequency, and the factor of 2 accounts for the fact that the input power spectrum has an identical interpretation at negative frequencies (not shown in Fig. 5.7). As plotted in Fig. 5.7(c), Eq. (5.58) is a distinctly-non-white triangular spectrum centered at zero frequency – *a distinctly non-Gaussian result*. But since the input noise is "white," we can write the fluctuations as:

$$(S_x)^2 \approx \langle (\Delta P_{in})^2 \rangle / \langle (\Delta\nu)^2 \rangle \quad (5.59)$$

Substitution into Eq. (5.58) then yields

$$S_y \approx 2\Re^2 \langle (\Delta P_{in})^2 \rangle (1 - f/\Delta\nu) / \Delta\nu \quad (5.60)$$

The quantity we need is the mean-square fluctuation in output signal y , which is given by:

$$\langle(\Delta y)^2\rangle = \int_0^\infty S_y(f)df \approx 2\Re^2(\Delta P_{in})^2 \int_0^{\Delta\nu} \frac{1-f/\Delta\nu}{\Delta\nu} df \quad (5.61)$$

For practically all THz detectors, the output electrical bandwidth Δf is much less than the THz bandwidth, which entails limiting the above integral to a range 0 to $\Delta f \ll \Delta\nu$. Under this condition, the rather slowly varying triangular spectrum can be treated like a constant at the peak value. We then get the result,

$$\langle(\Delta y)^2\rangle \approx 2\Re^2(\Delta P_{in})^2 \int_0^{\Delta f} \frac{df}{\Delta\nu} \approx 2\Re^2(\Delta P_{in})^2 \frac{\Delta f}{\Delta\nu} \quad (5.62)$$

Given terrestrial THz thermal radiation in the Rayleigh-Jeans limit, we can use Eq. (5.53) to re-write this as

$$\langle(\Delta y)^2\rangle = 2\Re^2(k_B T)^2 \Delta\nu \cdot \Delta f \quad (5.63)$$

As we should have expected, the mean-square output-signal fluctuation is proportional to the mean-square input power fluctuation, but also to the responsivity-squared and the product of the input and output bandwidths. So it is highly dependent on the detector type, and we will return to this later on.

5.3. THz Coupling

Devices operating at terahertz frequencies become inseparable from the circuit. Therefore, issues relating to coupling the device to electromagnetic radiation, to antenna structures for electromagnetic radiation and to device integration with THz circuits have to be addressed for practical consideration for the terahertz coupling. Since the device dimensions are often much smaller than the electromagnetic wave length, the antenna structure needed for coupling the device with the electromagnetic radiation has to be designed as a unique entity inseparably comprising both elements the device and the antenna. In the microwave regime this concept has been studied and defined as the active integrated antenna concept. Traditionally, antenna design for these receivers has been inefficient in terms of matching it with the diode. The antenna is often designed to be 50 Ω matched while the capacitive part of the diode is compensated including a RF filter or RF matching network with an inductive part. The Active Integrated Antenna (AIA) concept was proposed to optimize the matching between the antenna and the active device. In addition to the matching optimization, a further improvement on the overall antenna efficiency should be considered by increasing the total efficiency which is essential within the THz-band. As a consequence, the responsivity of the receiver working within the THz-band will be optimized.

However the coupling design is done, some degree of impedance mismatch inevitably occurs and the power delivery from free space to the detector is imperfect. So a useful metric is an external power coupling factor, η , analogous to the external quantum efficiency in photonic devices. It is defined simply by

$$P_{abs} = \eta P_{inc} \quad (5.64)$$

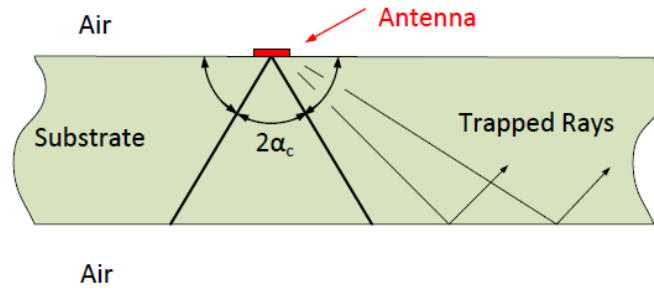


Fig. 5.8: Antenna on a dielectric substrate showing the rays trapped as substrate modes. The critical angle is represented as α_c .

where P_{inc} is the power incident on the device (or equivalently, the “available” power from free space).

5.3.1. THz Impedance Matching

The amount of power at THz frequencies is generally quite small so the coupling efficiency is a key factor that has to be taken into account. This analysis has to be undertaken in a two-fold way. On one hand the matching between the active device and the antenna itself has to be considered. On the other hand the matching and coupling between different media, guided and free-space, has to be taken into account. The first case will be developed in section 5.3.2. while the second one will be developed in this section.

The analysis of antennas embedded or just laying on the interface of two electrically different media has been carried out over a broad range of THz research [23,24]. The interest has increased in recent years due to, for example, the potential for new THz-generation devices based on the use of the substrate semiconductor properties, thereby avoiding the use of antennas. Antennas lying over a dielectric medium have some special features that distinguish them from antennas in free-space. In free-space, the power radiated from a planar antenna divides equally above and below the plane by symmetry, while an antenna on a dielectric mainly radiates primarily into the dielectric. This results both from the way in which elementary sources radiate and from the way in which waves propagate along metals at a dielectric interface. It can be explained by the impedance of both media. Let us assume that a dipole is placed on the interface between air and a dielectric. If the dipole is used as a receiving antenna it measures the electric field collinear to it. This electric field is the field transmitted through the interface. In the transmission-line model of wave propagation, a wave from a high impedance material (air) is incident on a low-impedance material (dielectric). The transmitted electric field is small, so the dipole response is small. The effect is the same as looking at the voltage across a coaxial cable terminated with a low resistance. When the wave is incident from the dielectric side, the transmitted electric field is large and the dipole response is large, since the material has a lower impedance than the air. The analogous situation in coaxial cable is the voltage seen at a high-impedance load. The result is that the response is larger when the wave is incident from the dielectric. By reciprocity, the power transmitted from the dipole is larger in the dielectric.

Another issue that has to be taken into account is the thickness of the substrates at THz frequencies which is on the order of a wavelength. Under this condition, the substrate is relatively thick and its effect has to be considered. From the ray point of view (Fig. 5.8), the rays incident with an angle larger than the critical angle are completely reflected and trapped as substrate modes.

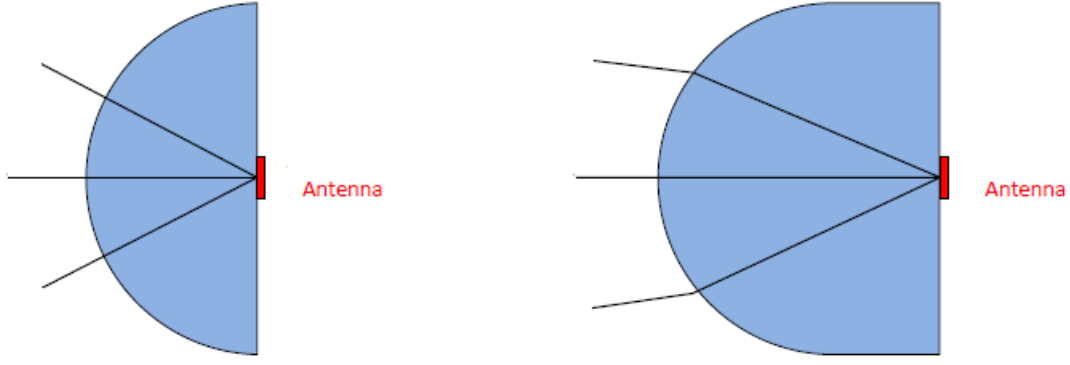


Fig. 5.9: Hemispherical (left) and hyperhemispherical (right) lenses.

These substrate modes can be suppressed by the inclusion of a lens on the back side of the substrate with the same dielectric constant. This eliminates the problem, because, as shown in Fig. 5.9, the rays are then incident nearly normal to the surface and do not suffer total internal reflection. The substrate lens takes advantage of the sensitivity of an antenna to radiation from the substrate side and eliminates the substrate modes. In addition, directivity is improved. The disadvantages of the substrate lens are those of any system using refractive optics: absorption and reflection loss. Among the available dielectric lenses, one is particularly attractive because it is aplanatic, adding no spherical aberration or coma. This is the extended (or hyper-) hemispherical lens with an extension of $d=r/n$.

5.3.2. Planar-Antenna Coupling

In the THz regime the detector often consists of a planar antenna integrated with a device (e.g., a zero-bias Schottky diode) together with a silicon lens [24] to suppress substrate modes due to the thick substrate and making the radiation pattern unidirectional. From the antenna point-of-view three parameters must be accounted for to determine the responsivity of the detector: the radiation propagation efficiency (ϵ_{rad}), impedance-matching efficiency (M), and the polarization matching efficiency (ϵ_{pol}):

$$\eta = \epsilon_{rad} \cdot M \cdot \epsilon_{pol} \quad (5.65)$$

This total radiative coupling efficiency expresses how close the performance is to ideal. Maximizing this efficiency will enhance the performance of the detector and it is a key factor that has been used to optimize the design of THz direct detectors integrated with planar antennas.

For lens-coupled planar antennas, the propagation efficiency is limited primarily by reflection-from and attenuation-in the lens. If it is made from high-resistivity silicon, as is common practice, the losses are usually negligible but the reflection can be substantial unless some form of anti-reflection coating is applied to the air-silicon interface. For incident radiation in the form of a Gaussian beam, the majority of power is incident on the Si lens near normal incidence so that the power reflectivity is given by $|I|^2 = [(n-1)/(n+1)]^2$, and $\epsilon_{rad} \approx 1 - |I|^2$, where n is the THz refractive index of silicon. This is well-known in the THz region, $n \approx (11.65)^{1/2} = 3.41$ [25], which leads to $|I|^2 = 0.30$, and $\epsilon_{rad} = 1 - |I|^2 = 0.70$.

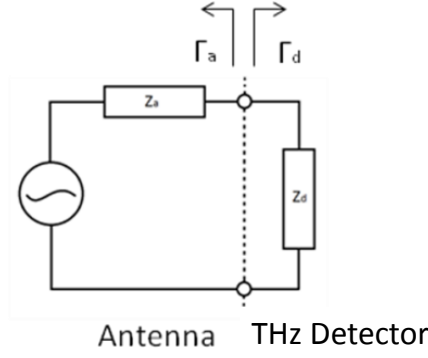


Fig. 5.10: THz-frequency impedance-matching view of the Schottky rectifier.

The impedance matching efficiency at the THz signal frequency can be analyzed as the ratio between the power delivered to the detector acting as a passive load (P_L) and the power available from the antenna terminals acting as a generator (P_{avs}):

$$M = \frac{P_L}{P_{avs}} = \frac{4R_D R_A}{(R_D + R_A)^2 + (X_D + X_A)^2} \quad (5.66)$$

where $Z_D = R_D + jX_D$ is the detector impedance and $Z_A = R_A + jX_A$ is the antenna impedance (see Fig. 5.10). When $M=1$, all the available power from the source is delivered to the load and it can be obtained the well-known conjugated matching condition which states that the maximum power transferred to the load is obtained when $Z_A = Z_D^*$ (conjugate match, i.e., $\Gamma_A = \Gamma_D^*$). The M -factor is a key parameter to estimate the performance of the detector, especially for highly reactive detectors like Schottky rectifiers. As will be shown later, typical Schottky rectifiers have much higher values of R_D than the R_A of the antenna, and also have significant capacitance making them very difficult to achieve the conjugate impedance matching condition. This tends to make M low at THz frequencies, of order 1% being typical.

5.3.3. Exemplary THz Coupling Structures

We have mentioned that planar antennas are now the preferred way of coupling room-temperature THz direct detectors to free space. Alternative methods such as metal-waveguide or integrating-cavity coupling also work and are actually older and therefore widespread in the literature. However, they are also tend to be more difficult (and expensive) to fabricate, can be multimodal, and do not lend themselves to proliferation of detector elements in focal-plane (imaging) arrays. Which is why we do not discuss them here. So in this section we graphically showcase two paradigm planar-antenna coupled detector structures, the first coupled to free space with a low-loss lens and the second coupled with a metal ground plane.

Fig. 5.11 shows perhaps the first successful THz planar-antenna structure: a microbolometer coupled to a self-complementary antenna. As will be emphasized later, in principle this structure can provide

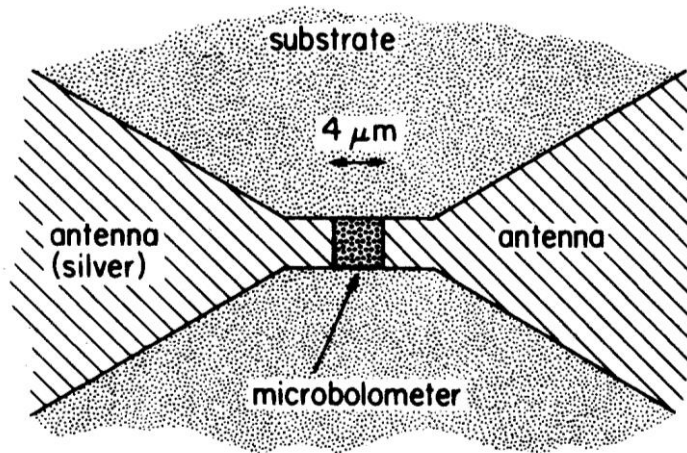


Fig. 5.11: Schematic of a microbolometer mounted at the driving gap of a planar antenna, Ref. [30].

near-perfect impedance matching ($M \approx 1$) between the thin-film metallic or semi-metallic microbolometer and the antenna on a dielectric substrate. A thin film of metal has very little reactance – just some kinetic inductance – at THz frequencies, so this structure should also have very wide bandwidth approaching 1.0 THz or beyond. And it integrates readily with hemispherical or hyperhemispherical lenses mounted on the back-side of the bolometer/antenna substrate as shown in Fig. 5.12(a), which also exemplifies the antenna as a self-complementary log-spiral – one of the widest-band planar antennas ever demonstrated.

On inspection of Fig. 5.12(b), we also see another beneficial feature of the planar antenna coupling, which is electrical isolation of the THz antenna circuit from the readout circuit. Notice that the readout circuit is coupled to the outer extremity of the spiral antenna after one full turn of the spiral. This means that the THz radiation will be coupled primarily to the driving gap, the readout circuit being isolated by the low-pass filtering properties of the spiral. So the rectified signal will flow through the spiral to the read-out arms relatively unimpeded. It also means that the readout circuit can be tailored to a specific load impedance, independent of the THz antenna. Shown in Fig. 5.12(a) is a standard 50 Ω coplanar waveguide (CPW) line, so the inclusion of an off-the-shelf, low-noise video output amplifier is straightforward. Shown in Fig. 5.12(c) is the same coupling structure but with a Schottky rectifier mounted in the driving gap. This presents more impedance-matching challenge than the microbolometer, but still works well in the sub-THz region.

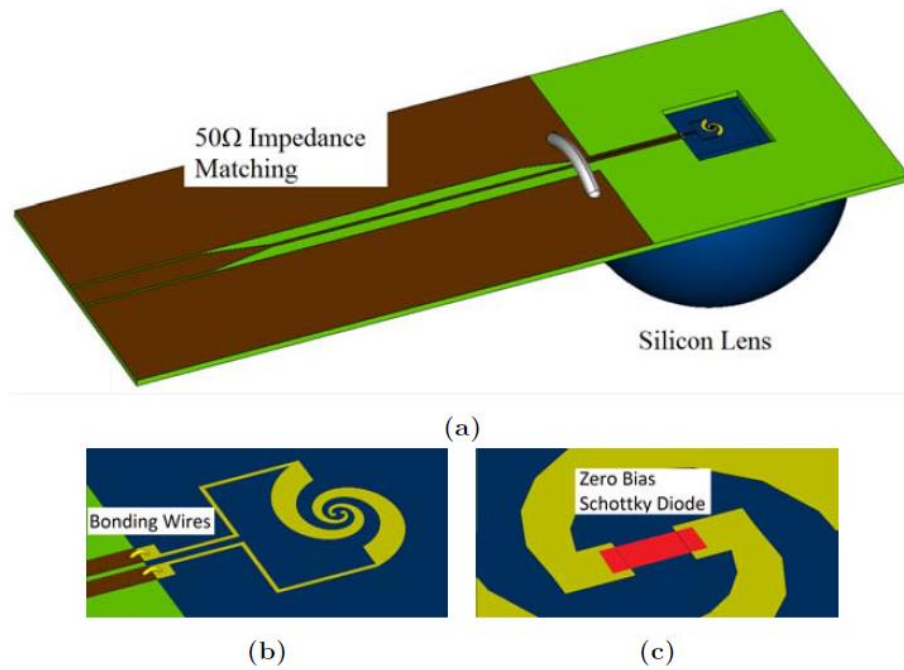


Fig. 5.12: Schematic of the Schottky rectifier quasi-optical QO detector with log-spiral antenna. Materials: dark blue part is the silicon lens and silicon substrate where the gold (yellow part) planar antenna is grown, while the green part is FR-4 material and the brown part is copper. The bonding wires are included to connect gold lines over silicon to copper lines over FR-4 and a metallic wire is inserted to interconnect both ground planes of CPW line. The Zero Bias Schottky diode is epoxied in the middle of the antenna. (a) Complete SBD QO Video Detector 3D schematic. (b) Antenna zoom. (c) Schottky diode zoom.

Fig. 5.13 shows the planar-antenna coupled with a ground plane. The antenna is a resonant patch and the detector is a CMOS-type similar to those discussed in Section 5.1.3, both designed for 300 GHz operation [26]. Two differential feeds occur between the patch and the CMOS detector, and each feed is recessed into the patch – a common practice with patch antennas to provide for impedance transformation and better overall coupling efficiency [27]. Being Si-CMOS based, this structure lends itself very well to proliferation into arrays via the super-successful technology of VLSI (very large scale integration). However, one drawback of this approach compared to Fig. 5.12 is the antenna pattern. Patch antennas generally provide much lower gain than lens coupled antennas of all types, and suffer from substrate-mode and other loss mechanisms when fabricated on the large dielectric constant ($\epsilon_r = 11.65$) of silicon substrates.

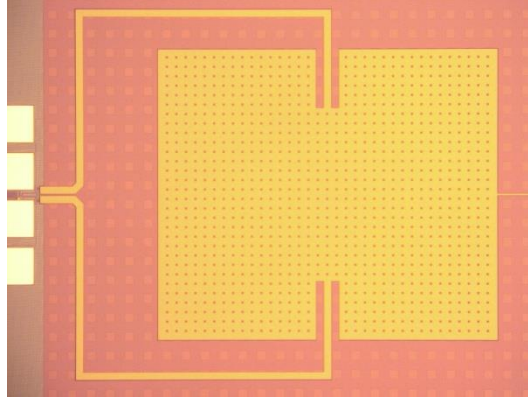


Fig. 5.13: Differential patch antenna integrated with a CMOS detector

5.3.4. Output-Circuit Coupling

Finally, there is the issue of output circuit coupling. While at first appearing to be a trivial issue, it can be relatively easy to lose signal strength (and signal-to-noise ratio) at the output owing to voltage or current division between the differential resistance of the detector and the input resistance of the low-noise amplifier connected to it. This is perhaps best shown by example, the Schottky rectifier once gain. Using Eq. (5.8) we can write the external responsivity of the rectifier as $\mathfrak{R}_{\text{ext}} = \eta \mathfrak{R}_V = \eta e R_D / (2 \alpha k_B T)$, which directly shows the presence of the diode differential resistance, typically of order 10 k Ω . If we now connect the rectifier to a load resistance (no reactance), there will then be a voltage divider effect such that the fraction across the load relative to the (open-circuit) voltage across the detector will reduce the external responsivity to

$$\mathfrak{R}_{\text{ext}} = \eta \frac{e R_D}{2 \alpha k_B T} \cdot \left[\frac{1}{1 + \frac{R_S}{R_D} + R_S R_D (\omega C)^2} \cdot \frac{R_L}{R_D + R_L} \right] \quad (5.67)$$

where ω is the angular output frequency and R_L is the load resistance. Unless high video bandwidth is required (say >10 MHz), the term in square brackets is close to unity since a high-input-impedance operational amplifier can be used having $R_L > 1 \text{ M}\Omega$ – a much higher value than practical Schottky rectifiers.

5.4. External Responsivity Examples

The combination of electrical responsivity introduced in Section 5.1 and external coupling efficiency form a crucial combination called the external, or optical, responsivity

$$\mathcal{R}_{\text{ext}} = \eta \mathcal{R}_X \quad (5.68)$$

where X is either current or voltage. In the small-signal operation that THz detectors usually experience, these two parameters are often separable, making the optical responsivity the most important metric with respect to signal processing (but not necessarily noise). We will now revisit each of the common detectors types from Section 5.2 with simultaneous consideration of coupling efficiency and electrical responsivity. Unfortunately, much of the THz literature on direct detectors does not make this distinction, instead limiting the analysis to electrical responsivity. So our discussion will be limited to the detector types best understood at this point in time, which are the rectifiers and the bolometers. By so doing, we will see design trade-offs that ultimately must be made to optimize THz detector performance – a task that is generally more complicated than it first appears.

5.4.1. Rectifiers

As mentioned in Section 5.1., rectifiers generally are the simplest of all THz direct detectors, which is why they are so popular. The general analysis done for impedance matching in Section 5.3.3. can be made more specific by considering the equivalent circuit model shown in Fig. 5.14 [2].

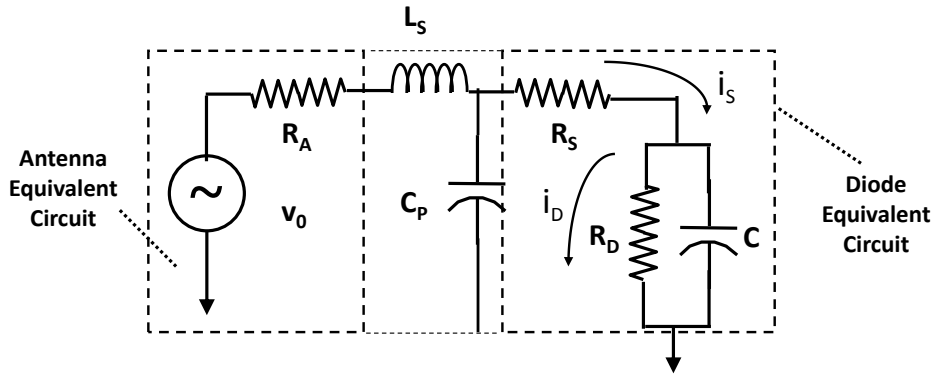


Fig. 5.14: Equivalent circuit planar-antenna-coupled Schottky rectifier.

It is a bit more sophisticated than Eq. (5.66) through the explicit addition of a diode series resistance R_S which becomes increasingly important as the frequency increases from the microwave bands toward the THz regime.

$$M = \frac{4 \cdot R_D R_A}{(R_D + R_A + R_S - \omega^2 L C R_D)^2 + [\omega L + \omega R_D C (R_A + R_S)]^2} \quad (5.69)$$

It is usually the sum of all dissipative components in the semiconductor outside the depletion layer of the Schottky diode itself. The model also includes parasitic elements L_S and C_P , which represent a

series inductance and packaging capacitance, respectively. Generally L_S is more important than C_P in THz air-bridge type planar devices where the finger that contacts anode is not contacting the high-permittivity GaAs (or InP) substrate. In that case we can assume $C_P \approx 0$ and can solve analytically for the impedance matching efficiency where R_A is the real part of the antenna driving-gap impedance. This ignores any reactive part of the antenna impedance, which is a fair assumption for self-complementary-type planar antennas such as bow-ties and log spirals.

It is informative to calculate Eq. (5.69) for typical values of all parameters at an interesting THz frequency – 650 GHz (which happens to be one of the highest-frequency useful atmospheric "windows". We do this for a zero-bias rectifier as shown in Fig. 5.1 for 20% - Al composition [28]. The diode anode area is $1 \mu\text{m}^2$. The I-V and R-V curves are shown in Fig. 5.15(a). The corresponding zero-bias differential resistance and capacitance are approximately $7.5 \text{ k}\Omega$ and 2.0 fF , respectively. The antenna resistance is assumed to be 72Ω consistent with the theoretical value according to Booker's relation for self-complementary antennas on a GaAs substrate: $R_A = \eta_{\text{eff}}/2$ where $\eta_{\text{eff}} \approx [\mu_0/\epsilon_{\text{eff}}]^{1/2}$, $\epsilon_{\text{eff}} = (1+\epsilon_r)/2$, and $\epsilon_r = 12.8$ for GaAs. For R_S we assume two values – 10 and 30Ω – which roughly defines a range of possible values that depends

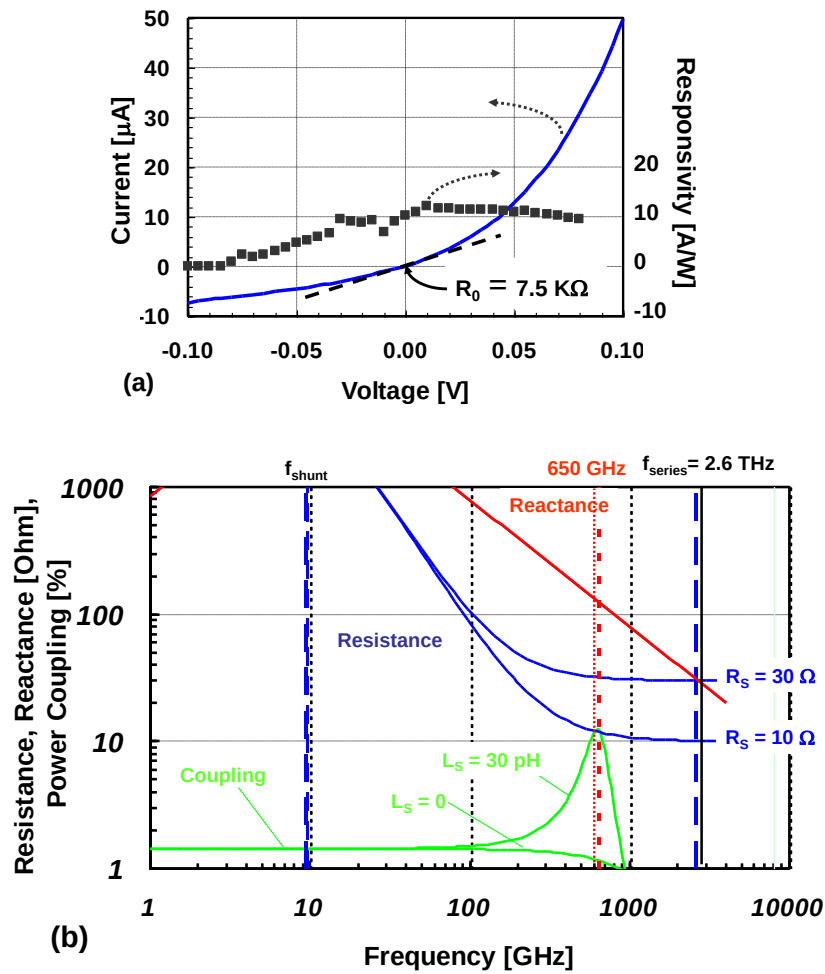


Fig. 5.15: (a) I-V and current electrical responsivity of Schottky rectifier. (b) THz impedance-matching coupling efficiency of Schottky according to equivalent circuit in Fig. 5.13.

on the specific fabrication technology. For L_S we consider 0 pH – an idealization, and 30 pH – a value found to resonate with the device (capacitance) at 650 GHz.

The resulting coupling efficiency is plotted in Fig. 5.15(b) where we see an interesting behavior vs frequency. For both $L_S = 0$ and 30 pH, and practically independent of R_S , we see that the coupling is approximately 1.5% up to 100 GHz. For $L_S = 0$ pH, it drop at higher values down to $\approx 1.0\%$ at 650 GHz. Hence, given the highest zero-bias electrical responsivity of ≈ 18 A/W from Fig. 5.1, the maximum external responsivity is ≈ 0.18 A/W at 650 GHz. For $L_S = 30$ pH, the coupling rises above 100 GHz to a peak value of $\approx 10\%$ at 650 GHz. Hence, the peak external responsivity at 650 GHz is 1.8 A/W. This is considered excellent coupling for Schottky rectifiers in spite of the limited bandwidth (~ 300 GHz) caused by the resonance. In contrast and calculated next, antenna-coupled bolometers can display coupling efficiencies approaching 100% and over bandwidths in excess of 1 THz. But because the zero-bias Schottky's are so "quiet" in physical noise relative to bolometers, the noise-equivalent power of the Schottkys remains superior.

5.4.2. Micro-Bolometers

As described in Section 5.1.2.1, bolometers have a great deal of design latitude through their electrical responsivity $\mathcal{R}_V \approx I_B R_0 \alpha / G_0$. But what happens if we also consider the external coupling efficiency so that $\mathcal{R}_{ext} = \eta I_B R_0 \alpha / G_0$? This generally makes the optimization complicated except in the case of the antenna- or waveguide-coupled "microbolometer", as developed in the 1980s. By definition, a microbolometer is physically much smaller than a wavelength but can still couple efficiently to THz radiation by integration with a waveguide or planar antenna. As described in Ref. [30], this leads to a simplification of the design if the bolometer is metallic because it should then display the same resistance R_0 both at THz and low (read-out) frequencies. Impedance matching then entails setting R_0 to the characteristic impedance Z_0 of the waveguide or antenna (a value generally less than 500Ω), so that $\eta \approx 1$ and $\mathcal{R}_{ext} \approx I_B Z_0 \alpha / G_0$. Aside from controlling I_B , the optimization then focuses simply on maximizing the ratio α / G_0 . As discussed previously, α has approximately the same absolute value in all metallic solids, so the design emphasis focuses on the reduction of G_0 along with reduction in heat capacity C so that the time constant $\tau = C / G_0$ remains usefully small.

An important development in this vein occurred in the 1980s with the air-bridge microbolometer. The air-bridge entails lifting the thin-film bolometer element off the solid substrate by elegant microfabrication techniques, thereby eliminating the thermal conduction path to the substrate [29]. This constrains the thermal conduction to heat flow into the metal planar antenna to which the bolometer element is necessarily connected. And assuming operation in the ambient environment, there is also some convection to the surrounding air. Assuming that the conduction to the antenna metal dominates (an assumption that must be checked in each air-bridge microbolometer design), a simple but insightful analysis can be carried out taking advantage of the fact that the air-bridge bolometer metal usually has rectilinear dimensions l (length), w (width), and t (thickness). In this case, the DC (and THz) resistance is $R_0 = \rho l / (w \cdot t)$ where ρ is the bulk electrical resistivity. The thermal conductance is a more complicated parameter since the temperature is not uniform across the bridge but rather peaks in the middle. Analysis has led to the approximate form $G_0 \approx 12 \cdot K \cdot w \cdot t / l$ where K is the bulk thermal conductivity [30]. Substitution of the electrical expression into the thermal expression yields

$$G_0 = 12 \cdot K \cdot \rho / R_0 \quad (5.70)$$

Now setting $R_0 = Z_0$ for the impedance matching leads to a rule-of-thumb to minimize G_0 and thereby maximize the electrical responsivity without jeopardizing the external responsivity:

$$G_0 = 12 \cdot K / (\sigma \cdot Z_0) \quad (5.71)$$

where σ is the electrical conductivity. So minimizing G_0 entails minimizing the ratio K/σ , which for bulk metallic solids is governed by the famous Wiedemann-Franz law of solid-state physics [14]:

$$K/\sigma = (\pi^2/3)(k_B/e)^2 \cdot T \quad (5.72)$$

However, nm-scale thin films generally display a lower value of K/σ than the Wiedemann-Franz prediction because the electrical conductivity drops faster with decreasing thickness than the thermal conductivity [31]. So we write

$$K/\sigma = d \cdot (\pi^2/3)(k_B/e)^2 \cdot T \quad (5.73)$$

where d is a metal- and dimension-dependent factor > 1 . Evaluation at 300 K yields $d \cdot 7.32 \times 10^{-6}$ W- Ω /K. Substitution of $Z_0 = 72 \Omega$ for a self-complementary antenna on GaAs then yields

$$G_0 = 12 \cdot d \cdot K / (\sigma \cdot Z_0) = 1.22 \times 10^{-6} \cdot d \text{ [W/K]} \quad (5.74)$$

For the sake of estimation, we set $d = 2.0$, which yields

$$G_0 = 2.45 \times 10^{-6} \text{ [W/K]} \quad (5.75)$$

Substitution of $\alpha \approx 1/T$ then yields for the external responsivity from Eq.s (5.13) and (5.75)

$$\mathcal{R}_{ext} \approx 12 \cdot I_B Z_0 \alpha / G_0 = 9.8 \times 10^4 \cdot I_B \text{ [V/W]} \quad (5.76)$$

where $\eta = 1.0$ (perfect coupling) has been assumed.

The external responsivity can now be estimated considering the limitations on I_B imposed by regenerative electrothermal feedback or overheating as discussed in Section 5.1.2.1. We make a plausible assumption that the maximum temperature rise ΔT of the microbolometer with respect to ambient is 100 K, which should occur at the geometric center of the device. All of the heat is created by the bias power, $P_B = (I_B)^2 R(T)$ which we set to $(I_B)^2 R_0$ for the sake of estimation. Assuming further that G_0 is independent of temperature, we can write $\Delta T = P_B / G_0 = (I_B)^2 R_0 / G_0 = (I_B)^2 Z_0 / G_0$, where the last step follows from antenna impedance matching condition. Solving for I_B using the nm-scale-corrected G_0 from above, we find, $I_B \approx 1.8$ mA – a typical value of microbolometer bias current in air-bridge microbolometers, but somewhat smaller than that in substrate-based devices which generally have higher G_0 so can support more current before overheating. Substitution of this into the external responsivity above yields $\mathcal{R}_{ext} \approx 180$ V/W. By comparison, the original air-bridge bisimulth bolometer displayed a dc responsivity of ≈ 100 V/W [29].

How then do THz researchers choose the metal for their microbolometers? The answer lies largely in practical considerations. To fabricate a metal air-bridge, the thickness must be large enough to obtain adequate strength and uniform electrical conductivity, but not so great that the microlithographic

fabrication becomes too difficult. A ballpark minimum thickness is 20 nm (2×10^{-6} cm). For similar reasons, the geometry of the bridge is roughly square, such that $w = l$. Hence, the bolometer can be assumed to be one "square" of metal roughly 20 nm thick. The issue is then to match to the antenna impedance Z_0 to such a square, i.e. $\rho/t \approx Z_0$, or $\rho \approx 1.4 \times 10^{-4} \Omega \text{ cm}$. Inspection of the metals commonly used in microfabrication, such as gold and aluminum, yields values of $\rho = 2.2 \times 10^{-6}$ and $2.7 \times 10^{-6} \Omega \text{ cm}$, respectively [16]. Other transition metals have higher resistivity (e.g., $\rho = 12.6 \times 10^{-6} \Omega \text{ cm}$ for chromium) but still nowhere near the target value. Semi-metals get much closer (e.g., $\rho = 1.1 \times 10^{-4} \Omega \text{ cm}$ for bismuth), so were the bolometer material of choice in early development of air-bridge microbolometers. However, many transition metals display significantly higher effective resistivity than the bulk values in such thin films, so some have been pursued (e.g., $\rho = 14.5 \times 10^{-6} \Omega \text{ cm}$ in niobium but roughly twice this value in 20 nm films [32]).

5.5. System Metrics

5.5.1. Signal-to-Noise Ratio

A useful metric for all types of sensors is the power signal-to-noise ratio (SNR).

$$\frac{S}{N} = \frac{\langle P \rangle}{\sqrt{\langle (\Delta P)^2 \rangle}} = \frac{\langle P \rangle}{S_P \cdot B_N} \quad (5.77)$$

where S_P is the power spectral density and B_N is the equivalent noise bandwidth at that point in the sensor

$$B_N = (G_{\max})^{-1} \int_0^{\infty} G(f) df \quad (5.78)$$

where $G(f)$ is the sensor gain function vs. frequency, and $G_{\max}$ is the maximum value of this gain B_N is generally dictated by sensor phenomenology, such as the resolution requirements and measurement time.

While at first appearing to add insurmountable complexity to sensor analysis, a great simplification results from the fact that radiation noise and two forms of physical noise discussed above are, in general, statistically Gaussian (the shot noise becomes Gaussian in the limit of large samples, consistent with the central limit theorem). A very important fact is that *any Gaussian noise passing through a linear component or network remains statistically Gaussian*. Hence, the output power spectrum S in terms of electrical variable X (current or voltage) will be white and will satisfy the important identity

$$\langle (\Delta X)^2 \rangle = \int_{f_0}^{f_0 + \Delta f} S_X(f) \cdot df \approx S_X(f) \cdot \Delta f, \quad (5.79)$$

where Δf ($\equiv B_N$) is the equivalent-noise bandwidth. Then one can do circuit and system analysis on noise added by that component at the output port by translating it back to the input port. In the language of linear system theory, the output and input ports are connected by the system transfer function $H_X(f)$, so the power spectrum referenced back to the input (reference) port becomes

$$S_x(in) = \frac{S_x(out)}{|H_x(f)|^2}. \quad (5.80)$$

Because the different Gaussian mechanisms are statistically independent, the total noise at the reference point can be written as the uncorrelated sum

$$\langle (\Delta X_{TOT})^2 \rangle = \sum_{j=1}^N \langle (\Delta X_i)^2 \rangle \text{ or } \langle (\Delta P_{TOT})^2 \rangle = \sum_{j=1}^N \langle (\Delta P_i)^2 \rangle \quad (5.81)$$

5.5.2. Sensitivity Metrics

5.5.2.1. Noise Equivalent Power and Its Properties

Unfortunately, noise figure (or factor) concept commonly applied in RF systems is not applicable or particularly useful to THz direct detectors since it presupposes linearity. THz direct detectors are inherently nonlinear elements in RF and THz systems, which is what allows them to extract information from incoming radiation in the form of a dc or low-frequency modulated signal. In this case, it is the SNR *at the point of detection or decision* that is generally the most important quantity in system performance. And because of reasons pertaining to statistical detection theory, generally this SNR must be greater than or equal to unity to have a reliable detection or decision. Therefore, a very useful metric for direct detection performance is to fix the “after detection” SNR, SNR_{AD} , at unity and then solve for the signal power *at the input to the system* sensor that achieves this. The resulting metric is the noise-equivalent power NEP which most simply put, is the *input* signal power to the sensor required to achieve a SNR of unity at the *output*.

A simple example is helpful at this point. Suppose we have an ideal “square-law” detector, such as the Schottky-diode rectifier or a metal-film microbolometer. By definition, the square-law detector has a circuit transfer function of

$$X_{out} = \mathcal{R}_X P_{in} \quad (5.82)$$

where X_{out} is the output signal (usually current or voltage), P_{in} is the input power, and $\mathcal{R}_X$ is the “responsivity”. The noise at the output of the detector is minimally given by the Nyquist generalized theorem, which for a Schottky rectifier is

$$P_N \propto \langle (\Delta i)^2 \rangle = 4k_B T_0 \text{Re}\{Y_D\} \Delta f \approx 4k_B T_0 \Delta f / R_D \quad (5.83)$$

where Δf is the post-detection bandwidth. The corresponding signal power at the output is just $(X_{out})^2$, so that the output SNR (with X chosen as i) is

$$SNR_{AD} = (\mathcal{R}_I P_{in})^2 / (4k_B T_0 \Delta f / R_D) \quad (5.84)$$

Setting this SNR to unity and solving for P_{in} , we get

$$NEP_{AD} = (4k_B T_0 \Delta f / R_D)^{1/2} / \mathcal{R}_I \quad (5.85)$$

For the purpose of comparing different sensor technologies, it is conventional to divide out the post-detection bandwidth effect (or equivalently, setting it equal to 1 Hz). This yields the normalized NEP_{AD} [in units of $W/(Hz)^{1/2}$], given by

$$NEP'_{AD} = NEP_{AD} \cdot \sqrt{\frac{1}{\Delta f}} = (4k_B T_0 / R_D)^{1/2} / \mathcal{R}_I \quad (5.86)$$

We deduce the generic expression for a square-law rectifier including the external power coupling factor:

$$NEP'_{AD} = [\langle (\Delta X_{out})^2 \rangle / \Delta f]^{1/2} / (\eta \mathcal{R}_X) \quad (5.87)$$

Although simple looking, there is a lot of interesting physics buried in this expression!

An equally fundamental effect is the NEP limited by incident radiation (photon) fluctuations in a single spatial mode according to the analysis conducted in Section 5.2.3. According to Eq. (5.63), $\langle (\Delta y)^2 \rangle = 2 \mathcal{R}_{ext,y}^2 \cdot (k_B T)^2 \Delta \nu \cdot \Delta f$, where y is the current or voltage. The average signal squared is simply $(y_s)^2 = (\mathcal{R}_{ext,y} \cdot P_{inc})^2$, so that the $NEP = k_B T (2 \cdot \Delta \nu \cdot \Delta f)^{1/2}$ and the specific “background limited” NEP is

$$NEP'_{AD} = k_B T (2 \cdot \Delta \nu)^{1/2} \quad (5.88)$$

As a “worst case” scenario, we imagine a very broadband THz detector such a microbolometer coupled to a self-complementary spiral antenna. In this case we can assume $\Delta \nu \approx 1.0$ THz, so that

$$NEP'_{AD} = 5.8 \times 10^{-15} \text{ W/Hz}^{1/2} \quad (5.89)$$

This is so much lower than the typical NEP' values in real THz detectors that it begs the question of why we bother deriving and analyzing it. The answer is physical completeness and demonstration of the richness of the NEP metric. If we were to do the same calculation for room-temperature detectors operating at higher frequencies into the infrared bands, we would find out that the background-limited NEP'_{AD} increases dramatically since the blackbody spectral density increases so rapidly (as ν^2). By 30 THz or so (10 μm wavelength), the background-limited NEP' begins to exceed common electrically- or thermally-limited NEP 's under room-temperature operation, making it the primary concern in the design of infrared sensors. Similarly, if one were to consider cryogenic operation at THz frequencies, then the electrically- or thermally-limited NEP s could drop down to values approaching $10^{-14} \text{ W/Hz}^{1/2}$ or lower (especially in bolometers), making the 300 K radiation noise comparable and requiring use of cold filters to reduce this noise component. For students or new researchers in the THz field, this trade-off can be quite mysterious because few if any books or review articles explain it as done here.

A useful system aspect of the NEP is its additivity. If there are N mechanisms contributing to the noise at a given node in the receiver, if the mechanisms are uncorrelated to the signal and to each other, if they obey Gaussian statistics, then the total NEP is the uncorrelated sum

$$NEP_{TOT}^2 = NEP_1^2 + NEP_2^2 + \dots + NEP_N^2 \quad (5.90)$$

This property applies to any node, pre- or post-detection, and will be used explicitly when we discuss the contribution from electronic noise. In reality there are cases where a noise mechanism is correlated to the signal (e.g., radiation noise) or to another noise mechanism (current and voltage noise in transistors), so one must be careful in applying this addition formula. In such cases, one can always fall back on the SNR as a useful measure of overall system performance. A good example

occurs when the rms power fluctuations associated with heat (phonon) transport are comparable to the electrical NEP of the detector. In this case, the system NEP after detection is given by:

$$\text{NEP}_{AD} \approx \eta^{-1} [\langle (\Delta P_T)^2 \rangle + (\text{NEP}_{elect})^2]^{1/2} \quad (5.91)$$

where P_T is the coupled thermal power.

5.5.2.2. Two Examples of THz Noise Equivalent Power

As a first practical example of the NEP metric, we consider a zero-bias, room-temperature Schottky rectifier coupled to a planar antenna. As described earlier, we know from Schottky-barrier physics that $\mathcal{R}_I \approx 25$ A/W or less. Let's assume $\mathcal{R}_I = 20$ A/W and $R_D = 8$ k Ω , and that the post-detection noise is dominated by Johnson-Nyquist fluctuations. We then can write $\text{NEP}'_{AD} = (4k_B T_0 / R_D)^{1/2} / (\eta \mathcal{R}_I)$ to obtain $\text{NEP}'_{AD} \approx \eta^{-1} \cdot 7.0 \times 10^{-14}$ W/(Hz)^{1/2} – a performance obviously dependent on power coupling. For the 8 k Ω Schottky diode operating at 650 GHz, we can apply the results shown in Section 5.4.1 whereby the resonant and non-resonant coupling efficiencies were approximately 10% and 1%, respectively. The corresponding NEP'_{AD} values are 7.0×10^{-13} and 7.0×10^{-12} W/Hz^{1/2}, respectively, which roughly straddle the best experimental results ever achieved in this frequency region for Schottky rectifiers.

As a second example of the NEP metric, we consider an air-bridge microbolometer designed according to the principles given above in Section 5.4.2, specifically with perfect THz impedance matching. This means that $\mathcal{R}_{ext} \approx 12 \cdot I_B Z_0 \propto G_0 = 9.8 \times 10^4 \cdot I_B$ [V/W] = 180 V/W for $I_B = 1.8$ mA, independent of THz frequency. We consider two separate cases: (1) a Johnson-Nyquist-noise limited NEP, and (2) a phonon-noise-limited NEP. For the first case, we have $\text{NEP}'_{AD} = (4k_B T_0 R_D)^{1/2} / \mathcal{R}_{ext} = (4k_B T_0 Z_0)^{1/2} / (180 \text{ V/W}) = 6.0 \times 10^{-12}$ W/Hz^{1/2}. For the phonon noise-limited case, we have $\text{NEP}'_{AD} = [4k_B (T_0)^2 G_0]^{1/2} = 3.5 \times 10^{-12}$ W/Hz^{1/2}. This example is simplified in that it ignores some practical noise effects in bolometers, especially $1/f$ noise and ambient air fluctuations. So the best experimental NEP values are higher than these fundamental limits (see Section 5.7). Nevertheless, this example shows that the phonon noise can be the fundamental limiting mechanism when operating at room temperature, as opposed to Johnson-Nyquist noise in rectifiers.

5.5.2.3. Noise Equivalent Temperature Difference

For radiometric and thermal imaging systems, it is often convenient to express the sensitivity in terms of the change of temperature of a thermal source at the input that produces a post-detection SNR of unity. The resulting metric is the noise equivalent temperature difference NETD, also called the noise equivalent change of temperature NE Δ T. Regardless of title, it is given mathematically by

$$\text{NETD} = \text{NEP}_{AD} / (dP_{inc}/dT|_{PB}) \text{ or } \text{NETD}' = \text{NEP}'_{AD} / (dP_{inc}/dT|_{PB}) \quad (5.92)$$

where P_{inc} is the incident power and P_B is the background power. If we assume that the receiver accepts just one spatial mode with power coupling η (valid for antenna-coupled detectors), and that the thermal source satisfies the Rayleigh-Jeans limit and fills the field-of-view of the sensor, then $P_{inc} = k_B T_B \Delta\nu$ and $P_{abs} = \eta k_B T_B \Delta\nu$ so the NETD' becomes

$$\text{NETD}' = \text{NEP}'_{AD} / (k_B \cdot \Delta\nu) \quad (5.93)$$

where $\Delta\nu$ is the spectral equivalent noise bandwidth, not to be confused with the corresponding electrical quantity Δf given by Eq. (5.78). Clearly the NETD metric favors detectors having large spectral bandwidth, such as bolometers coupled to broadband self-complementary planar antennas. The NETD metric brings some subtleties with respect to its accurate characterization. First is knowledge of $\Delta\nu$, which technically is given by

$$\Delta\nu \equiv (\eta_{max})^{-1} \int_0^\infty \eta(\nu) d\nu. \quad (5.94)$$

In most THz detectors, $\Delta\nu$ is not well known and very difficult to measure because $\eta(\nu)$ is a complicated function of frequency, especially for highly impedance-mismatched detectors like rectifiers. Furthermore, the external NEP'_{AD} may not be known either because of the lack of a powerful-enough THz coherent source to accurately measure it at any (spot) frequency ν_0 , and therefore a lack of knowing η in Eq. (5.87). So researchers sometimes will resort to measurement of NEP'_{AD} and NETD by direct characterization with a THz blackbody. It is then important to be sure that the incident power from the blackbody is known accurately. This is particularly true for non-antenna coupled detectors where the number of spatial modes may exceed unity and therefore the denominator of Eq. (5.92) may significantly exceed $k_B T \cdot \Delta\nu$ as assumed in Eq. (5.93). Because most THz thermal imaging is intended for the terrestrial environment, it is then best to do the NEP'_{AD} and NETD characterization with a room-temperature blackbody having brightness temperature as-close-as possible to 300 K.

Another subtlety of the NETD characterization is electrical bandwidth. For imaging applications, the output of the detector will be often be sampled at a time $t_{\text{samp}} = 1/30$ s to accommodate the frame-rate standard (30 frame/s) generally considered to be compatible with comfortable human viewing. This leads to the common usage of a *reference* NETD [59], NETD'' , given by the NEP'_{AD} of Eq. (5.92) and (5.93) measured specifically at an electrical center frequency of 30 Hz, but multiplied by $[1/(2t_s)]^{1/2}$ - the equivalent electrical bandwidth. Clearly this will always lead to a value of NETD'' [units of K] significantly higher than the NETD' [units of $\text{K}/(\text{Hz})^{1/2}$].

5.6. Effect of Amplifier Noise

No discussion of THz detectors would be complete without addressing the effect of the essential low-noise amplifier (LNA) connected to the output. In fact this applies to direct detectors in general, independent of wavelength or operating temperature. It's a wonderful thing to have a high-responsivity, low-NEP detector, and quite another to have it integrated into a working sensor system and achieve comparable sensitivity. LNAs necessarily introduce their own noise which adds to the detector noise and degrades the overall NEP in a manner that depends on the specific amplifier noise mechanisms and the detector impedance. We will focus only on voltage-amplification LNAs, such as operational amplifiers (opamps), because their relative simplicity of analysis compared to other types (e.g., transimpedance amplifiers). But our conclusions will be rather universal and hopefully convince the reader that the LNA performance should be considered early on when pursuing THz detectors in systems.

As well known in discrete-electronics, voltage amplifiers can often be represented by the simple equivalent circuit shown in Fig. 5.16 [33] where the input impedance of the amplifier is assumed infinite and all the noise it creates is represented by two uncorrelated, white noise sources connected at the input. The voltage noise density generator $V_{N,A}$ has zero internal impedance and is quantified in units of $V/Hz^{1/2}$, uniform across the amplifier bandwidth. The current noise density generator $I_{N,A}$ has infinite impedance and is quantified in units of $A/Hz^{1/2}$. Assuming that the LNA bandwidth extends down to near DC, the THz detector can be represented by its low-frequency differential resistance.

This is generally valid for rectifiers, bolometers, and other resistive-type detectors, which minimally generate Johnson-Nyquist noise as discussed above and represented by the noise generator $V_{N,D} = (4k_B T_0 R_D)^{1/2}$, uncorrelated with the two amplifier noise sources. In this situation, the total voltage

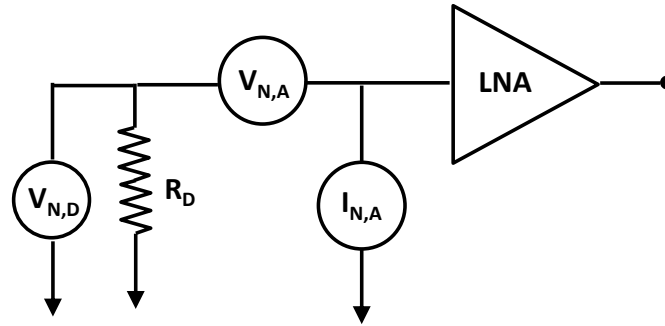


Fig. 5.16: Noise equivalent circuit of THz detector connected to opamp-LNA.

noise generated at the output of the detector (where the NEP_{AD} is referenced) is given by:

$$V_{N,tot} = [(4k_B T_0 R_D) + (V_{N,A})^2 + (I_{N,A} \cdot R_D)^2]^{1/2} \quad (5.95)$$

And the impact on the NEP' is given by

$$NEP' = V_{N,tot} / \mathcal{R}_{ext} \quad (5.96)$$

To proceed further we have to get specific. We choose a popular low-frequency, low-noise opamp that the author has used with low-noise THz rectifiers and bolometers for about 10 years. It has the following characteristics $V_N \approx 1.0$ nV/Hz, $I_N \approx 2.0$ pA/Hz (above ~ 100 Hz) [34]. We can then evaluate Eq. (5.95) using the detector dynamic resistance as a parameter. The resulting plot is shown in Fig. 5.17(a) where we show the detector noise alone and the total noise including LNA contributions. The total noise is very close to the detector noise alone at a resistance $R_D = 500 \Omega$, which is equal to V_{NA}/I_{NA} , and called the amplifier noise resistance R_N . This is the source resistance into the LNA that creates the best "noise match," thereby minimizing the effect of the LNA. At much lower R_D , the total noise is dominated by the LNA voltage noise, and at much higher R_D it is dominated by the LNA current noise. This behavior persists in the analysis of the NEP', as calculated for Eq. (5.96) and plotted in Fig. 5.17(b), except now we need to include the detector external responsivity $\mathcal{R}_{ext}$ as an additional parameter. We include four such values in the plot, 10, 100, 1000, and 10,000 V/W in the plot, which spans the values usually provided by THz room-temperature direct detectors. Again, the minimum impact of the amplifier noise occurs at $R_D = R_N$, but at much lower or higher values there is significant degradation of the NEP' rather independent of $\mathcal{R}_{ext}$. The best Schottky rectifiers are described reasonably well by the $\mathcal{R}_{ext}=1000$ V/W curves, and the best microbolometers by the 100 V/W curves.

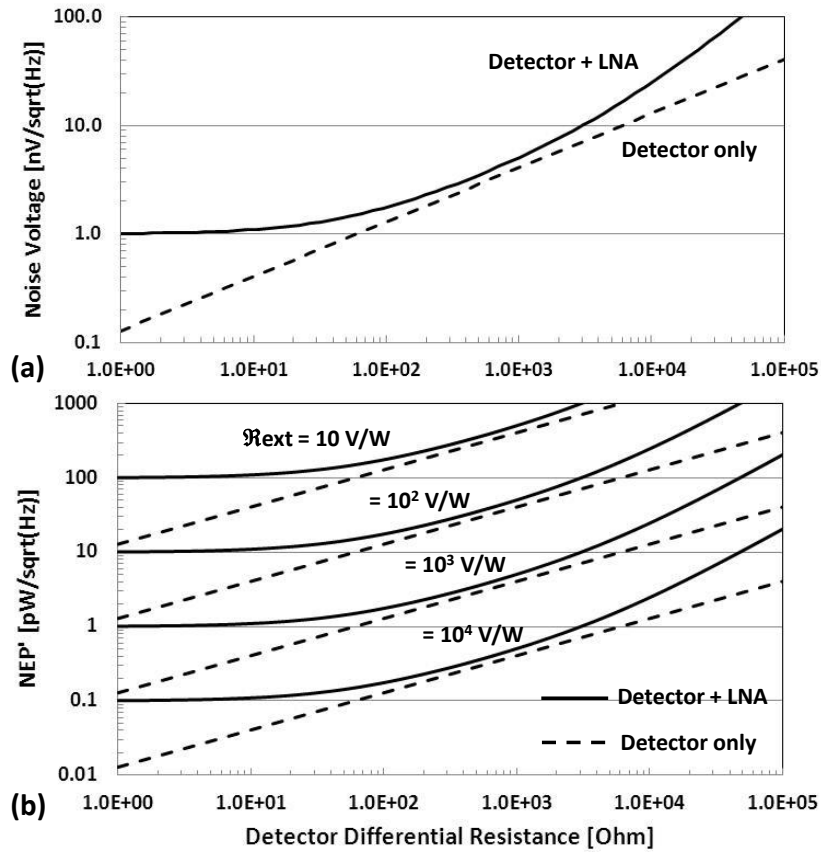


Fig. 5.17: (a) Noise characteristics with THz detector connected to an opamp-LNA assuming noise characteristics of the AD797 (above ~ 100 Hz) and plotted against detector differential resistance. (b) Specific NEP' values for THz detector vs its differential resistance given the AD797 noise characteristics and parameterized by detector external responsivity.

5.7. A Survey of Experimental THz Detector Performance

In this section we tabulate experimental results on THz room-temperature direct detectors. Strong preference is given to results that include measured values of NEP or NETD. As this Chapter has proposed, at THz frequencies external or optical responsivity is far more telling than internal or electrical responsivity, and many publications only report the latter. Furthermore, responsivity is only part of the story in THz detectors, NEP being far more important from a sensor-system standpoint. As shown in the following Tables, there is plenty of experimental NEP data for rectifier-type detectors, and even more for microbolometers, much of the microbolometer data being over 10 years old. The results tabulated for CMOS-based and plasma-wave detectors are recent and rapidly growing, testament to the fact that the CMOS technology is more generic and widely available. The data for pyroelectric detectors and Golay cells is the most sparse of all, which is mostly why we have not analyzed their performance in any detail. In fact, the 530 GHz data points at the end of each Table was obtained by the author in cross-calibration experiments conducted several years ago.

The very best results to date for each detector type are compared in the bar-chart of Fig. 5.18. In composing this, a decision had to be made about the test frequency. We chose to seek frequencies as close-as possible to 650 GHz as this is the center frequency of a well-known atmospheric window – perhaps the last one useful for terrestrial sensing before the atmosphere becomes prohibitively lossy. All values shown are external NEP's, meaning that they include the effects of the external coupling efficiency η . Interestingly, all the Si-CMOS or SiGe based devices are quite close to each other, hovering around $50 \text{ pW/Hz}^{1/2}$. In contrast the best pyroelectric detector and Golay cell lie just above and below $1 \text{ nW/Hz}^{1/2}$ which begs the question why they are so popular. The answer lies in their ease-of-use and their benchmarking utility. Both have physical wide apertures (5 or 6 mm) so it is quite easy to focus THz radiation into them and be assured that the majority of power is being collected (although not necessarily absorbed). Given this benefit and their $\sim \text{nW}$ sensitivity, they are useful devices to characterize THz sources of all types, which generally needs to be done with a signal-to-noise ratio of at least 10 to be confident in the output power measurement. This means that

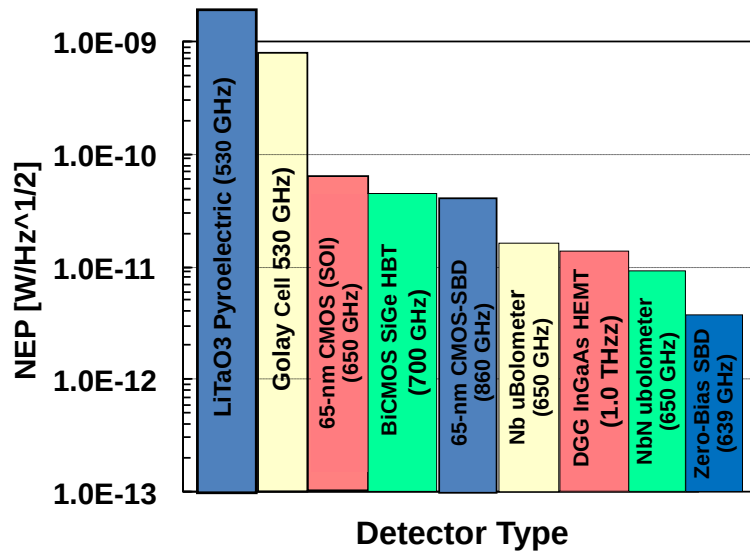


Fig. 5.18: Comparison of the best room-temperature direct detectors operating > 300 GHz and as close to 650 GHz as possible (SOI = silicon on insulator; HBT = heterojunction bipolar transistor; SBD = Schottky barrier diode); DGG = dual grating gate.

the pyroelectric and Golay cell can establish whether or not a given source is putting out ~ 10 nW or more – an average power level thought to be minimally useful in THz sensor systems.

Also in the Tables a strong preference is given to reported detector results in the conventional THz region, 300 GHz and above, motivated large by this frequency being the low-frequency edge of the older "submillimeter-wave" region. There are many excellent detector results in the literature that claim-to or allude-to THz technology impact based on experimental results measured in W-band (75-110 GHz) – one of the popular "millimeter-wave" bands for many decades. This approach is often dictated by the availability of source, control, and calibration components and instrumentation (usually in rectangular metal waveguide) in W band, but not at any higher frequency. The problem with this is two-fold. First, one of the great challenges with THz detectors is radiative coupling efficiency, which tends to drop rapidly as the frequency increases and for a variety of reasons. Some detector types (e.g., rectifiers) are more susceptible to this than others (e.g., bolometers), but there are always fundamental issues such as skin-effect losses, planar-antenna feed losses, etc, that come to bear as the frequency is increased from ~ 100 GHz to ~ 1.0 THz.

The second problem is that W-band direct detectors are of increasingly questionable importance because of technological competition. Integrated circuit technologies (both GaAs- and InP-based MMIC, Si RFIC, and SiGe) all work rather well up to 100 GHz already, with InP-based MMICs operating well above 300 GHz. One of the key components from all of these is transistor-based LNAs with useful levels of gain (> 10 dB) and noise figure (< 6 dB). Given affordable and commercially available RF LNAs, the likelihood of engineers building sensor systems with direct-detection front-ends diminishes. This is particularly true for the Si-based RFICs that can rather easily be proliferated into 1D- or 2D-receiver arrays.

Similarly, there are many publications and sometimes allusion to THz impact of mid-infrared measurements of direct detectors, especially thermal types in the LWIR (8-12 μm band). Again the reason for many such measurements has to do largely with the availability of instrumentation, such as QCL and CO₂ lasers, sensitive power meters, and video-rate uncooled imaging arrays in the LWIR region. The problem in extending these results to THz wavelengths also includes fundamental issues such as radiative coupling. For example, microbolometers in the LWIR can be constructed in which the absorbing element and the thermistor element are one and the same. The dimensions of both can be made small enough for high sensitivity (low G) and high speed (small C/G), but large enough for diffraction-limited coupling to free-space optics in the LWIR. The diffraction-limited coupling then becomes a big problem in the THz region.

5.7.1. Rectifiers

Table I. Performance Characteristics of Room Temperature Rectifying Detectors				
Device Type	Electrical Responsivity	Conditions	NEP [$\text{W/Hz}^{1/2}$]; NETD (where provided)	Ref
ErAs/InGaAlAs Schottky in square spiral planar antenna		zero bias, 639 GHz, 300 K	4.0E-12; NETD = 120 mK	[28]
Zero-bias Schottky diode, waveguide mounted	4,000 V/W at 100 GHz to 400 V/W at 900 GHz	zero bias	1.5E-12 near 150 GHz, 20E-12 at 800 GHz	[35]
Zero-bias InGaAs Schottky diode with log-spiral antenna	~ 200 V/W for system, estimate 10^3 V/W intrinsic for the diode	0.8 THz	5.0E-10	[36]
GaAs Schottky with traveling wave antenna	200 V/W at 337 μm , 20 V/W at 119 μm		0.1E-9 and 1.0E-9	[37]
Silicon diode thermal detector array	5.3 V/W		NETD = 1.7 $^{\circ}\text{C}$	[38]
VDI Model: WR2.8 ZBD	1500 V/W	260-400 GHz	2.7E-12	[39]
VDI Model: WR2.2 ZBD	1250 V/W	325-500 GHz	3.2E-12	[39]
VDI Model: WR1.9 ZBD	1000 V/W	400-600 GHz	4.1E-12	[39]
VDI Model: WR1.5 ZBD	750 V/W	500-750 GHz	5.1E-12	[39]
VDI Model: WR1.2 ZBD	250 V/W	600-900 GHz	16E-12	[39]
VDI Model: WR1.0 ZBD	200 V/W	750-1100 GHz	20E-12	[39]
VDI Model: WR0.8 ZBD	100 V/W	900-1400 GHz	40E-12	[39]
VDI Model: WR0.65 ZBD	100 V/W	1100-1700 GHz	40E-12	[39]
W-InSb point contact diode	25-30 V/W	313 μm , zero bias, room temp	2.0E-09	[40]
GaAs Schottky diode	2000 V/W	214 μm		[41]
Ni-NiO-Ni MoM diode		3 kHz mod 98 GHz & 28.3 THz	50E-12 and 72E-12	[42]

5.7.2. Thermal Detectors

5.7.2.1. Bolometers

Table II. Performance Characteristics of Room Temperature Bolometers				
Device Type	Electrical Responsivity	Conditions	NEP [W/Hz ^{1/2}]	Citation
bismuth, room temp	1 V/W	70-Ω device; bias = 1 V; video mod of 300 Hz @ 185 GHz	3 E-9	[43]
Hg _{0.8} Cd _{0.2} Te HEB	0.30 V/W at 17 mV bias, 36 GHz; 95 mV/W for 0.89 THz, 12 mV bias	room temperature,	2.4E-9 for 17 mV bias, 36 GHz; 7.4E-9 for 0.89 THz, 12 mV bias	[44]
Nichrome/tellurium composite microbolometer	120 V/W	Ambient, 30kHz	6.7 E-9	[45]
Si _x Ge _y :H	170 V/W	0.934 THz, uncooled	0.2 E-9	[46]
HgCdTe HEB		37 GHz, uncooled	4.00E-10	[47]
Vanadium oxide		uncooled	320E-12 at 4.3 THz, 9E-13 @ 7.5-14 μm	[48]
Nb		uncooled	4.00E-10	[49]
Hg _{0.8} Cd _{0.2} Te HEB	95 mV/W	0.89 THz, 200 Hz mod, 300 K	7.40E-09	[50]
niobium film	21 V/W	3.6 mA bias, 1 kHz mod, uncooled	1.10E-10	[32]
Ti; antenna-coupled microbolometer		10kHz chop, 1.04 mA bias, room temp	1.50E-11	[51]
Nb	8 V/W	0.25 V bias, room temp, 3.6 mA, 1 KHz	4.5E-10	[52]
Nb5N6	400 V/W	0.4 mA bias, > 10 KHz; electrical NEP	9.8E-12	[53]
vanadium oxide array	1.5e4 V/W	1 V bias, 130 μm, uncooled	2.00E-10	[54]
Nb, polyimide; antenna coupled	450 V/W	<1THz, no cryogenics mentioned	1.50E-11	[55]
Nb5N6	100.5 V/W	2 mA bias, 300 K, 1 kHz modulation	3.98E-10	[56]

Al/Nb; antenna coupled	85 V/W	1 kHz mod; 1.6 mA bias,	2.50E-11	[57]
Free-standing Nb bridge; antenna coupled	163 V/W (average over 5 devices)		17 E-12 (average over 5 devices); 650 GHz	[58,59]
Free-standing NbN bridge; antenna coupled	210 V/W (average over 10 devices)		12.5 E-12 (average over 10 devices); 650 GHz	[58,59]
Nb		uncooled, 3 kHz modulation	8.30E-11	[60]

5.7.2.2. Pyroelectrics.

Device Type	Electrical Responsivity	Conditions	NEP [W/Hz ^{1/2}]	Citation
Philips P5219 deuterated L-alanine TGS pyroelectric	321 V/W	10 Hz modulation; amplifier with gain of 4.8, 91 GHz	3.1E-08	[61]
QMC Instruments	18300 V/W	10 Hz modulation, room temp	4.40E-10	[62]
	1200 V/W	1.89 THz, <20 Hz modulation		[63]
LiTaO3 w/ chrome		Room temp, 530 GHz, Molelectron Model SPH-45	2.0e-9	[64]

5.7.2.3. Golay Cells

Table IV. Performance Characteristics of Room Temperature Golay Cells				
Device Type	Electrical Responsivity	Conditions	NEP [W/Hz ^{1/2}]	Citation
Tydex GOLAY CELL GC-1X	100000 V/W	21 Hz Chopper	1.4E-10	[65]
Microtech Instruments Golay cell	10 ⁴ V/W	12.5 Hz Chopper	10E-8	[66]
micro-array, LbL polymer membranes over Si		30 Hz mod 105 GHz	300E-9	[67]
Tydex Golay cell, 6-mm-diam Diamond Window		10 Hz modulation	7.0E-10	[63]

5.7.3. CMOS-Based and Plasma-Wave Detectors

Table V. Performance Characteristics of Plasma waves and CMOS based detectors				
Device Type	Electrical Responsivity	Conditions	NEP [W/Hz ^{1/2}]	Citation
BiCMOS SiGe 0.25μm, HBT	Current RI 1 A/W at 0.7 THz	Array 3x5, chopper 125 KHz	50 E-12 at 0.7 THz	[68]

BiCMOS SiGe 0.13 μ m, HBT	Voltage Rv 11 kV/W at 0.17 THz	chopper 0.16 KHz	10 E-12 at 0.17 THz	[69]
BiCMOS SiGe 0.25 μ m, NMOS	Voltage Rv 80 kV/W at 0.6 THz	3x5 array, chopper 16 KHz	300 E-12 at 0.6 THz	[70]
CMOS, 65 nm, NMOS	Voltage Rv 140 kV/W at 0.87 THz	32x32, chopper 5 KHz	100 E-12 at 0.87 THz	[71]
CMOS, 65 nm, NMOS	Voltage Rv 0.8 kV/W at 1 THz	3x5, chopper 1 KHz	66 E-12 at 1 THz	[72]
CMOS, 65 nm, SOI	Voltage Rv 1 kV/W at 0.65 THz	3x5, chopper 1 KHz	54 E-12 at 0.65 THz	[73]
130-nm CMOS-SBD	Voltage Rv 0.323 kV/W at 0.28 THz	4x4, chopper 1000 KHz	29 E-12 at 0.28 THz	[74]
65-nm CMOS-SBD		1 element; 1 MHz mod	42E-12 at 0.86 THz	[75]
CMOS, 150 nm, NMOS	Voltage Rv 11 V/W at 4.1 THz	1 element	1330 E-12 at 4.1 THz	[76]
InGaAs HEMT	Rv = 23 KV/W at 200 GHz	1 element	0.5E-12 at 200 GHz	[76,78]
Asymmetric Dual-Grating Gate InGaAs HEMT	Rv = 2.2 KV/W at 1.0 THz	1 element	15E-12 at 1.0 THz	[77,79]
Asymmetric Dual-Grating Gate InGaAs HEMT	Rc = 6.4 KV/W at 1.5 THz	1 element	50eE12 at 1.5 THz	[78,79]

References

- [1] H.C. Torrey and C.A. Whitmer, *Crystal Rectifiers*. New York: McGraw-Hill, 1948, Chap. 11.
- [2] E.R. Brown, A.C. Young, J. Zimmerman, H. Kazemi, and A.C. Gossard, "Advances in Schottky Rectifier Performance", *IEEE Microwave Magazine* **8**, p. 54 (2007)
- [3] Yun-Shik Lee, *Principles of Terahertz Science and Technology*. New York: Springer, 2009, p. 151-154.
- [4] Paul W. Kruse, "Principles of Uncooled Infrared Focal Plane Arrays" in *Uncooled Infrared Imaging Arrays and Systems*, vol. 3, Paul W. Kruse and David D. Skatrud, eds. San Diego: Academic Press, 1997, p. 25-29.
- [5] Marcel J. E. Golay, "A Pneumatic Infra-Red Detector," *Rev. of Sci. Inst.* **18**, p. 357 (1947).
- [6] http://www.tydexoptics.com/en/products/thz_optics/golay_cell/
- [7] M. Dyakonov and M. Shur, "Detection, mixing, and frequency multiplication of terahertz radiation by two-dimensional electronic fluid," *IEEE Trans. Electron Devices* **43**, p. 380 (1996).
- [8] A. Lisauskas, U. Pfeiffer, E. Ojefors, P.H. Bolivar, D. Glaab and H. Roskos, "Rational design of high responsivity detectors of terahertz radiation based on distributed self-mixing in silicon field effect transistors", *Journal of Applied Physics* **105**, 114511 (2009)
- [9] A. Gutin, V. Kachorovskii, A. Muraviev and M. Schur, "Plasmonic terahertz detector response at high intensities", *Journal of Applied Physics* **112**, 014508 (2012)

- [10] W. Knap, S. Rumyantsev, M.S. Vitiello, D. Coquillat, S. Blin, N. Dyakonova, M. Schur, F. Teppe, A. Trediuci and T. Nagatsuma. "Nanometer size field effect transistors for terahertz detectors," *Nanotechnology* **24**, 214002 (2013).
- [11] E. Ojefors, U.R. Pfeiffer, A. Lisauskas and H.G. Roskos; "A 0.65 THz Focal Plane Array in a Quarter-Micron CMOS process technology," *IEEE Journal of Solid State Circuits* **44**, (2009).
- [12] S. Maas, *The RF and Microwave Circuit Design Cookbook*. Norwood, MA: Artech House, 1998.
- [13] H. Nyquist, "Thermal Agitation of Electrical Charge in Conductors," *Phys. Rev.* **32**, p. 110 (1928).
- [14] R.H. Kingston, *Detection of Optical and Infrared Radiation*. Springer: New York, 1978, Sec. 2.2.
- [15] N.W. Ashcroft and N.D. Mermin, *Solid State Physics*. New York: Holt, Rinehart, and Winston, 1976, pp. 25.
- [16] W. Schottky, "Über spontane Stromschwankungen in verschiedenen Elektrizitätsleitern." *Annalen der Physik* (in German) **57**, p. 541 (1918).
- [17] A van der Ziel, *Noise in Solid State Devices and Circuits*. New York: John Wiley, 1986, Chap. 3.
- [18] C. Kittel, *Introduction to Solid State Physics* 7th Ed. New York: John Wiley & Sons, 1996, Chap. 5, and Chap. 6 Table 3.
- [19] yet another example in the vast physics literature of "the blind leading the blind."
- [20] W.S. Boyle and K.F. Rodgers, Jr. "Performance Characteristics of a New Low-Temperature Bolometer," *J. Opt. Soc. America* **49**, p. 66 (1959).
- [21] W. B. Davenport, W. L. Root, *An Introduction to the Theory of Random Signals and Noise*, New York: McGraw-Hill, 1958, Sec. 12-2.
- [22] J.D. Kraus, *Radio Astronomy* 2nd Ed., Powell, OH: Cygnus-Quasar, 1986, Sec. 7-1.
- [23] K. Chang, R. A. York, P. S. Hall, and T. Itoh, "Active Integrated Antennas", *IEEE Trans. Microwave Theory and Tech* **50**, p. 937 (2002).
- [24] D. B. Rutledge, D. P. Neikirk, and D. Kasilingam, "Integrated-Circuit Antennas," in *Infrared and Millimeter Waves, vol. 11: Millimeter Components and Techniques, Part II*. Editor: K.J. Button: New York: Academic Press, 1984.
- [25] P.H. Bolivar, M. Brucherseifer, J. G. Rivas, R. Gonzalo, I. Ederra, A. L. Reynolds, M. Holker, and P. de Maagt, "Measurement of the Dielectric Constant and Loss Tangent of High Dielectric-Constant Materials at Terahertz Frequencies," *IEEE Trans. Microwave Theory and Tech.*, **51**, p. 1062 (April 2003).
- [26] D. Segovia-Vargas, J. Montero-de-Paz, J. Crooks, P. G. Huggard, L.E. García, V. González-Posadas, B. Alderman and R. Turchetta, "300 GHz CMOS Video Detection using Broadband and Active Planar Antennas", European Conference on Antennas and Propagation 2012, EUCAP 2012.
- [27] C.A. Balanis, *Antenna Theory* 3rd Ed, Hoboken, NJ: John Wiley, 2005, Sec. 14.2.1.
- [28] E.R. Brown, A.C. Young, J.E. Bjarnason, H. Kazemi, J. Zimmerman, and A.C. Gossard, "Millimeter and Submillimeter Wave Performance of an ErAs:InAlGaAs Schottky Diode Coupled to a Single-Turn Square Spiral," *Int. J. High Speed Electronics and System* **17**, p. 383 (2007).
- [29] D. P. Neikirk and D. B. Rutledge, "Air-Bridge Microbolometer for Far-Infrared Detection," *Appl. Phys. Lett.*, **44**, p. 153 (1984).

- [30] D. P. Neikirk, D. B. Rutledge, and W. Lam, "Far-Infrared Microbolometer Detectors," *Int. J. Infrared Millimeter Waves*, **5**, p. 245 (1984).
- [31] N. Stojanovic, D. H. S. Maithripala, J. M. Berg, and M. Holtz, "Thermal conductivity in metallic nanostructures at high temperature: Electrons, phonons, and the Wiedemann-Franz law," *Phys Rev. B* **82**, 075418 (2010).
- [32] M. E. MacDonald and E. N. Grossman, "Niobium microbolometers for far-infrared detection" *IEEE Trans. Microwave Theory and Tech.* **43**, p. 893 (1995).
- [33] P. Horowitz and W. Hill, *The Art of Electronics* 2nd Ed, (Cambridge Press, Cambridge, UK, 1989), Fig. 7.42.
- [34] AD797: Analog Devices Corporation, Norwood, MA 02062, http://www.analog.com/static/imported-files/data_sheets/AD797.pdf
- [35] J.L. Hesler, and T.W. Crowe, "NEP and responsivity of THz zero-bias Schottky diode detectors" *Int. Conference on Infrared and Millimeter Waves, (IRMMW-THz)*. **2007**, p. 844 (2007).
- [36] A. Semenov, O. Cojocari, H.-W. Hubers, F. Song, A. Klushin, and A.S. Muller, "Application of Zero-Bias Quasi-Optical Schottky-Diode Detectors for Monitoring Short-Pulse and Weak Terahertz Radiation" *IEEE Electron Device Letters* **31**, p. 674 (2010).
- [37] V. V. Kubarev, G. M. Kazakevitch, Y. Uk Jeong, B.C. Lee "Quasi-optical highly sensitive Schottky-barrier detector for a wide-band FIR FEL," *Nuclear Instruments and Methods in Physics Research Section A: Accelerators, Spectrometers, Detectors and Associated Equipment* **507**, 523 (2003).
- [38] Y-P. Xu, R-S. Huang, and G. A. Rigby, "A silicon-diode-based infrared thermal detector array," *Sensors and Actuators A*, **37-38**, p.226 (1993).
- [39] <http://vadiodes.com/index.php/en/products/detectors>
- [40] A. Bertolini, G. Carelli, C. A. Massa, A. Moretti, and F. Strumia "Detection and Mixing Properties of an InSb Metal-Semiconductor Point Contact Diode" *Int. J. Infrared and Millimeter Waves*, **20**, p. 1121 (1999).
- [41] R. U. Titz, H. P. Roser, G. W. Schwaab, H. J. Neilson, P. A. Wood, T. W. Crowe, W.C.B. Peatman, J. Prince, B. S. Deaver, H. Alius, and G. Dodel "Investigation of GaAs Schottky Barrier Diodes in the THz Range," *Int. J. of Infrared and Millimeter Waves* **11**, p.809 (1990).
- [42] M. R. Abdel-Rahman, B. Monacelli, A. R. Weeks, G. Zummo, and G. D. Boreman "Design, fabrication, and characterization of antenna-coupled metal-oxide-metal diodes for dual-band detection" *Opt. Eng.* **44**, 066401 (2005).
- [43] C.C. Ling and G.M. Rebeiz, "A wideband monolithic submillimeter-wave quasi-optical power meter," *IEEE Microwave Symposium Digest* **1990.**, p. 1315 (1990).
- [44] V.N. Dobrovolsky, F.F. Sizov, Yu. E. Kamenev, G. Sichevska, and A.B. Smirnov, "Millimeter and Submillimeter Semiconductor Hot Electron Bolometer" *Sixth International Kharkov Symposium on Physics and Engineering of Microwaves, Millimeter and Submillimeter Waves* 2007 **1**, p. 198 (2007).
- [45] S.M. Wentworth and D.P. Neikirk, "Composite microbolometers with tellurium detector elements" *IEEE Transactions on Microwave Theory and Tech.* **40**, 196 (1992).
- [46] A. Kosarev, S. Rumyantsev, M. Moreno, A. Torres, S. Boubanga, and W. Knap "SixGey:H-based micro-bolometers studied in the terahertz frequency range," *Solid-State Electronics* **54**, 417 (2010).

- [47] V. Dobrovolsky, F. Sizov, V. Zabudsky, N. Momot "Mm/sub-mm Bolometer Based on Electron Heating in Narrow-gap Semiconductor," *Terahertz Science and Technology* **3**, p. 33 (2010).
- [48] Real-time, continuous-wave terahertz imaging using a microbolometer focal-plane array, United States Patent 7692147.
- [49] Q. Hu, G. de Lange, S. Verghese, E. Duerr, I. Lyubomirsky, A. Rahman, F. Rana, R. A. Wyss, B. Xu, N. D. Zamdmer "High-Frequency (>100 GHz) and High-Speed (<10 ps) Electronic Devices" *Massachusetts Institute of Technology. Research Laboratory of Electronics. Progress Report* **139**, (1996).
- [50] V. Dobrovolsky and F. Sizov "A room temperature, or moderately cooled, fast THz semiconductor hot electron bolometer" *Semicond. Sci. Technol.* **22**, p. 103 (2007).
- [51] A. Luukanen "High Performance Microbolometers and Microcalorimeters: from 300 K TO 100 mK," Ph.D. Dissertation, University of Jyvaskyla, Finland (2003).
- [52] A. Rahman, G. de Lange, and Q. Hu "Micromachined room-temperature microbolometers for millimeter-wave detection" *Applied Physics Letters* **68**, p. 2020 (1996)
- [53] L. Kang, X. H. Lu, J. Chen, B. B. Jin, P. H. Wu, Q. J. Yao and S. C. Shi "Room temperature Nb₅N₆ bolometer for THz detection" *34th Int. Conf. on Infrared, Millimeter, and Terahertz Waves*, p. 1-2 (2009).
- [54] M. A. Dem'yanenko, D. G. Esaev, V. N. Ovsyuk, B. I. Fomin, A. L. Aseev, B. A. Knyazev, G. N. Kulipanov, and N. A. Vinokurov, "Microbolometer detector arrays for the infrared and terahertz ranges," *J. Opt. Technol.* **76**, p. 739 (2009).
- [55] A. Luukanen, V-P. Viitanen, "Terahertz imaging system based on antenna-coupled microbolometers" *Proc. SPIE* **3378**, p. 34 (1998);
- [56] X.H.Lu, L.Kang, J.Chen, Y.Y.Zhong, N.He, L.B.Zhang, B.B.Jin, W.W.Xu, P.H.Wu, Q.J.Yao, S.C.Shi "A terahertz detector operating at room temperature" *Proc. SPIE* **7277**, 72770N (2008);
- [57] A. J. Miller, A. Luukanen, E. N. Grossman "Micromachined antenna-coupled uncooled microbolometers for terahertz imaging arrays" *Proc. SPIE* **5411** (2004).
- [58] C. Dietlein, A. Luukanen, J.S. Penttila, H. Sipola, L. Gronberg, H. Seppa, P. Helisto, and E.N. Grossman, "Performance Comparison of Nb and NbN Antenna-Coupled Microbolometers," *Proc. SPIE* **6549** (2007).
- [59] Dr. Erich Grossman, private communication.
- [60] A. Rahman, E. Duerr, G. de Lange, and Q. Hu "Micromachined room-temperature microbolometer for mm-wave detection and focal-plane imaging arrays" *Proc. SPIE* **3064**, p. 122 (1997).
- [61] M. R. Webb "A Millimeter Wave Pyroelectric Detector," *Int. J. of Infrared and Millimeter Waves* **12**, p. 1225 (1991).
- [62] http://www.qmcinstruments.co.uk/index.php?option=com_content&view=article&id=236&Itemid=530
- [63] J. Yang, X-J. Gong, Y-D. Zhang, L. Jin "Research of an infrared pyroelectric sensor based THz detector and its application in CW THz imaging," *Proc. SPIE* **7385**, 738521 (2009).
- [64] E.R. Brown and D. Dooley, unpublished.
- [65] Tydex GOLAY CELL GC-1X, http://tydexoptics.com/pdf/Golay_cell.pdf
- [66] Microtech Instruments, <http://www.mtinstruments.com/downloads/Golay%20Cell%20Datasheet%20Revised.pdf>

- [67] D.R. Denison, M. E. Knotts, M. E. McConney, V. V. Tsukruk "Experimental characterization of mm-wave detection by a micro-array of Golay cells," *Proc. SPIE* **7309**, 7309J (2009).
- [68] R. Al Hadi, J. Grzyb, B. Heinemann and U.R. Pfeiffer; "A Terahertz Detector Array in SiGe HBT Technology," *IEEE J. of Solid State Circuits* **48**, pp. 2002-2010 (2013).
- [69] E. Dacquay, A. Tomkins, K. Yau, E. Laskin, P. Chevalier, A. Chantre, B. Sautreuil, and S. Voinigescu, "D-band total power radiometer performance optimization in an SiGe HBT technology," *IEEE Trans. Microwave Theory Tech.* **60**, p. 813 (2012).
- [70] U. Pfeiffer and E. Ojefors, "A 600-GHz CMOS focal-plane array for terahertz imaging applications," in *Proc. Eur. Solid-State Circuits Conf.*, p. 110 (2008).
- [71] R. Al Hadi, H. Sherry, J. Grzyb, Y. Zhao, W. Forster, H. Keller, A. Cathelin, A. Kaiser, and U. Pfeiffer, "A 1 k-pixel video camera for 0.7–1.1 terahertz imaging applications in 65-nm CMOS," *IEEE J. Solid-State Circuits* **47**, p. 2999 (2012).
- [72] R. Al Hadi, H. Sherry, J. Grzyb, N. Baktash, Y. Zhao, E. Öjefors, A. Kaiser, A. Cathelin, and U. Pfeiffer, "A broadband 0.6 to 1 THz CMOS imaging detector with an integrated lens," in *IEEE MTT-S Int. Microwave Symp. Dig.*, p. 1 (2011).
- [73] E. Öjefors, N. Baktash, Y. Zhao, R. Al Hadi, H. Sherry, and U. Pfeiffer, "Terahertz imaging detectors in a 65-nm CMOS SOI technology," in *Proc. Eur. Solid-State Circuits Conf.*, p. 486 (2010).
- [74] R. Han, Y. Zhang, Y. Kim, D. Y. Kim, H. Shichijo, E. Afshari, and K. K. O, "280 GHz and 860 GHz image sensors using Schottky-barrier diodes in 0.13 m digital CMOS," in *IEEE Int. Solid-State Circuits Dig.*, p. 254 (2012).
- [75] S. Boppel, A. Lisauskas, D. Seliuta, L. Minkevicius, L. Kasalynas, G. Valusis, V. Krozer, and H. Roskos, "CMOS-integrated antenna-coupled field-effect-transistors for the detection of 0.2 to 4.3 THz," in *IEEE SiRF Tech. Dig.*, p. 77 (2012).
- [76] Y. Kurita, G. Ducournau, D. Coquillat, A. Satou, K. Kobayashi, S.A. Boubanga-Tombet, Y.M. Meziani, V.V. Popov, W. Knap, T. Suemitsu, and T. Otsuji, Ultrahigh sensitive sub-terahertz detection by InP-based asymmetric dual-grating-gate high-electron-mobility transistors and their broadband characteristics, *Appl. Phys. Lett.* **25**, 251114 (2014)
- [77] T. Watanabe, S. Boubanga Tombet, Y. Tanimoto, D. Fateev, V. Popov, D. Coquillat, W. Knap, Y. Meziani, Y. Wang, H. Minamide, H. Ito, and T. Otsuji, "InP- and GaAs-based plasmonic high-electron-mobility transistors for room-temperature ultrahigh-sensitive terahertz sensing and imaging," *IEEE Sensors J.* **13**, p. 89 (2013).
- [78] T. Otsuji, S.A Watanabe, B. Tombet, A. Satou, V. Ryzhii, V. Popov, and W. Knap, "Emission and detection of terahertz radiation using two dimensional plasmons in semiconductor nano-heterostructures for nondestructive evaluations," *Opt. Eng.* **53**, 031205 (2013) .
- [79] Dr. Michael Shur, private communication.