ADVANCED TERAHERTZ DETECTORS AND FOCAL-PLANE ARRAYS BASED ON

SB-HETEROSTRUCTURE BACKWARD DIODES

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by

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Abstract

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In this dissertation, the design, fabrication and characterization of advanced terahertz (THz) quasi-optical detectors and focal-plane arrays (FPAs) based on monolithically integrated heterostructure backward diodes (HBDs) are presented for THz sensing and imaging applications. In order to develop highly sensitive room temperature THz detectors and FPAs, zero bias Sb-HBDs are attractive to be employed owing to their relatively high curvature coefficient (γ), high sensitivity, low noise equivalent power (NEP), high cut-off frequency, and room temperature operation. To develop high performance detectors and FPAs, HBDs with submicron-scale device areas are preferred for their high detector bandwidth. Since submicron scale HBDs have high device impedances at terahertz frequencies, lens-coupled planar folded dipole antennas (FDAs) which have a wide range impedance tuning capacity at THz frequencies are adopted in the detector design to achieve conjugate impedance matching for maximum detector sensitivities without additional matching networks. For a prototype demonstration, HBDs
with 0.16 $\mu m^2$ and 0.1 $\mu m^2$ active areas have been employed for the detector design at 200 and 585 GHz respectively. Simulation results show that maximum detector sensitivities of 21,000 V/W at 200 GHz and 9,500 V/W at 585 GHz could be achieved. The corresponding minimum NEPs ($\text{NEP}_{\text{min}}$) of these detectors are estimated to be 0.42 pW/\text{VHz} and 1.3 pW/\text{VHz} respectively.

In this work, HBDs are integrated with FDAs using submicron-scale airbridges and anode-to-mesa spacing to minimize parasitic capacitance and frequency dependent spreading resistance respectively. On the basis of modeling results, a novel, scalable, and robust fabrication process has been developed using mix-and-match electron beam and optical lithography. A record high zero bias curvature co-efficient of -58V$^{-1}$ has been obtained for HBDs developed using this process. The measurements of the lens-coupled HBD detector with a 0.7× 0.7 $\mu m^2$ device area show that a peak detector sensitivity of approximately 2400 V/W and a $\text{NEP}_{\text{min}}$ of 2.14 pW/\text{VHz} have been obtained at 170 GHz without applying anti-reflection coating on the silicon lens. If an antireflection coating was used, a sensitivity of approximately 3500 V/W and a $\text{NEP}_{\text{min}}$ of 1.48 pW/\text{VHz} are projected. The radiation patterns of the quasi-optical detector in both the E- and H-planes have been measured and good agreement has been achieved between simulation and measurement.

Finally, for imaging applications, the single element design has been expanded into full 2D THz FPAs and the off-axis radiation patterns, angular resolution, and mutual coupling have been studied. The reported approach using monolithically integrated heterostructure backward tunneling diodes is promising for developing high performance
and compact detectors and FPAs for millimeter-wave and THz sensing and imaging applications.
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1.1 Terahertz technology and its potential applications

Conventionally, electromagnetic signals with frequencies between 30 GHz and 300 GHz are called millimeter waves, and the submillimeter wave range is defined from 300 GHz to 3 THz. The term “terahertz” broadly refers to a frequency range from 100 GHz to 10 THz. The wavelengths of radiation in the terahertz band correspondingly range from 3 mm to 0.03 mm (or 30 μm). The development of electronics in this frequency range is challenging because this frequency band lies between the microwave and infrared regime. This is too high for conventional electronics and the photon energy is too small for classical optics. In addition, the atmospheric propagation path is also limited for this frequency range [1].

Since the latter half of the twentieth century, radio astronomers have used terahertz technology for observing millimeter- and submillimeter-wave frequencies emitted by astronomical objects. By observing the universe in this frequency region, astronomers have been able to examine basic clues as to the formation and evolution of star, galaxies and our universe [2]. Fig. 1.1 shows the cosmic microwave background (CMB) radiation (the remnants of Big Bang radiation) of the universe observed by COBE.
(Cosmic Background Explorer) at millimeter-wave frequencies (Noble Prize 2006),
together with 1.89 THz spectral lines emitted by interstellar gas [3,4]. The CMB data was
used for interpreting the cosmic evolution of the universe from the Big Bang to its current
form, explaining why THz frequency spectrum is so important to radio astronomers [3].
Over the last several decades, scientists and engineers have developed several space and
ground based radio astronomy observatories [5]— including NASA’s Submillimeter Wave
Astronomy Satellite (SWAS) [6], ESA’s Planck and Herschel Space Observatory [7,8],
airborne observatory SOFIA [9,10] and Atacama Large Millimeter Array (ALMA) [11] — in
order to study the universe in this frequency range.

Figure 1.1. The cosmic microwave background of universe observed by COBE at
millimeter-wave frequencies together with 1.89 THz spectral lines (black lines) emitted by
interstellar gas [4].

THz technology has also been driven by applications in earth and planetary science
[12-22]. This frequency range is rich in spectral signatures of atmospheric gases which are
involved in ozone destruction, global warming, air pollution and precipitation [15-17].
Remote analysis of atmospheric gases at terahertz frequencies is also a powerful tool to study the atmosphere of other planets and to assess their potential for hosting life [18,19]. Several satellites have been deployed to study the atmosphere of earth and other planets at this frequency range including the NASA’s Microwave Limb Emission Sounder [20], Swedish Odin satellite [21], and Japanese Experiment Module [22].

Besides radio astronomy and space science, terahertz technology has also opened up numerous commercial applications recently due to the emergence of advanced THz solid-state devices, circuits and systems [23-37]. There has been an increased interest in THz technology for detecting concealed weapons, as THz radiation can readily be transmitted through most non-metallic mediums such as packaging, clothing, shoes, and bags, and it poses no health risk to a person being scanned by THz systems [38]. In addition, explosives, chemical agents, and illegal drugs have spectroscopic signatures in this frequency range and can be detected using THz systems [39]. Moreover, THz systems offer higher spatial resolution than the currently used imaging systems for security applications (e.g. scanning systems in airports) which is beneficial to identify hidden objects accurately.

The non-ionizing nature of THz radiation lends THz techniques to biological, medical, and pharmaceutical applications [40-44]. Biological tissues have exceptionally high absorption losses at THz frequencies due to the presence of water contents [41]; this high absorption coefficient creates an extreme contrast between tissues with lesser or higher degree of water contents. This can help to make a distinction between cancerous
and healthy tissues for planning surgery without biopsy [42]. In addition, it was demonstrated that the complex refractive index of DNA molecules is strongly dependent on their hybridization state in the THz frequency range that enables label-free characterization of unknown DNA sequences [43]. Moreover, this technology can be applied to monitor the quality of pharmaceutical products nondestructively (i.e. crystalline structure, thickness and chemical composition) [44].

Figure 1.2. Applications of terahertz band communication systems for cellular networks (left) and local area networks (right).

THz technology has been given significant attention for use in high bandwidth wireless communication systems [45,46]. THz communication systems with data rates up to terabit per second (Tbps) are advantageous for a wide variety of applications including high bandwidth cellular networks, wireless local area networks (WLAN), wireless personal area networks (WPAN), and the in-home distribution of high-definition television (HDTV). For example, high speed interconnections between fiber optical links and personal devices (such as laptops and tablet-like devices) will facilitate the use of bandwidth-
intensive applications in indoor scenarios. Fig. 1.2 shows short-range terahertz band communication systems for cellular and local area networks.

The applications of THz technology are not limited to the above areas and are growing with the advancement THz electronics [47]. In order to develop high performance THz sensing and imaging systems for these applications, highly sensitive detectors and focal plane arrays are desirable. Recently, scientists and engineers intensified their efforts into the development of detectors and focal plane arrays (FPAs) for THz sensing and imaging applications.

1.2 Overview of THz sensing and imaging detectors and FPAs

Two major classes of systems are currently available for THz sensing and imaging applications: active and passive detection and imaging systems [48-53]. For an active detection and imaging system, the imaging object is illuminated by an external THz source to enhance the signal level and the reflected signals from the imaging object are detected by THz detectors. The difference in reflectivity from different objects increases the contrast of a THz image. However, an active system has several significant drawbacks. For example, such a system requires additional electronics (e.g. THz sources, low noise amplifiers, and related optics to illuminate the imaging object) which significantly increases the system’s complexity, cost, and power consumption. In addition, active THz systems can produce glint (bright regions in images from specular reflection), making it difficult to obtain natural looking high-fidelity images. In comparison, passive THz detection and imaging systems do not require external illumination but instead rely on
natural blackbody radiation from objects in the THz region. These systems do not suffer from glint due to the absence of illumination, providing natural looking THz images. As a result, highly sensitive, extremely low noise detectors are required to realize room temperature THz passive detection and imaging systems.

The most widely used technique in the THz systems is based on single point detection configuration. However, in many applications, only one pixel of object information from a receiver is not enough. To effectively map the distribution of intensity of an imaging object, many pixels of information are needed. Although mechanical raster scanning can be applied to a single element detector to form two-dimensional THz image [54], in some cases, it is not applicable due to long scanning time. For example, some radio astronomy events happen on the scale of milliseconds [55] and it is very difficult to catch such events by a single element detector with mechanical scanning. In addition, mechanical scanning is not advantageous for surveillance and security screening since that system generally requires image acquisition time approaching video rates with sufficient imaging resolution. As a direct alternative, imaging FPAs are demanded for these applications where a large number of pixels are used in parallel to collect incoming THz radiation from an imaging object. This approach is essential to reduce observing and processing time [56,57], leading to compact and portable imaging systems with reduced overall complexity and manufacturing cost.
1.3 Goal of this research

The goal of this research is to design, fabricate and characterize THz detectors and FPAs at WR-5.1 (140-220 GHz) and WR-1.5 (500-750 GHz) frequency bands based on Sb-based heterostructure backward diodes (Sb-HBDs) for THz sensing and imaging applications. These two frequency bands were chosen for prototype demonstration for two reasons: 1) they are of great interest to the science/engineering community; and 2) they required instrumentation (i.e. local oscillators, sources, VNA) that are readily available in our THz circuits and systems laboratory. Consequently, the detectors and FPAs developed in this work should be viewed as scaled-model prototypes for sensing and imaging at terahertz frequencies.

1.4 Dissertation organization

In Chapter 2, different types of THz detection technologies are presented on the basis of their figures of merit. To assess the performance of HBDs at THz frequencies, the dynamic range, sensitivity, noise equivalent power (NEP), cutoff frequencies and device impedances are studied based on their lumped element circuit model. In order to develop THz detectors and FPAs with optimized performance, lens-coupled folded dipole antennas (LC-FDAs) are proposed in Chapter 3. The embedding impedance, far-field radiation patterns, directivity, and Gaussian coupling efficiency are studied to evaluate the performance of LC-FDAs. Chapter 4 focuses on the design, fabrication, and characterization of fully integrated HBD detectors. On the basis of the design results, a unique fabrication process is developed for integrating HBDs with FDAs using submicron-
scale airbridges. To demonstrate the effectiveness of reported integration and fabrication process, DC, RF, and quasi-optical characterization results are presented. In Chapter 5, the design, fabrication, and characterization results of two-dimensional FPAs are reported. The conclusion and future work that could be undertaken are discussed in Chapter 6.
CHAPTER 2:
HIGH PERFORMANCE HETEROSTRUCTURE BACKWARD DIODES FOR TERAHERTZ SENSING
AND IMAGING

In this chapter, comparative assessments of different types of detector and FPA technologies are discussed on the basis of their important figures of merit for terahertz sensing and imaging applications. In order to develop highly sensitive room temperature detectors and imaging FPAs, submicron-scale Sb-HBDs are attractive to be employed owing to their higher curvature coefficient ($\gamma$), higher sensitivity, lower noise equivalent power ($NEP$), higher cut-off frequency, and room temperature operation [33,58-60]. The performance of submicron-scale HBDs (i.e. dynamic range, device impedance, sensitivity, $NEP$, cutoff frequency) is assessed based on their lumped element circuit model using Advanced Design System (ADS) software. The results suggest that the performance of HBDs is superior over the competing technologies for THz sensing and imaging applications.

2.1 THz heterodyne and direct detection schemes

Two types of detection schemes are widely used in the THz regime, “heterodyne” or coherent detection and “direct” or incoherent detection [61]. At THz frequencies, heterodyne detectors typically exhibit higher sensitivity and greater dynamic range. Heterodyne receivers are frequency converters that translate high frequency or RF signals
into a lower frequency (intermediate frequency or “IF”) signals. The output IF signal preserves both the magnitude and phase information of the original signal. Fig. 2.1 shows the schematic of a THz heterodyne detection system where the incoming THz radiation is mixed with the local oscillators (LOs) and the IF signals are generated using non-linear mixing elements (i.e. diode, transistor). The IF signals can further be amplified using multistage low noise amplifiers (LNAs) if the detected signals are extremely weak. As seen in Fig. 2.1, heterodyne detectors require LOs that are difficult to realize in THz frequency range. In addition, the use of additional circuit elements (i.e., LNAs, matching network, etc.) significantly increases system complexity, cost, and power consumption. As a consequence, it is very challenging to realize large-scale FPAs based on heterodyne receiver systems for THz sensing and imaging applications.

Figure 2.1: Simplified detection scheme of heterodyne detector for THz imaging applications.
Figure 2.2: Simplified detection scheme of direct detector for THz sensing and imaging application.

For developing simple but high-performance THz detection circuits suitable for sensing and imaging applications, direct (or incoherent) detectors that translate incident RF signals into DC (i.e., current or voltage) output are attractive. As shown in Fig. 2.2, a direct detector essentially works by rectifying an applied AC (or RF) signal using a nonlinear device such as a diode. Different from the mixers discussed previously, the output of a direct detector is DC and all phase information of the incoming RF signal is lost. In addition, the output DC signal is proportional to the input RF power, and because of this, direct detectors are usually called square law detectors. In comparison to complicated mixer configurations, the direct detection scheme is much simpler and easier to realize, making it suitable for developing detectors and FPAs for THz sensing and imaging applications.

The two most important figures of merit for direct detectors are responsivity (current or voltage) and NEP. The responsivity of a direct detector is defined as the ratio of output detected signal (either voltage or current) to incident RF power. NEP quantifies
the noise power generated by a detector itself and in principle places a limit on the smallest power that can be detected. Direct detectors with high responsivity, low NEP and without biasing are very attractive to the THz community and are preferred for developing THz detectors and imaging arrays. Direct detectors that operate at room temperature (room-T) have attracted significantly more attention in recent years. It is well known that to achieve a high-resolution imaging array, a high circuit packing density is required, leaving very limited space for accommodating circuitry such as low-pass filters (LPFs) and LNAs [62]. LNAs are not necessary if high-responsivity and low-noise detectors are chosen, thus reducing the single pixel circuit size. For large-scale imaging arrays, biasing is always problematic and undesired. Zero-bias detectors are of particular interest since they can further reduce the footprint of the single element circuit, system complexity and power consumption. In addition, 1/f noise that results from external biasing can be avoided by employing zero-bias detectors.

2.2 Review of device technologies for direct detectors

Different types of devices were extensively studied and employed for THz sensing and imaging applications including superconducting hot-electron bolometers (HEBs), pyroelectric devices, Schottky barrier diodes (SBDs) and field effect transistors (FETs).

HEBs operate based on thermal principles, which sense the change in temperature associated with the incident RF energy absorbed. HEBs are widely used by radio astronomers for their high sensitivity and extremely low NEP (≈10^{-17} W/Hz^{1/2}) [63-65]. However, HEBs require liquid helium based cryogenic cooling systems to suppress the
thermal fluctuation noise arising from thermal resistance between the bolometer and its heat sink. As a result, the overall detector becomes physically large and very expensive. For commercial imagers and portable systems, the use of cryogenic cooling system is a significant limitation. In addition, the response time of HEBs is fairly long (1-15 ms) which significantly reduces the imaging speed of a THz system. Another THz detector that operates based on thermal principle is the pyroelectric sensor whose output current source is inversely proportional to the thickness of absorbing material and the rate of temperature change. Some inherent advantages of pyroelectric detectors are room temperature operation, small size, and low cost. However, their sensitivity is generally lower than 100 V/W, NEP is in the order of $10^{-9}$-$10^{-8}$ W/Hz$^{1/2}$, and response time is significantly longer. The most widely used semiconductor devices, operating at terahertz frequencies, are the Schottky barrier diodes that have long been used in heterodyne and direct detection systems [19,66,67]. State-of-the-art SBDs exhibit very high cut-off frequency (≈16 THz) and have the merits of high sensitivity, low noise performance, fast response, and room temperature operation. However, the main limitation of low barrier SBDs for direct detection is their theoretically-bounded DC curvature coefficient which cannot exceed $q/kT$ [68], resulting in limited detector sensitivity. Another limitation of SBDs is their use of bias circuitry to achieve strong non-linearity which introduces additional $1/f$ noise from the device, resulting in poor detector performance. Recently, field effect transistor-based THz detectors (e.g. CMOS FET detectors) attracted significant attention for developing THz FPAs for room temperature THz imaging systems [69].
Although these devices are highly sensitive, their performance is typically inferior (as compared to SBDs) with NEPs in the order of tens or hundreds of pW/Hz$^{1/2}$. Moreover, CMOS-based systems are extremely challenging to be scaled to frequencies above 1 THz.

Sb-based heterostructure backward diodes (Sb-HBDs) are attractive to be employed as direct detectors and are promising candidates for developing advanced THz sensing and imaging systems [33,70-72]. The main advantage of HBDs is that these devices can be engineered to greatly improve the curvature co-efficient at zero bias. Moreover, their zero-bias operation also bring the advantage of eliminating the complication caused by external bias circuits. In the following, a detail discussion of Sb-HBDs is presented.

2.3 High performance heterostructure backward diodes (HBDs)

Scientists and engineers have investigated the potential of Esaki tunnel diodes for microwave detection and mixing applications for many years [73-78]. Modified from Esaki diodes, HBDs with engineered band structures may pave the way for achieving high sensitivity and low noise detectors at THz frequencies [79-85]. Since the responsivity of HBDs is directly proportional to their curvature coefficient, the performance of zero-biased HBDs is expected to be superior than the thermionic-emission based devices. Zero bias HBDs have already been experimentally demonstrated with curvature coefficients higher than the theoretical limit of thermionic-emission based devices such as SBDs and planar-doped barrier (PDB) diodes. Furthermore, zero bias HBDs do not require bias circuitry, and the elimination of bias circuitry significantly reduces $1/f$ noise. These devices
are relatively insensitive to changes in ambient temperature and can operate in wide temperature range (cryogenic to 4000 K) [86]. In addition, these devices exhibit extremely low NEP, high cutoff frequency (~8 THz) and fast response (video rate), making them suitable for THz detection and imaging at room temperature [87-92].

2.3.1 Device physics and operation

Fig. 2.3(a) shows the typical device epitaxial layer structure consisting of a n+-InAs cathode layer, a GaSb anode layer and an AlSb (~1.1 nm) tunnel barrier between the InAs cathode and GaSb anode. Ohmic contact for the cathode and anode is formed on the top n+-InAs layer. This structure results in a broken-gap energy band alignment between cathode and anode, allowing the flow of highly nonlinear tunneling current near zero bias. In order to further improve the device nonlinearity, a p-type δ-doping plane in the n-InAs for adjusting the cathode band bending near the tunnel barrier, and an Al$_{0.12}$Ga$_{0.88}$Sb layer for precisely tuning the band alignment between the cathode and anode are added. As shown in Fig. 2.3(b), under forward bias the InAs conduction band edge becomes higher than the Al$_{0.12}$Ga$_{0.88}$Sb valence band, blocking electrons from tunneling from the cathode through the barrier to the anode. Under reverse bias, the valence band edge of Al$_{0.12}$Ga$_{0.88}$Sb is higher than the InAs conduction band, allowing electrons to tunnel freely from the anode to the cathode, resulting in rapidly increasing backward current with increasing reverse bias voltage.
Figure 2.3. (a) The device epitaxial layer structure and (b) band diagram of the InAs/AlSb/Al$_{0.12}$Ga$_{0.88}$Sb/GaSb backward diode. The high performance heterostructure backward diode consists of an AlSb (~1.1 nm) tunnel barrier between the InAs cathode and GaSb anode.
A typical measured $I$-$V$ characteristic for a high-performance Sb-HBD with submicron device area is shown in Fig. 2.4 [58]. The carefully engineered energy band alignment in the device heterostructure results in a strong nonlinear $I$-$V$ characteristic than that cannot be achieved in conventional Schottky diodes. Nonlinearity of HBD, which is defined as $\gamma = (d^2I/dV^2)/(dI/dV)$, exceeds the theoretical maximum of 38.5 V$^{-1}$ at room temperature for a thermionic device (e.g., Schottky diodes). For the heterostructure design shown in Fig. 2.3(a), HBDs are demonstrated with a high zero bias curvature coefficient of $-47$ V$^{-1}$, an unmatched sensitivity of 4500-5800 V/W, a projected impedance-matched sensitivity of $\sim 50$ kV/W at 100 GHz and a record low NEP$_{\text{min}}$ of 0.18 pW/Hz$^{1/2}$ [58].
Figure 2.5. Lumped element nonlinear equivalent circuit model of HBD together with GSG probe pad for the heterostructure design shown in Fig. 2.3(a).

Figure 2.6. ADS harmonics balance simulation setup for unmatched and matched detector sensitivity, NEP and device impedance calculation.

2.3.2 Lumped element circuit model of HBDs

In order to evaluate the performance of HBDs in the THz frequency range, an accurate scalable nonlinear equivalent circuit model was developed by Ze Zhang for the heterostructure design shown in Fig. 2.3(a) [59]. The equivalent circuit model of HBD is
shown in Fig. 2.5, where $R_j$ and $C_j$ are the nonlinear bias dependent junction resistance and junction capacitance respectively and $R_s$ is the series resistance associated with the contact resistance and frequency dependent spreading resistance caused by current crowding effects within the device. The $C_p$ and $L_p$ are assumed in the model to be bias independent capacitance and inductance of the ground-signal-ground (GSG) probe pad respectively due to their geometric structure.

The performance of HBD detectors have been assessed by estimating their dynamic range, sensitivity, $NEP$, cut-off frequency, and device impedance at THz frequencies. To evaluate the performance of HBDs, ADS harmonic balance simulations with order $n = 3$ have been performed for a frequency range from 10 GHz to 1 THz. The simplified diagram for harmonic balance simulations is shown in Fig. 2.6, where the HBD is represented by its lumped element circuit model. $P_{in}$ is THz power provided to intrinsic HBD and $Z_s$ is the source impedance.

2.3.3 Dynamic range

Since HBDs are square-law device, their rectified output voltage is directly proportional to the incident RF power. However, under high RF illumination, the device output deviates from the linear region, resulting in severe degradation of the detector’s performance. The range of input RF power for which HBDs work as direct detectors, known as dynamic range, was assessed for RF input power ranging from -50 dBm to 0 dBm at 200 GHz for a HBD device with 0.5 µm$^2$ active area for the heterostructure design shown in Fig 2.3(a). As show in Fig. 2.7, the simulation result shows that the diode
operates in the linear region for low input RF power (< -23 dBm) and the 1 dB compression point was found to be -20 dBm, as shown in Fig. 2.7.

Figure 2.7. Simulated detector output voltage as a function of input RF power for HBD device with 0.5 \( \mu \)m\(^2\) active area at 200 GHz. The 1 dB compression point was estimated to be -20 dBm.

2.3.4 Intrinsic sensitivity

Sensitivity is the most important figure of merit of a direct detector. Sensitivity is defined as the ratio of generated output DC signal (voltage or current) to the incident RF power. The maximum intrinsic device sensitivity of HBD under impedance matching condition was derived to be [60],
\[
\beta_{v,\text{max}} = \frac{\gamma R_j}{2(1 + \frac{R_s}{R_j})(1 + \frac{R_s}{R_j} + \omega^2 C_j^2 R_s R_j)}.
\] (2.1)

The curvature coefficient used in Eq. 2.1 is the intrinsic curvature coefficient of HBDs with higher junction resistance compared to their series resistance (i.e. \(R_j >> R_s\)). Eq. 2.1 shows that the maximum voltage sensitivity is directly proportional to the curvature coefficient \(\gamma\).

Figure 2.8. Simulated unmatched and matched sensitivities of submicron-scale HBDs for a frequency range from 10 GHz to 1 THz. The simulations were performed using ADS harmonic balance simulations.

In order to estimate the maximum sensitivity \(\beta_{v,\text{max}}\) of the HBD detector at THz frequencies, the scalable HBD model [59] for the heterostructure design shown in Fig 2.3
(a) was simulated using S-parameter simulations in ADS. An input RF power ($P_{in}$) of 0.1 pW was applied to the HBD detector for a frequency sweep from 10 GHz to 1 THz. A RF source impedance ($Z_s$) of 50 Ω was used in simulations. The unmatched sensitivity was calculated from the detector output voltage and incident RF input power using,

$$\beta_{v} = \frac{v_d}{P_{in}},$$  \hspace{1cm} (2.2)

where $v_d$ is the detector output dc voltage and $P_{in}$ is the incident power. The maximum detector sensitivity under impedance matching condition was determined by:

$$\beta_{v,\text{max}} = \frac{v_d}{P_{abs}},$$  \hspace{1cm} (2.3)

where $P_{abs}$ is the power absorbed by the diode. The simulations were performed for submicron-scale HBDs with two different active areas, 0.4×0.4 μm$^2$ (0.16 μm$^2$) and 0.3×0.3 μm$^2$ (0.1 μm$^2$) for developing detectors and FPAs at 200 GHz and 585 GHz operational frequencies respectively. Fig. 2.8 shows the simulated unmatched sensitivities ($\beta_{v}$) of submicron-scale HBDs for a frequency range from 10 GHz to 1 THz. The results show that HBDs with 0.4×0.4 μm$^2$ and 0.3×0.3 μm$^2$ device active areas have unmatched sensitivities of ~3800 V/W at 200 GHz and ~3000 V/W at 585 GHz operational frequency respectively. The maximum detector sensitivities ($\beta_{v,\text{max}}$) were estimated from their unmatched sensitivities using Eq. 2.3. The results show that the corresponding HBDs have maximum sensitivities of ~21000 V/W and ~9500 V/W at 200 GHz and 585 GHz operational frequencies respectively, as shown in Fig. 2.8. Although the performance of HBDs improve with scaling, HBDs have a fringing capacitance component that does not scale with device
active area and has been ignored in this analysis. This capacitance contribution becomes increasingly significant as device area is scaled into the submicron regime. As a result, further scaling may not improve detector performance significantly.

![Figure 2.9. Simulated noise equivalent power (NEP) of submicron HBDs under impedance matching for a frequency range from 10 GHz to 1 THz.](image)

2.3.5 Noise performance

Two types of noise play significant roles on the performance of HBDs: thermal noise arising from their internal resistance and bias dependent $1/f$ noise. The noise performance of HBDs is assessed by estimating their $NEP$s. The minimum detectable power or $NEP_{min}$ is a measure of minimum input power required to generate an output voltage equal to the HBD’s RMS noise voltage. Since $R_s \ll R_j$ for submicron scale HBDs, the contribution of noise from the series resistance has been ignored. Since HBDs are bias...
independent detector, the contribution of 1/f noise can also be ignored. Minimum NEP could be estimated from maximum detector sensitivity and junction resistance using [59],

\[ \text{NEP}_{\text{min}} = \frac{\sqrt{4k_BT R_j}}{\beta_{v,\text{max}}} \]  

(2.4)

where \( \beta_{v,\text{max}} \) is the maximum detector sensitivity under impedance matching condition. The \( \text{NEP}_{\text{min}} \) of HBDs were estimated for a frequency range from 10 GHz to 1 THz. Fig. 2.9 shows that HBDs have \( \text{NEP}_{\text{min}} \) of \(~0.42\) pW/Hz\(^{1/2}\) (for 0.16 \( \mu \)m\(^2\) HBD) and \(~1.3\) pW/Hz\(^{1/2}\) (for 0.1 \( \mu \)m\(^2\) HBD) at 200 GHz and 585 GHz respectively.

Although \( \text{NEP} \) defines the minimum detectable power at a single operational frequency, the sensitivity of HBD detectors is determined by their noise equivalent temperature difference (\( \text{NETD} \)) which is often used to assess the performance of direct detectors. In fact, \( \text{NETD} \) describes a minimum temperature difference that can be discerned within a scene being imaged, under the assumption of blackbody radiation and constant emissivity. This parameter relates the \( \text{NEP} \) to the detector’s RF bandwidth and the video bandwidth. For an unamplified direct detector, Johnson-noise limited \( \text{NETD} \) was derived to be [59]:

\[ \text{NETD} = \frac{\sqrt{4k_BT_0 R_d \Delta f_{\text{vid}}}}{\alpha k_B \beta_{v,\text{max}} \Delta f_{rf}} \]  

(2.5)

where \( T \) is the detector noise temperature, \( k_B \) is the Boltzmann constant, \( \alpha \) is the predetection RF attenuation factor to account for any loss in the matching network, \( \beta_{v,\text{max}} \) is the optimum voltage sensitivity under conjugate impedance matching conditions, \( \Delta f_{\text{vid}} \) is
the video bandwidth and $\Delta f_{rf}$ is the equivalent RF bandwidth. For passive THz imaging applications, NETDs of 0.5K or less are preferred.

2.3.6 Cutoff frequency

The maximum operational frequency of imaging systems is defined by their cutoff frequency or the 3-dB point of their sensitivity. Detectors with higher cut-off frequencies provide wider RF equivalent bandwidth which improves the noise performance and sensitivity of imaging systems significantly. The intrinsic cut-off frequency of HBD device is defined as

$$f_c = \frac{1}{2\pi R_s C_j}, \quad (2.6)$$

where $R_s$ and $C_j$ are the device series resistance and junction capacitance respectively. Eq. 2.6 shows that HBD’s cutoff frequency can be minimized by reducing $R_s$ and $C_j$. Although $C_j$ can be reduced by scaling device active area laterally to submicron regime, the scaling trend increase their series contact resistance. Therefore, a design trade-off is needed between the $R_s$ and $C_j$ to achieve highest cut-off frequency for each heterostructure design. For the heterostructure design shown in Fig. 2.3(a), the intrinsic cut-off frequency of a 0.16 $\mu m^2$ active area HBD fabricated at Notre Dame was demonstrated to be 650 GHz ($R_s=103 \ \Omega$ and $C_j=2.4 \ \text{fF}$) [59]. The series resistance of HBDs can be approximated by $R_s= R_c+R_a+R_{sp}$, where $R_c=\rho_c/A$, $\rho_c$ is the specific contact resistivity of cathode ohmic contact, $A$ is the device active area, $R_a$ and $R_{sp}$ are the anode resistance and frequency dependent spreading resistance respectively. For submicron-scale HBDs $R_c \gg R_a+R_{sp}$ and the contact
resistance can be reduced by lowering specific contact resistivity of cathode contacts. Fig. 2.10 shows the projected intrinsic cut-off frequency of 0.16 \( \mu \text{m}^2 \) HBDs as a function of specific contact resistivity. The results approximated that a highest intrinsic cut-off frequency of \(~8\) THz could be achieved by lowering the specific contact resistivity to \(1\times10^{-8} \Omega\cdot \text{cm}^2\) [93-95].

![Graph showing cut-off frequency vs. specific contact resistivity](image)

Figure 2.10. Projected cut-off frequencies of submicron HBDs as a function of specific contact resistivity for 0.16 \( \mu \text{m}^2 \) HBD detector.

2.3.7 Device impedance

As discussed previously, HBDs with laterally scaled sub-micron areas are required to obtain sufficiently low junction capacitance for operation at THz frequencies. However, device impedances of submicron-scale HBDs increase with their lateral scaling at THz frequencies [60]. Therefore, an accurate analysis of HBD’s device impedance is required.
for designing integrated HBD detectors and FPAs at THz frequencies. The impedance analysis results provide valuable insights for designing planar antennas to be integrated with HBDs.

Figure 2.11. Simulated device impedance of 0.16 \( \mu \text{m}^2 \) HBD for 200 GHz detector design (a) real part (b) imaginary part.

Figure 2.12. Simulated impedance of 0.1 \( \mu \text{m}^2 \) HBDs for 585 GHz detector design (a) real part (b) imaginary part.
The impedance analysis of submicron-scale HBDs, with device active areas of 0.16 \( \mu m^2 \) and 0.1 \( \mu m^2 \), was performed using ADS harmonic balance simulations of the scalable diode model of HBD [59], as shown in Fig. 2.11 and 2.12, respectively. The input impedance was estimated from first harmonic input current \( (I_1) \) and voltage \( (V_1) \) using \( Z_{in} = V_1/I_1 \). The simulation results show that the impedance of a 0.16 \( \mu m^2 \) active area HBD varies from 140 \( \Omega \) to 115 \( \Omega \) for its real part and -440 \( \Omega \) to -260 \( \Omega \) for its imaginary part for a frequency sweep from 150 GHz to 250 GHz. Similarly, the impedance of a 0.1 \( \mu m^2 \) active area HBD varies from 105.5 \( \Omega \) to 108.5 \( \Omega \) for its real part and -210 \( \Omega \) to -150 \( \Omega \) for its imaginary part for a frequency sweep from 500 GHz to 700 GHz. As can be seen, the device impedances of submicron HBDs are relatively higher and planar antennas that can potentially achieve high impedances at THz frequencies are usually required in a detector design for developing THz detectors.

<table>
<thead>
<tr>
<th>Device Area</th>
<th>Frequency</th>
<th>Device Impedance</th>
<th>Maximum Sensitivity</th>
<th>NEP</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.4x0.4 ( \mu m^2 )</td>
<td>200 GHz</td>
<td>125-j330 ( \Omega )</td>
<td>21000 V/W</td>
<td>0.42 pW/( \sqrt{Hz} )</td>
</tr>
<tr>
<td>0.3x0.3 ( \mu m^2 )</td>
<td>585 GHz</td>
<td>107-j202 ( \Omega )</td>
<td>9500 V/W</td>
<td>1.3 pW/( \sqrt{Hz} )</td>
</tr>
</tbody>
</table>
2.4 Summary

In this chapter, the figures of merit of high performance Sb-HBDs with active areas of 0.16 $\mu$m$^2$ and 0.1 $\mu$m$^2$ based on a nonlinear lumped element circuit model have been discussed. The ADS simulation results show that HBDs have maximum detector sensitivities of 21000 V/W and 9500 V/W at 200 GHz and 585 GHz respectively. Another important figure of merit is NEP which has been calculated, the results show that the HBD detectors have minimum NEP of 0.42, and 1.3 pW/Hz$^{1/2}$ at 200 and 585 GHz respectively. Moreover, the impedances of HBDs have been analyzed in ADS for the THz detector design. Table 2.1 summarizes the projected performance of submicron-scale HBDs at THz frequencies.
CHAPTER 3:

LENS-COUPLED FOLDED DIPOLE ANTENNAS FOR TERAHERTZ SENSING AND IMAGING

In chapter 2, the device impedances of HBDs at THz frequencies were analyzed. The analysis results show that HBDs have relatively high device impedance at THz frequencies and planar antennas that have high embedding impedance need to be employed for developing HBD based THz detectors and imaging FPAs. Planar FDAs are known to have high embedding impedance, and their impedance depends on the geometry of antenna structure. The unique feature of FDAs is that a wide range antenna impedance could be achieved by changing its antenna geometry (for both real and imaginary part) [96]. In addition, planar FDAs can be fabricated on semiconductor substrate for direct integration with HBDs and other passive components for developing THz detectors and FPAs.

In this chapter, the design, simulation, analysis, and testing results of LC-FDAs for THz sensing and imaging applications are presented. For an accurate analysis of LC-FDAs, even-odd mode analysis (EOA) is first performed to extract the related mode currents. On the basis of EOA, the antenna embedding impedance is evaluated from their relative mode currents and compared with the full-wave simulation results using ANSYS high frequency structure simulator (HFSS) [97]. HFSS simulation results show that a wide range of impedance values (both real and imaginary parts) can be achieved by modifying the geometry of FDAs for impedance matching to a variety of devices (e.g. Schottky diodes
and HBDs) without additional matching networks. In addition, the far-field radiation patterns, directivity and Gaussian coupling efficiency of LC-FDAs have been studied using ray-tracing techniques. Good agreement between calculated results and measurements has been obtained, demonstrating the effectiveness of this analysis approach.

3.1 THz quasi-optical configuration

Classical microwave technologies based on metallic waveguides are widely used and attractive for developing highly sensitive detectors at microwave frequencies for its low loss characteristics. However, the detection of RF signals becomes quite challenging at terahertz frequencies largely because of difficulties associated with extending classical technologies and techniques at this frequency regime. As the frequency increases, it is more challenging in fabricating waveguide circuits for use due to cost and tolerances associated with machining small waveguide structures. In addition, the surface resistivity of metallic waveguides, which is proportional to $\sqrt{\omega \mu / 2 \sigma}$, introduces high frequency losses at THz frequencies. Moreover, waveguide detector modules are bulky and difficult to expand into two-dimensional and large scale focal-plane arrays (FPAs). As an alternative, quasi-optical technology, which combines both the microwave and optics techniques, is attractive and suitable for developing highly sensitive detectors and imaging FPAs for its simplicity and low manufacturing cost.
THz quasi-optical detectors are typically designed and fabricated on dielectric substrate such as GaAs, silicon, or quartz. The detector on dielectric substrate is comprised of detection device and receiving antenna. The incident THz signal is guided by an objective lens and focused onto the dielectric substrate to illuminate the detectors. In order to maximize the directivity of the receiving antenna, the detector is illuminated from the dielectric side. However, the receiving antenna generates trapped surface waves in the dielectric substrate, which significantly decreases antenna efficiency as well as increase cross-talk between adjacent antennas in a two-dimensional array [98-101]. The crosstalk between adjacent antennas limits the imaging resolution of an imaging array. Fig. 3.1 illustrates the trapped surface waves in dielectric substrate using geometric optics where the transmitting antenna lies in the air/dielectric interface and radiates into the substrate. The radiation from the transmitting antenna with incident angles of $\theta < \theta_c$ can go through the air-dielectric interface, where $\theta_c$ is the critical angle. For incident angles
\[ \theta > \theta_c, \text{ the radiation reflected back into the substrate contributes to surface-wave loss. The total surface-wave loss can be estimated using, } P_{s,\text{loss}} = 3\varepsilon_r \lambda_0 (2m + 1)/32h + \lambda_0^3 (m^2 + m)(2m + 1)/128h^3, \text{ where } \varepsilon_r \text{ is the dielectric constant of the substrate, } m \text{ is the number of the highest propagation mode and } h \text{ is the substrate thickness [99].} \]

![Diagram](image)

Figure 3.2. Antennas mounted on extended-hemispherical silicon lens to eliminate surface wave losses.

In order to obtain good antenna efficiency, surface wave losses have to be reduced and the quasi-optical configuration based on the “reverse-microscope” concept, first proposed by Rutledge and Muha in 1982 [98], can be adopted for THz sensing and imaging applications. In this configuration, an extended hemispherical silicon lens (which has similar dielectric constant as antenna substrate) is attached to the detector substrate for minimizing surface wave losses, as shown in Fig. 3.2. Both objective lens and substrate
lens are utilized in this configuration and the approach is capable of demonstrating two-dimensional diffraction limited imaging.

Different types of lens-coupled planar antennas including bow-tie antennas, double-slots, annular-slots, sinuous and log-periodic antennas have been employed to THz sensing and imaging applications [29,102]. However, these planar antennas tend to exhibit relatively low embedding impedances (i.e., < 100 Ω) at THz frequencies or limited flexibility. On the other hand, THz detectors and imaging focal-plane arrays (FPAs) at room-temperature have been designed and demonstrated using semiconductor devices such as Schottky diodes and HBDs [88,103]. These devices typically have high impedances (both real and imaginary parts) at frequencies (~1 THz) below their cutoff frequencies. However, the power coupling efficiency of these detectors is limited due to the substantial impedance mismatch between detection devices and low-impedance planar antennas, resulting in relatively low detector performance. Although impedance matching networks can be utilized to maximize detector’s performance, this unavoidably leads to reduced detection bandwidth and large circuit footprints and increases the system complexity. To overcome these issues, lens-coupled planar antennas that can potentially achieve high embedding impedances are needed for realizing high-performance, room-temperature THz detectors and imaging focal-plane arrays.

3.2 Lens-coupled planar folded dipole antennas

Planar FDAs are known to have high embedding impedance at THz frequencies, and their embedded impedance depends on the antenna geometry. The unique feature
of FDAs is that a wide range impedance variation (for both real and imaginary part) could be achieved by changing their antenna geometry. In addition, these antennas can be fabricated on semiconductor substrates for which planar detectors using FDAs are realizable. Moreover, FDAs permit design flexibility to integrate with high impedance detection devices (i.e. HBDs, Schottky diodes) to achieve conjugate impedance matching condition and maximum detector sensitivity without additional impedance matching network.

Figure 3.3. Planar folded dipole antenna with number of arm \((N)\), arm length \((L)\), arm width \((w)\) and arm gap \((g)\).

Conventional FDAs consist of a closed loop geometry [104,105], which results in a DC short-circuit for the output, making them not well suited for detection applications. As an alternative, planar FDA design with an open geometry, as shown in Fig. 3.3, has been proposed and demonstrated for THz photomixers applications for improved matching condition to enhance output power [106-108]. In addition, some preliminary research has been carried out for Schottky-diode-based detector designs using lens-
coupled planar FDAs (LC-FDAs) at 75 GHz and 200 GHz, showing promising performance for detection applications [109,110]. In the following sub-sections, the RF parameters of LC-FDAs at THz frequencies such as FDA theory, embedding impedance, far-field radiation pattern, directivity and Gaussian coupling efficiency will be discussed in detail.

3.2.1 Even-odd mode analysis

FDAs with N number of arms can be considered as an array of N number linear dipole antennas assuming that the arm gap, \( g<<\lambda_d \) (guided wavelength). To analyze the LC-FDA (see Fig. 3.2), even-odd mode analysis (EOA) has been performed [104,105]. This technique makes use of the physical symmetry of the FDA. As shown in Fig. 3.4, the analysis begins by decomposing the current components of the FDA into two distinct modes, namely, even-mode and odd-mode. By applying EOA, the single-port FDA with N number of arms can be decomposed into even-mode and odd-mode N-port networks. The solution of the original single-port problem can be treated as a superposition of the

![Diagram](image)

Figure 3.4. Even-odd mode analysis: decomposition of FDA into even-mode and odd-mode circuits for extraction of the radiating antenna currents [96].
solutions of the two N-port sub-problems. This approach permits the direction calculation of the embedding impedance and radiation patterns of the LC-FDAs from the extracted mode currents.

As an example, an LC-FDA with a number of arms \( N = 3 \), antenna length of \( l = 285 \) μm, arm width \( w = 12 \) μm, arm gap \( g = 12 \) μm has been analyzed from 150 GHz to 250 GHz. Assuming the FDA is excited by a rms voltage \( V \) (1.0 V used in the analysis) at the feed point (middle of the center antenna arm), the total rms current flowing through the antenna is \( I \). Fig. 3.4 illustrates the decomposition of the FDA’s feed point voltage \( V \) and total current \( I \) into even-mode and odd-mode using superposition theorem. The even-mode has voltages \( V/3 \), \( V/3 \) and \( V/3 \) (from left to right) for the three ports, and the associated currents in three antenna arms are \( I_{e1} \), \( I_{e2} \), and \( I_{e3} \) respectively. Similarly, the odd-mode has voltages of \(-V/3\), \( 2V/3 \) and \(-V/3\), and currents of \( I_{o1} \), \( I_{o2} \), and \( I_{o3} \) in each arm. Since the FDA radiates largely from the even-mode currents and the parallel arms of the FDA cause the odd-mode currents to cancel out in the far field, the two mode currents are also called the radiation mode current and transmission line mode current, respectively [104,105]. The current and voltage for each mode can be related using a linear combination of the self and mutual impedances of the \( N \)-port network, where the ports are located at the center of each arm. For the even-mode circuit, the resulting current and voltage equations are given by (for the case of \( N = 3 \):
where $I_{e1}$, $I_{e2}$, and $I_{e3}$ are the even-mode terminal currents and $Z$ represents the impedance matrix of the 3-port network which has been computed using conventional full-wave simulations. For the odd mode circuit, the relations are given by:

$$\begin{pmatrix}
\frac{-V}{3} \\
\frac{2V}{3} \\
\frac{-V}{3}
\end{pmatrix} =
\begin{bmatrix}
Z_{11} & Z_{12} & Z_{13} \\
Z_{21} & Z_{22} & Z_{23} \\
Z_{31} & Z_{32} & Z_{33}
\end{bmatrix}
\begin{pmatrix}
I_{o1} \\
I_{o2} \\
I_{o3}
\end{pmatrix},\quad (3.2)$$

where $I_{o1}$, $I_{o2}$, and $I_{o3}$ represent the odd mode terminal currents, and the $Z$ matrix remains unchanged. Fig. 3.5(a) shows the resulting even-mode currents for this example FDA for a frequency range from 150 GHz to 250 GHz. At the resonant frequency, the antenna radiates from the real part of the even-mode currents and provides distortion-less radiation patterns [105]. The detailed computation of the far-field radiation patterns using the even-mode currents will be described in subsection 3.2.3. Once the mode currents (both even-mode and odd-mode) are determined, the embedding impedance ($Z_{in}$) of the FDAs (at the feed point of the original antenna problem shown in Fig.3.4) can be calculated from the current ratios and impedance matrix. The resulting equation for $Z_{in}$ is given by:
\[ Z_{in} = \frac{V}{I_2} = \frac{I_1}{I_2} Z_{21} + \frac{I_3}{I_2} Z_{23}, \]  

where \( I_1 = I_{e1} + I_{o1} \), \( I_2 = I_{e2} + I_{o2} \) and \( I_3 = I_{e3} + I_{o3} \). Using the above formulation, the real and imaginary parts of the driving point impedance (embedding impedance) have been calculated as shown in Fig. 3.5(b). The embedding impedance of the FDA has also been simulated directly using full-wave HFSS simulation. As can be seen in Fig. 3.5(b), the results obtained from EOA show quite good agreement with HFSS simulation, demonstrating the effectiveness of the proposed analysis approach.

3.2.2 Embedding impedance

On the basis of the EOA described in section II, it is seen that the LC-FDA \((N=3)\) can achieve much higher embedding impedance at its resonant frequency (i.e., 750 Ω at 215 GHz) as compared to a simple dipole antenna. According to antenna theory, the embedding impedance of a FDA is given by \( Z_{in} = N^2 Z_0 \), where \( Z_0 \) is the impedance of the (conventional) single dipole [104]. Owning to the above property, the proposed FDA offers additional flexibility to present a wide range of impedances by varying the antenna geometry. This provides an opportunity to conjugate match high-impedance devices such as HBDs (e.g., typical device impedance of 124- j 330 Ω at 200 GHz) for maximum sensitivity without additional matching networks. This approach may significantly reduce the footprint of single detectors (or imaging pixels) and enable large-scale THz FPAs. In order to tune the embedding impedance of the FDA, the antenna geometry (i.e., \( N, w \) and \( g \) as shown in Fig. 3.3) needs to be adjusted.
Figure 3.5. Analysis of FDAs using EOA: (a) Real and imaginary part of even-mode currents of the 200 GHz FDA for a frequency range from 150 GHz to 250 GHz calculated using EOA. (b) Comparisons of simulated embedding impedance of FDA using EOA and full-wave simulations [96].

To investigate and demonstrate the impedance tuning capacity of the planar FDAs, HFSS full-wave simulations have been performed for FDAs on semi-infinite substrate with appropriate boundary conditions (since the lens dimension is much larger...
than that of the antenna) for a frequency range from 100 GHz to 300 GHz, with a goal to match the HBD device impedance of \(\sim 124\,\text{j330} \,\Omega\) at 200 GHz. Fig. 3.6 shows the simulation results for the embedding impedance of a FDA designed with a center frequency of \(\sim 200\,\text{GHz}\) by varying its number of arms, \(N\). The antenna length was fixed at \(l = 285 \,\mu\text{m}\). When \(N\) increases from 1 to 9, the real part of the impedance at resonance varies from 10 Ω to 1800 Ω, while the maximum imaginary impedance (near resonance) ranges from 10 Ω to 900 Ω. The effect of arm width was also investigated. Fig. 3.7 shows how the embedding impedance of an FDA varies with arm width \(w\) for antennas with \(N=3\) and \(g=12 \,\mu\text{m}\). As can be seen in Fig. 3.7 (a) and (b), as \(w\) increases from 12 \,\mu\text{m} to 36 \,\mu\text{m}, the real part of the antenna impedance at resonance changes from 10 Ω to 700 Ω, while the imaginary part near resonance varies from 10 Ω to 300 Ω. Finally, HFSS simulation results also show that the FDA impedance can be tuned by varying its arm gap from 12 \,\mu\text{m} to 36 \,\mu\text{m} for antennas with \(N=3\) and \(w=12 \,\mu\text{m}\). As shown in Fig. 3.8 (a) and (a), the real part of impedance varies from 10 Ω to 730 Ω and its imaginary part changes from 10 Ω to 300 Ω, for \(g\) from 12 \,\mu\text{m} to 36 \,\mu\text{m}. These results demonstrate the capability to design an FDA impedance-matched to a wide range of load impedances (e.g., \(\sim 124\,\text{j330} \,\Omega\) at 200 GHz for Sb-HBDs), without the need for intervening separate matching networks. In addition, the proposed LC-FDA is also suitable for operation at higher THz frequencies. The geometry of the FDA (e.g., antenna length, arm number, arm gap and arm width) needs to be scaled and modified to achieve impedance-tuning capacity for conjugate matching with devices.
Figure 3.6. Simulation results for a FDA: (a) real part and (b) imaginary part of antenna embedding impedance with varying number of arms (N) from N=1 to N=9 [96].

Figure 3.7. Simulation results for a FDA: Real part (a) and imaginary part (b) of antenna embedding impedance for FDAs (N=3) with arm widths ranging from 12 µm to 36 µm [96].
Figure 3.8. Simulation results for a FDA: Real part (a) and imaginary part (b) of antenna embedding impedance for FDAs ($N=3$) with arm gaps ranging from 12 $\mu$m to 36 $\mu$m [96].

Figure 3.9. Lens-coupled FDA mounted on an extended hemispherical silicon lens of radius $R$ and extension length $L$ for ray-tracing analysis [96].
3.2.3 Radiation properties

In order to achieve high directive gain and antenna efficiency, a high-resistivity (20000 Ω.cm) silicon lens was chosen since it has the same dielectric constant ($\varepsilon_r=11.7$) as the GaAs antenna substrate. By utilizing an extended hemispherical lens (extension length $L$ as shown in Fig. 3.9), the gain and efficiency can be optimized for different applications, and the surface wave losses can be minimized. As noted in Section I, the radiation properties of LC-FDAs in the THz region have not yet been fully studied and reported. For these studies, one could consider performing full-wave simulation of the complete lens-coupled antenna geometry (e.g., using HFSS), but this would require significant computer resources and long simulation times. An alternative approach is to apply the ray-tracing technique described in [111]. However, application of the ray-tracing technique requires a closed-form solution for the antenna patterns without lenses. Using the EOA discussed in Section II, an array of currents for the even (radiation) mode has already been computed; these currents are used to study the LC-FDA radiation properties using the ray-tracing technique in the following sub-sections.

3.2.3.1 Far-field radiation pattern

The far-field radiation patterns of the LC-FDA have been calculated using the ray-tracing technique [96,111] on the basis of the EOA. If we first assume the LC-FDA is driven by a sinusoidal excitation current at the antenna feed point, then the antenna arms can be treated as an array of current sources radiating into the silicon, and the normalized element pattern inside the dielectric associated with each antenna arm is given by [96]:

\[ \text{pattern} \]
\[ R_n(\theta) = \frac{\sin\theta}{k_m^2 - ke^2 \cos^2 \Theta} \left[ \cos (ke l \cos \Theta) k_m - \cos l \right] , \quad (3.4) \]

where \( k_e \) and \( k_m \) are the propagation constants for the dielectric side and the air side respectively and \( \Theta \) is the angle with respect to the z axis. The resultant radiation patterns of the full FDA can be calculated by including an array factor in the H-plane direction to combine the effects of each arm into the overall antenna pattern. For \( N=3 \), the array factor becomes

\[ A_f = I_{e2} + (I_{e1} + I_{e3}) \cos (k_e d \sin \Theta \cos \phi) , \quad (3.5) \]

where \( I_{e1}, I_{e2} \) and \( I_{e3} \) are the radiating even-mode currents calculated using Eq. (1), \( d \) is the spacing between two adjacent arms, and \( \phi \) is the angle from the x axis in the x-y plane. Fig. 3.10 shows the calculated radiation patterns inside the lens dielectric for different planes.

![Radiation Patterns](image)

Figure 3.10. Calculated radiation intensity of FDA mounted on silicon substrate at 200 GHz.
Figure 3.11. Co-ordinate systems for the analysis of lens-coupled FDAs using ray tracing techniques.

Once the radiation patterns inside the dielectric medium are calculated, the far-field radiation patterns of the entire structure can be obtained from the lens surface field [111] using the ray-tracing technique. The field components at air/dielectric interface are subsequently decomposed into parallel/perpendicular components. The electric and magnetic field just outside of the air/dielectric interface are found by Fresnel equations for the transmission coefficient:

\[
I_\parallel = \frac{n\sqrt{1 - n^2\sin^2\theta_i^2} - \cos\theta_i}{n\sqrt{1 - n^2\sin^2\theta_i^2} + \cos\theta_i}, \quad (3.6)
\]

\[
\tau_\parallel = (1 + I_\parallel) \frac{\cos\theta_i}{n\sqrt{1 - n^2\sin^2\theta_i^2}}, \quad (3.7)
\]
\[ \Gamma_\perp = \frac{n \cos \theta_i - \sqrt{1 - n^2 \sin^2 \theta_i^2}}{n \cos \theta_i + \sqrt{1 - n^2 \sin^2 \theta_i^2}} \]  
(3.8)

\[ \tau_\perp = (1 + \Gamma_\perp), \]  
(3.9)

where \( n \) is the dielectric constant, \( \theta_i \) is the angle of incidence from the normal to the spherical lens. \( \Gamma \) and \( \tau \) are the reflection and transmission coefficients for the parallel and perpendicular polarization respectively. The equivalent electric and magnetic surface current densities at the lens surface can be calculated from electric and magnetic field components at the lens/air interface using:

\[ J_s = \hat{n} \times H, \]  
(3.10)

\[ M_s = \hat{n} \times E, \]  
(3.11)

where \( \hat{n} \) is the normal to the interface, as shown in Fig. 3.11. In the far-field, the transverse electric field can be calculated by:

\[ E_\theta = -\frac{jk e^{-jk r}}{4\pi r} \left( L_\phi + \eta N_\theta \right), \]  
(3.12)

\[ E_\phi = -\frac{jk e^{-jk r}}{4\pi r} \left( L_\theta + \eta N_\phi \right), \]  
(3.13)

where \( L \) and \( N \) are defined by:

\[ N = \int_S J_s e^{jk r' \cos \psi} dS', \]  
(3.14)
\[ L = \iint_S M_s e^{i k r' \cos \psi} dS', \]  

where \( S' \) is the closed surface just outside the lens, \( r' \) is the distance from the center of the sphere to the equivalent electric and magnetic currents, \( r \) is the distance from the center of the sphere to the far-field point, and \( \psi \) is the angle between \( r \) and \( r' \).

Figure 3.12. Simulated E and H-plane radiation patterns of a 200 GHz LC-FDA with \( N=3, l = 285 \ \mu m, g=12 \ \mu m, \) and \( w=12 \ \mu m \) for extension lengths \( L \) varying from 1.8 mm to 2.5 mm using the ray-tracing technique [96].
Fig. 3.12 shows the simulated far-field radiation patterns of the lens-coupled \((R = 6 \text{ mm})\) FDA by varying the extension length \((L)\) from 1.8 mm to 2.5 mm. It is clearly seen that the main beam of the far-field patterns in both the E- and H-planes first become narrower and then broader as the extension length is increased from 1.8 mm to 2.5 mm. The calculated radiation patterns in the E- and H-planes exhibit the highest directivity for a 2.2 mm extension length (or \(L/R \sim 0.37\)).

3.2.3.2 Antenna directivity, Gaussian coupling efficiency and radiation efficiency

Once the radiation patterns of the LC-FDAs are obtained, the gain (or directivity) and Gaussian coupling efficiency of the lens-coupled FDA are calculated for varying lens geometry, especially the extension length \((L)\) and radius \((R)\) of the silicon lens. For rotationally symmetric radiation patterns, the directivity can be calculated from the 3-dB beam width of E- and H-plane radiation patterns using [104]:

\[
D_0 = \frac{32 \ln 2}{\theta_E^2 + \theta_H^2} \quad (3.16)
\]

As shown in Fig. 3.13(a), the directivity of the lens-coupled FDA has been calculated (on the basis of the radiation pattern calculations using the ray-tracing technique) for different lens geometries by varying both \(L\) (from 1.0 mm to 3.5 mm) and \(R\) (from 4 mm to 8 mm). For fixed lens radius, the antenna directivity initially increases and then drops as \(L\) increases. Each directivity curve (for fixed \(R\)) in Fig. 3.13(a) reaches a maximum approximately centered at \(L/R \sim 0.37\). Similar trends have been reported in [111] for lens-coupled double-slot antennas.
Although larger lens radius generally results in higher antenna directivities, Gaussian coupling efficiency (or Gaussicity, defined as the coupling efficiency of an antenna radiation pattern to the far-field pattern of an ideal Gaussian beam) must also be considered for designing a LC-FDA suitable for THz detection or imaging. The Gaussicity can be estimated with a double integration of electric-field patterns of the antenna \(E_{antenna}\) and the incident Gaussian beam \(E_{Gauss}\) [111]. However, the computation of Gaussicity using numerical code is very time consuming. To quickly estimate the Gaussicity for radiation coupling efficiency calculation, the Gaussian beam parameter in the far-field can be estimated using [111]:

\[
E_{Gauss}(\theta, w) = \exp\left(-\left(\frac{kw}{2} \sin \theta\right)^2\right),
\]

where, \(k\) is the propagation constant and \(w\) is the beam waist. The Gaussicity is found to be,

\[
\eta = \frac{K^2}{N \cdot G},
\]

where \(K\), \(N\) and \(G\) are defined by,

\[
K = \iint E_{antenna} \cdot E_{Gauss} \cdot \sin \theta d\varphi d\theta,
\]

\[
N = \iint E_{antenna} \cdot E_{antenna} \cdot \sin \theta d\varphi d\theta,
\]

\[
G = \frac{\pi}{2} \int_0^\frac{\pi}{2} E_{Gauss}^2 \cdot \sin \theta d\varphi d\theta.
\]
Figure 3.13. Directivity and Gaussicity calculation: (a) Simulated directivity of the LC-FDA for different lens radii and extension lengths, and (b) simulated maximum directivity and Gaussicity for lens radii from R=4 mm to R=8 mm [96].
The $E_{\text{antenna}}$ is calculated from Eqn. 3.12 and 3.13. The computed directivity and Gaussicity of LC-FDAs for varying lens radius is shown in Fig. 3.13(b). The computed results show that the peak directivity increases progressively from 24 dB to 30 dB with increase in lens radius from 4 mm to 8 mm. On the other hand, the Gaussicity of the LC-FDA decreases from 94% to 84%. For a single-pixel THz detector design using LC-FDAs, $L$ and $R$ should be chosen to provide the highest Gaussian coupling efficiency (to achieve maximum sensitivity) while still obtaining a good antenna gain (directivity). In contrast, however, higher gain/directivity (by optimizing $L$ and $R$) is desired in imaging FPAs to achieve high imaging resolution.

In addition to directivity and Gaussicity, HFSS simulations have been performed to estimate the radiation efficiency of FDAs at 200 GHz. Due to computational limitations for simulating the lens-coupled structure, the simulations have been performed for the FDAs on semi-infinite high-resistivity silicon substrate. The result shows that a radiation efficiency of 91% can be obtained at 200 GHz under conjugate matching condition. It can be expected the lens-coupled configuration with a matching cap layer at the air/lens interface would increase the radiation efficiency of FDA significantly.

3.2.3.3 Antenna characterization

In order to experimentally validate the antenna analysis, a 200 GHz ($l = 285 \ \mu m$) FDA with $N = 3$, arm width $w=12 \ \mu m$ and arm gap $g=12 \ \mu m$ was fabricated on a 500 $\mu m$ thick high resistivity ($\geq 20,000 \ \Omega.cm$) silicon wafer ($\varepsilon_r = 11.7$, using a conventional photolithography and Au plating process. A zero-bias Schottky diode (ZBD) provided by Virginia
Diodes Inc. (VDI) was flip-chip mounted at the antenna feed point for use as a simple power detector (not optimized with impedance conjugate matching), as shown in Fig. 3.14 (a) and (b). The antenna was connected to DC output pads using two meander lines. The antenna was mounted to a silicon lens with $R = 5$ mm and $L = 1.7$ mm (total extension length including the wafer thickness) for quasi-optical measurements and characterization. The far-field radiation patterns of the LC-FDA have been measured at $\sim 200$ GHz using VDI frequency extension modules (FEMs, 190-210 GHz) for providing the THz radiation. During the measurement, the detector was mounted on a computer-controlled rotation stage. The incident RF was amplitude-modulated, and the detector output DC signal was detected and amplified using a lock-in amplifier.

Figure 3.14. Antenna characterization results: (a) The fabricated 200 GHz FDA circuit mounted to the backside of a silicon lens for antenna characterization. (b) Higher magnification view of the FDA with a flip-chip mounted zero-bias Schottky diode as a direct power detector. Inset: top view of Schottky diode [96].
Figure 3.15. Antenna characterization results: Measured and calculated radiation patterns of LC-FDA at 200 GHz in the (a) E-plane (b) H-plane [96].

As shown in Fig. 3.15 (a) and (b), the measured radiation patterns (both E- and H-plane) show nearly Gaussian-shape main beams with side-lobe levels less than -14 dB. The 3-dB beam width of the FDA in the H-plane is ~10° at 200 GHz. The measured E-plane radiation pattern is slightly broader than the pattern in the H-plane, as expected [111].
Simulated results for E- and H-plane radiation patterns are included for comparison. The discrepancies between the simulation and measurement results may be attributed, at least in part, to the DC readout meander lines (see Fig. 3.14) that are present in the experimental configuration and slightly affect the antenna properties but were not considered in the theoretical analysis. In addition, although the antenna has been designed for integrated THz detectors, flip-chip mounted Schottky diode detector (with large pad) has been used in this measurement and contributed to discrepancies. In a practical THz detector design using LC-FDAs, integrated power detection diodes (e.g., Schottky or HBD) are preferred (vs. the flip-chip mounting used here) and a more sophisticated DC signal output circuit (e.g., using a coplanar strip (CPS) low-pass filter structure) would be advantageous to suppress the RF currents on the DC lines.

3.3 Summary

In this chapter, the LC-FDAs have been evaluated for THz detection and imaging applications. Even-odd mode analysis (EOA) was performed to analyze the antennas and extract the mode currents for determining the embedding impedance and facilitating radiation property calculations. The antenna embedding impedance was studied for different antenna geometries; the results show that a wide range of impedance values (both real and imaginary parts) can be achieved. This unique property makes LC-FDAs especially attractive for high-performance THz detectors in which impedance matching between antennas and small-area detectors is required. On the basis of the currents calculated using EOA, the LC-FDA far-field radiation patterns, antenna directivity and
Gaussian coupling efficiency for different lens structures have been studied using the ray-tracing technique. Good agreement between calculation results and experimental measurements has been obtained, demonstrating the effectiveness of the above analysis. In order to evaluate the FDAs for two-dimensional (2-D) THz FPAs with high imaging resolution, the lens structure has been optimized and the off-axis radiation patterns of the FPAs are evaluated. The results suggest that LC-FDAs are promising for realizing high-performance THz detectors and FPAs.
CHAPTER 4:

SINGLE ELEMENT HBD DETECTORS: DESIGN, FABRICATION AND CHARACTERIZATION

In this chapter, the design, fabrication, and characterization results of single element HBD detector are reported. As discussed in chapter 3, planar FDAs, which are known to have high embedding impedance at THz frequencies, are attractive for integration with HBDs for their impedance tuning capacity over a wide frequency range in the THz region. Direct integration of HBDs into FDAs not only improves detector’s reliability, but also minimizes circuit losses (i.e. losses due to dielectric, impedance mismatch and parasitic effects) at THz frequencies. For prototype demonstration, HBDs with 0.16 μm² and 0.1 μm² device active areas have been employed for developing THz detectors at 200 and 585 GHz operational frequencies respectively [112,113]. In order to achieve detector performance close to the intrinsic detectors, the parasitic capacitance introduced by metal interconnects and the frequency dependent spreading resistance for current path in the InAs anode layer are studied. The integrated detectors with sufficiently low parasitic effects were developed using a novel, scalable, and robust fabrication process to achieve high accuracy and reproducibility for HBD detectors. DC and quasi-optical characterization were performed to evaluate the performance of the HBD detectors. The performance of these detectors was assessed by measuring their sensitivity, noise equivalent power (NEP), and radiation patterns at 170 GHz and 200 GHz [114].

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Figure 4.1. Lens-coupled quasi-optical configuration of a single element HBD detector where a substrate lens (e.g., extended hemispherical dielectric lens, with same or similar dielectric constant as the antenna substrate) is attached to GaAs substrate containing an HBD detector [114].

Figure 4.2. Equivalent circuit model of the lens-coupled HBD detector where $P_{in}$ is the incident radiation, $Z_s$ is the antenna impedance. The diode is represented by its nonlinear equivalent circuit model.
4.1 Single element HBD detector

Fig. 4.1 shows the lens-coupled quasi-optical configuration of a single element HBD detector where a substrate lens (e.g., extended hemispherical dielectric lens, with same or similar dielectric constant as the antenna substrate) is attached to GaAs substrate containing an HBD detector. The equivalent circuit model of the lens-coupled detector is shown in Fig. 4.2 where $P_{in}$ is the incident radiation, $Z_s$ and $Z_d$ are the antenna and diode impedance respectively. The diode is represented by its nonlinear equivalent circuit model. The power transferred to the detector ($P_d$) can be expressed by,

$$P_d = \frac{1}{2} \frac{|v_s|^2}{|Z_d + Z_s|^2} \text{Real}(Z_d).$$  \hspace{1cm} (4.1)

where $v_s$ is the source voltage. For conjugate impedance matching condition, the power delivered to the detector is the power available from the antenna, i.e. $P_o = P_d$.

Figure 4.3. Non-linear $I$-$V$ characteristics of a typical Sb-HBD device. The detector output is a rectified DC signal.
Since HBD is a non-linear device, the incident RF signals generate a DC voltage (or current) across the device, as shown in Fig. 4.3. In order to analyze the characteristics of HBD detector, its \( I-V \) characteristic can be expanded using Taylor series about the bias voltage \( V=V_0 \),

\[
i(V) = i(V_0) + \frac{di}{dV} \bigg|_{V=V_0} (V - V_0) + \frac{1}{2!} \frac{d^2i}{dV^2} \bigg|_{V=V_0} (V - V_0)^2 \\
+ \frac{1}{3!} \frac{d^3i}{dV^3} \bigg|_{V=V_0} (V - V_0)^3 \ldots \ldots,
\]  

(4.2)

For a zero-bias detector \( V_0 = 0 \) V and equation 4.2 can be simplified to

\[
i(V) = \frac{di}{dV} \bigg|_{V=0} V + \frac{1}{2!} \frac{d^2i}{dV^2} \bigg|_{V=0} V^2 + \frac{1}{3!} \frac{d^3i}{dV^3} \bigg|_{V=0} V^3 \ldots \ldots,
\]  

(4.3)

The applied voltage across the diode is sinusoidal and can be expressed by

\[
V(t) = v_{RF} \cos \omega t.
\]  

(4.4)

The applied voltage is small and the first two terms in eqn. (4.3) mainly represent the diode nonlinearity (i.e. high order terms have negligible contribution to nonlinearity). The current flowing through the device can be shown to be

\[
i(t) = \frac{di}{dV} \bigg|_{V=0} v_{RF} \cos \omega t + \frac{1}{2} \frac{d^2i}{dV^2} \bigg|_{V=0} v_{RF}^2 \cos^2 \omega t \\
- \frac{1}{4} \frac{d^2i}{dV^2} \bigg|_{V=0} v_{RF}^2 + \frac{1}{4} \frac{di}{dV} \bigg|_{V=0} v_{RF} \cos \omega t + \frac{1}{4} \frac{d^2i}{dV^2} \bigg|_{V=0} v_{RF}^2 \cos 2\omega t
\]

Taking the time average of current, the detector current can be expressed by
We can see that the detector output current (or voltage) is proportional to the incident input RF power and these detectors are therefore called “square law detectors”.

For the HBD based unamplified direct detector shown in Fig. 4.2, it is very important to achieve best signal-to-noise ratio to insure maximum power transfer to the diode detector. This can be achieved using an impedance matching network between the diode and antenna. Since FDAs have high embedding impedance at THz frequencies, maximum power to the HBD can be delivered by achieving conjugate impedance matching condition without any additional matching networks. Although impedance matching can be achieved at resonance frequency, the impedance mismatch between the diodes and antennas at the both sides of resonance frequency decreases detector’s sensitivity. The reflection co-efficient due to impedance mismatch can be expressed as a function of frequency by,

\[
|\Gamma(f)| = \left| \frac{Z_d(f) - Z_s^*(f)}{Z_d(f) + Z_s^*(f)} \right|,
\]

(4.6)

and the frequency dependent detector sensitivity can be obtained using,

\[
\beta_v(f) = \beta_{v,max}(f) (1 - |\Gamma(f)|^2),
\]

(4.7)

where \(\beta_{v,max}\) is the intrinsic sensitivity of HBD device. The average output voltage over the RF bandwidth is related to frequency dependent detector sensitivity and power absorbed by the diode which can be expressed by [58]:

\[
\Delta i = \langle i(t) \rangle = \frac{1}{4} \left. \frac{d^2i}{dV^2} \right|_{V=0} v_{RF}^2
\]

(4.5)
where \( S_a \) is the spectral density of the power available from the source.

\[
V_{out} = \int_{f_{min}}^{f_{max}} \beta_v(f) S_a(f) df, \tag{4.8}
\]

Figure 4.4. (a) FDA design with \( N=5 \) for conjugate impedance matching (both real and imaginary) to an HBD device with 0.16 \( \mu m^2 \) active device area at 200 GHz. (b) Simulated detector sensitivity and NEP under conjugate impedance matching condition at 200 GHz.

Figure 4.5. (a) FDA design with \( N=3 \) for conjugate impedance matching (both real and imaginary) to an HBD device with 0.16 \( \mu m^2 \) active device area at 200 GHz. (b) Simulated detector sensitivity and NEP under conjugate impedance matching condition at 200 GHz.
4.1.1 THz HBD detector design

On the basis of above analysis, highly sensitive HBD detectors employing FDAs can be designed and developed for THz sensing and imaging applications. For a prototype demonstration, FDAs were employed for impedance matching to HBDs with device active areas of 0.16 \( \mu m^2 \) and 0.1 \( \mu m^2 \) at 200 and 585 GHz respectively. The HBD with 0.16 \( \mu m^2 \) active area has a device impedance of 125-j330 \( \Omega \) at 200 GHz, as discussed in chapter 2. In order to achieve conjugate impedance matching for maximum detector sensitivity at 200 GHz operational frequency, two FDAs were designed by modifying their antenna geometry. Since FDAs are resonant dipole antenna, the length of antenna \( (l) \) was kept slightly shorter than \( \lambda_d/2 \) for inductive antenna impedance for impedance matching, where \( \lambda_d \) is guided wavelength. The FDA for 200 GHz detector has five arms \( (N=5) \) with antenna arm length \( (l) \) of 275 \( \mu m \), arm width \( (w) \) of 15 \( \mu m \), and arm gap \( (g) \) of 12 \( \mu m \). Simulation results show that, nearly perfect conjugate impedance matching condition has been achieved, as shown in Fig 4.4 (a). Under such matching condition, a maximum detector sensitivity of 21000 V/W and a NEP\(_{\text{min}}\) of 0.42 pW/\( \sqrt{\text{Hz}} \) have been projected at 200 GHz from simulations [112], as shown in Fig. 4.4 (b). Another FDA with \( N=3 \) was designed for the 200 GHz detector. That antenna has \( l=275 \mu m, w=12 \mu m, \) and \( g=10 \mu m \). The real and imaginary parts of the antenna impedance are shown in Fig. 4.5 (a) together with HBD’s device impedance. The results show that nearly conjugate impedance matching condition has been achieved without any additional matching networks and a maximum detector sensitivity of 19500 V/W and a NEP\(_{\text{min}}\) of 0.44 pW/\( \sqrt{\text{Hz}} \) have been
obtained at 200 GHz from simulations, as shown in Fig. 4.5 (b). Since FDAs are resonant antennas, the detector bandwidth increases from 35 GHz to 55 GHz as the number of arms increases from N=3 to N=5.

Figure 4.6. (a) FDA design with N=5 for conjugate impedance matching (both real and imaginary) to an HBD device with 0.1 μm² active device area at 585 GHz. (b) Simulated detector sensitivity and NEP under conjugate impedance matching condition at 585 GHz.

Since lateral scaling of HBDs improves detector’s sensitivity (assuming that the effect of fringing capacitance is negligible), an HBD device with reduced active area of 0.1 μm² has been employed for developing 585 GHz detectors. The HBD with 0.1 μm² active area with has an impedance of ~107–j202 Ω at 585 GHz. An FDA with N=3 has been designed for the 585 GHz detector. The antenna has antenna arm length (l) of 94 μm, arm width (w) of 6 μm, and arm gap (g) of 4 μm. As shown in Fig. 4.6(a), nearly perfect conjugate impedance matching condition has been achieved and a maximum detector sensitivity of ~9,500 V/W and a NEP min of 1.3 pW/√Hz have been obtained at 585 GHz from simulations [113], as shown in Fig. 4.6(b). For many systems, this level of detector
sensitivity would allow pre-amplifiers to be eliminated, greatly simplifying the system design and reducing cost for developing room temperature THz detectors and FPAs.

![Graphs showing impedance and sensitivity](image)

Figure 4.7. (a) FDA design with $N=5$ for conjugate impedance matching (both real and imaginary) to an HBD device with $0.5 \mu m^2$ active device area at 170 GHz. (b) Simulated detector sensitivity and NEP under conjugate impedance matching condition at 170 GHz.

Although we have designed 200 and 585 GHz detectors using HBDs with $0.16 \mu m^2$ and $0.1 \mu m^2$ device active areas, HBDs with larger device active areas are preferred for prototype demonstration due to ease of fabrication. For this purpose, HBDs with $0.7 \times 0.7 \mu m^2$ ($0.5 \mu m^2$) device active area were designed at 170 GHz. These devices have a device impedance of $49-j132 \Omega$ at 170 GHz. The FDA employed for impedance matching with this HBD has same antenna geometry as that is designed for the 200 GHz detectors. Simulation results show that, nearly ideal conjugate impedance matching has been achieved at 170 GHz and peak detector sensitivity of $\sim 3,800$ V/W and a $\text{NEP_{min}}$ of $\sim 1.37$.
pW/√Hz at 170 GHz have been estimated under impedance matching condition, as shown in Fig. 4.7 (a) and (b).

![Diagram of a low-pass filter](image)

**Figure 4.8.** Coplanar stripline stepped impedance low pass filter to pass DC signal and suppress incoming THz signals.

### 4.1.2 Low-pass filter design

In order to extract the rectified DC signals, a stepped impedance coplanar stripline (CPS) low-pass filter was designed and integrated with FDAs which can effectively attenuates the incident THz radiation and permits the DC signals to pass through LPF [115, 116]. Two co-planar strip-line stepped-impedance low-pass filter have been designed for the 200 and 585 GHz HBD detectors to extract the DC signals. The length of each LPF section for 200 GHz attenuation is 135 μm. For the high impedance sections, the strip line width ($t$) = 2 μm and the gap between the lines ($h$) = 82 μm, while for the low impedance sections, $t = 41$ μm and $h = 4$ μm. To effectively attenuate the RF signals, five sections are used. The LPF structure is shown in Fig. 4.8 which is simulated using ADS momentum. Fig. 4.9 (a) shows the simulated $s$- parameters of the 200 GHz LPF and the obtained RF
attenuation at around 200 GHz is as high as -32 dB which sufficient for attenuating 200 GHz signals. In addition, at low frequencies the insertion loss is nearly zero for which the DC signal can be extracted. To attenuate the RF signals of the 585 GHz detector, high impedance section with $h=38 \, \mu m$, $t=2 \, \mu m$ and low impedance section with $h=2 \, \mu m$ and $t=20 \, \mu m$ has been designed. The length of each section is 50 $\mu m$. Fig. 4.9(b) show the simulation results of 585 GHz LPF and RF attenuation of -26 dB at around 585 GHz can be achieved using the designed filter.

![Simulation Results](image)

Figure 4.9. ADS simulation results of the designed CPS LPF for the (a) 200 GHz and (b) 585 GHz HBD detector.
Figure 4.10. CPW structure on the device/detector substrate (GaAs) (a) top view (b) cross-sectional view [117].

4.2 Submicron-scale airbridge finger for integrated HBDs

Although submicron-scale Sb-HBDs have superior intrinsic performance at THz frequencies, THz sensing and imaging systems implemented with such devices using conventional (e.g. flip-chip mounting) technology suffer significant performance deterioration due to additional losses introduced by impedance mismatch and interconnect parasitics (i.e. parasitic finger capacitance ($C_f$) and frequency dependent
spreading resistance \((R_{sp})\). These additional losses can be effectively minimized by developing monolithically integrated Sb-HBDs using submicron-scale airbridges and anode to device mesa spacing to achieve the detector performance close to the intrinsic HBD detectors [117].

Figure 4.11. (a) Parasitic components of airbridge integrated backward diode and (b) equivalent circuit of HBD with its extrinsic parasitic components [117].

In order to evaluate in detail, the impact of these extrinsic parasitic component, a co-planar waveguide (CPW) as shown in Fig. 4.10, was designed (for coupling RF/THz signal to the 0.16 \(\mu m^2\) HBD device) on the device/detector substrate (GaAs) as was simulated using HFSS full-wave simulation where the input port (A-A’ plane) impedance
was set to 50 Ω. The vertical cross section of the corresponding structure (C-C’ in Fig. 4.10(a)) is shown in Fig. 4.10(b) where \( l, t \) and \( h \) are airbridge length, thickness, and height from the GaAs substrate, respectively and B-B’ represents the input port of the HBD device (intrinsic). The thickness \( (t) \) and height \( (h) \) of the airbridge are chosen based on skin depth of bridge metal and epitaxial layer thickness respectively. To include the possible effect of loss from the extrinsic semiconductor device area, a 500 nm thick lossy InAs mesa layer is considered in these simulations.

The extrinsic parasitic components associated with the detector interconnects and layout are shown in Fig. 4.11(a). In this diagram, \( C_{pp}, C_{f1}, C_{f2}, L_f \) and \( Z_{sp} \) are the CPW pad to mesa capacitance, finger (airbridge) to isolation mesa capacitance, finger to anode capacitance, finger inductance, and frequency-dependent spreading impedance (for lossy InAs mesa layer), respectively. The CPW pad to mesa capacitance \( C_{pp} \) varies with the length of airbridge. The length \( (l) \) for the airbridge is chosen so that \( C_{pp(l)}/C_{pp(l=1\mu m)}<0.12 \) to minimize the impact of \( C_{pp} \). However, the self-inductance of airbridge can be increased for an increase in airbridge length and the LC low pass filter introduced by the airbridge may affect the performance of the detector. Therefore, a design tradeoff is needed to choose airbridge length depending on the desired operational frequency of HBD detector.

A simplified equivalent lumped-element circuit model of the entire structure is shown in Fig. 4.11(b), where A-A’ and B-B’ represent the corresponding extrinsic input port (shown in Fig. 4.10(a)) and intrinsic diode port (shown in Fig. 4.9(b)) respectively. The CPW pad to mesa capacitance \( C_{pp} \) is not included since the input port in HFSS simulation is placed at
A-A’ (see Fig. 4.10(a)) plane in this analysis. The total bridge capacitance \( C_f \) consists of both \( C_{f1} \) and \( C_{f2} \). The HBD is represented by its equivalent circuit model inside the dotted box where \( R_s \), \( R_j \) and \( C_j \) are series resistance, junction resistance and junction capacitance respectively. \( R_s \) is comprised of cathode (\( R_c \)) and anode (\( R_a \)) contact resistance.

\[ Z_{\text{sp}} \] in the frequency dependent spreading impedance.

![Figure 4.12](image)

**Figure 4.12.** Equivalent circuit model for parasitic (a) inductance, (b) capacitance and (c) spreading resistance extraction.

In order to extract the circuit parameters in Fig. 4.11 (b), HFSS full-wave simulations were performed by exciting the CPW at the airbridge end (see port 1 in Fig. 4.10(a)). The circuit parameters were then extracted analytically from the simulation. The geometry-dependent inductance (\( L_f \)) and capacitance (\( C_f \)) of the airbridge finger have been extracted by simulating the input impedance of short- and open-circuited CPW (i.e., short and open at plane BB’ in Fig. 4.10(a)). In this simulation, the contact resistances between metal and semiconductor are considered to be zero (ideal ohmic contact), and the anode, cathode and airbridge metals are considered to be lossless (i.e series resistances of metals are neglected) for simplification. The equivalent short and open circuits for \( L_f \) and \( C_f \) extraction are shown in Fig. 4.12(a) and (b).
To further simplify this analysis, the impedance associated with the semiconductor spreading resistance \( R_{sp} \) is neglected by considering lossless InAs mesa layer (i.e. perfect conductor). The bridge inductance and capacitance can then be extracted using:
\[ L_f = \frac{1}{2\pi f Z_{\text{short}}} \]  \hspace{1cm} (4.9)

\[ C_f = -\frac{1}{2\pi f (Z_{\text{open}} - 2\pi f L_f)} \]  \hspace{1cm} (4.10)

where \( Z_{\text{short}} \) and \( Z_{\text{open}} \) are the simulated input impedance of the CPW circuit with airbridge for short and open circuit terminations (at plane BB’), respectively. Using the above approach, the parasitic \( L_f \) and \( C_f \) values for airbridges with thickness \( t=0.5 \, \mu\text{m} \), length \( l=8 \, \mu\text{m} \), height \( h=1 \, \mu\text{m} \) were extracted for varying bridge widths, \( w \), from 0.5 \( \mu\text{m} \) to 5 \( \mu\text{m} \). The results are shown in Fig. 4.13 (a) and (b). The simulation results show that as the finger width increases from 0.5 \( \mu\text{m} \) to 5 \( \mu\text{m} \), the finger inductance decreases from 6.4 pH to 3.1 pH. At the same time, increasing the airbridge width over the same range increases the parasitic capacitance from 0.25 fF to 1.45 fF. The 6.4 pH inductance and the 0.25 fF capacitance overall lead to a LC low pass filter cutoff frequency of approximately 4 THz which is well above the designed detector operation frequency (i.e., 200 and 600 GHz). To put this capacitance in perspective, the intrinsic junction capacitance of a typical submicron-scale HBD with an area of 0.16 \( \mu\text{m}^2 \) is 2.4 fF. So one can see that for finger resistance was extracted from the lumped element circuit model (see Fig. 4.10 (b)) widths below 1 \( \mu\text{m} \), the finger capacitance is much smaller than that of the HBDs, but larger interconnect widths can significantly impact the total capacitance. Therefore, by introducing submicron-scale airbridge fingers, the parasitic capacitance can be made much smaller than the intrinsic device capacitance. This allows the full intrinsic performance potential of these devices to be realized.
### TABLE 4.1
PARAMETERS OF INAS FOR SPREADING RESISTANCE CALCULATION

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dielectric permittivity</td>
<td>12.7</td>
</tr>
<tr>
<td>Doping concentration</td>
<td>1.3×10^9 cm^-3</td>
</tr>
<tr>
<td>Electron mobility</td>
<td>3000 cm^2/V.s</td>
</tr>
<tr>
<td>Bulk conductivity</td>
<td>6.24×10^5 S/m</td>
</tr>
</tbody>
</table>

In addition to interconnect-related parasitics, device layout can also have a significant effect on detector performance. In particular, the frequency dependent spreading resistance arising from the semiconductor current path between the anode metallization and device mesa also contributes to high frequency losses and affects the HBD device performance [118, 119]. The spreading resistance has been extracted by simulating the input impedance of short-circuited CPW (i.e. short at plane BB’ in Fig. 4.10(b)) for lossy InAs mesa. The equivalent circuit model is shown in Fig. 4.12 (c). The effect of $C_f$ is neglected in this analysis. A bulk conductivity ($\sigma$) of 6.24×10^5 S/m is used for the InAs semiconductor current path based on the material parameters shown in table 4.1 [120] using,

$$\sigma = q\mu_e N_d,$$

(4.11)

where $\mu_e$ and $N_d$ are electron mobility and doping concentration respectively. The simulations have been performed for a frequency range from 100 GHz to 600 GHz to calculate the input impedance of short-circuited HBD and the frequency dependent spreading resistance has been extracted using,

$$R_{sp} = Re \left( Z_{in} - i2\pi f L_f \right),$$

(4.12)
where $Z_{in}$ is the simulated impedance of short-circuited HBD with for lossy InAs mesa and $L_f$ is the finger inductance. The capacitance between the cathode contact and InAs mesa has been ignored in this analysis. Since frequency dependent spreading resistance is dependent on the current path between the anode and device mesa, its high frequency losses can be minimized by reducing the anode-mesa lateral spacing $d$ for reduced current crowding and proximity effects [119]. Our simulation results show that by decreasing anode to diode mesa spacing frequency dependent spreading resistance can be minimized and a very low spreading resistance can be achieved with an anode to mesa spacing of 500 nm for a frequency range from 100 GHz to 600 GHz, as shown in Fig. 4.14. Though the spreading resistance can further be minimized by decreasing mesa-anode spacing, however, the detector performance can be degraded for increasing finger to anode capacitance ($C_{f2}$ in Fig. 4.11(a)). Therefore, a design tradeoff is needed to improve detector performance. In addition, the anode to device mesa spacing of 500 nm can be well defined using our EBL process.

Figure 4.14. Extracted frequency dependent spreading resistance for a frequency range from 100 GHz to 600 GHz for a mesa-anode spacing of 500 nm [117].
4.3 Integration and fabrication of HBDs

From the discussion in Section 4.2, one can see that by introducing sub-micron airbridges between the antenna and HBD, the parasitic capacitance could be reduced significantly, and by optimizing the layout of the anode metallization to reduce the anode-to-mesa spacing, the frequency dependent spreading resistance can also be minimized. This allows the full potential performance of the HBDs to be achieved. To experimentally validate these projections, monolithically integrated HBDs and planar FDAs were designed and demonstrated. As demonstrations, antenna-coupled detectors at 200 GHz and 585 GHz as were implemented. A schematic cross section of the integrated HBD detector with its associated circuit components (i.e. antenna, airbridge, output low-pass filter) is shown in Fig. 4.15. Since the HBD is directly integrated at the feed point of the FDA, the integrated detector has significantly reduced circuit footprint compared to competing hybrid-integration approaches. This makes this approach especially suitable for THz FPA implementation.
Figure 4.16. Process flow diagram for the development of submicron-scale airbridge integrated heterostructure backward diode (dimensions are not scaled) [117].
For the demonstrations reported here, the results of the previous analysis were used to set the airbridge geometry and device layout. The cathode contact and antenna are connected using a submicron-scale airbridge finger (w = 0.6 \mu m, t = 0.5 \mu m, l = 8 \mu m), and an anode lateral spacing, d, of 500 nm was used to minimize spreading resistance contributions. In order to extract the output the DC signal, a low-pass filter (LPF) was also included; this was integrated on a 1 \mu m thick polyimide isolation layer. The LPF and antenna are connected using continuous metal interconnects through metal vias in the polyimide.

The fabrication process for the integrated antenna-coupled HBDs includes 1) epitaxial layer growth on semi-insulating GaAs wafer, 2) HBD active area definition, 3) mesa isolation, 4) antenna layer deposition, 5) polyimide layer coating and planarization, 6) air-bridge formation, and 7) LPF and DC pad formation. The process flow is shown pictorially in Fig. 4.16. The lithography steps are a blend of optical lithography and electron-beam direct write; the submicron-scale cathode contacts, mesa, anode and airbridge fingers were defined using electron-beam lithography to ensure accurate dimensional control and to achieve sufficient resolution. In contrast, I-line projection photolithography was used for the antenna, LPF and DC pad development since these elements do not require the precision afforded by electron-beam lithography.

The epitaxial layer structure of Sb-HBDs was grown metamorphically on a semi-insulating GaAs substrate by molecular beam epitaxy (MBE). A self-aligned process was used to fabricate the submicron HBD active area. The top contact cathode layer was
defined using electron-beam lithography in a bilayer MMA/PMMA resist stack. Evaporated metal contacts were deposited by electron-beam evaporation; a Ti/Au/Ti (20nm/460nm/20nm) stack was used. The bottom 20 nm Ti layer increases the adhesion between the metal/semiconductor interface. The contact layer serves as an etch mask for self-aligned wet chemical etching of the InAs and GaSb layers for the active device junction fabrication. Selective etchants were used to etch the InAs and Sb-bearing materials sequentially. Citric acid/H\textsubscript{2}O\textsubscript{2} (1:2) was used to etch the InAs layer, while NH\textsubscript{4}OH/H\textsubscript{2}O (1:5) was used for the Sb-bearing layers. This wet etch produces lateral undercut in the diode mesa, but it should be noted that the formation of the device junction requires a precisely-controlled active area to control junction capacitance and junction resistance. The top Ti of the contact layer minimizes the lateral undercut by minimizing the electrochemical effects [121]. Fig. 4.17(a) shows SEM image of a submicron device after chemical wet etching of the InAs cathode material and Sb-bearing material, and Fig. 4.17(b) shows cross-sectional focused ion beam (FIB) image of the submicron-scale HBD. The image shows that the device active area is reasonably well controlled. More FIB imaging results show that most devices fabricated from this process have symmetric undercut and the undercut varies from 150 nm to 200 nm with an average of 170 nm. Therefore, a large number of nearly identical devices can be fabricated using this process.
In order to isolate the detectors, InAs layer was then wet etched for mesa formation, as shown in Fig. 4.16(c). Fig. 4.18 show the SEM image of HBD device after diode mesa etch. This wet etch results in crystallographic etch profile in the <110> crystal direction. It is preferred that the metal interconnects run directly from the anode to antenna over a mesa sidewall without difficulties with step-edge coverage. Otherwise, broken metal at step edge will result in additional parasitic effects. A three-sided wrap-around Ti/Au (250 nm thick) anode contact, separated from the mesa edge by a spacing of 500 nm was defined using electron-beam lithography and deposited by evaporation, as shown in Fig. 4.16(d). Fig. 4.19 show the SEM image of the U-shaped anode contact wrap around the HBD mesa.
Figure 4.18. SEM image of HBD device after chemical wet etching of InAs isolation mesa etch.

Figure 4.19. Fabricated U-shaped anode structure using with an anode to device mesa spacing of 500 nm to minimize high frequency losses due to frequency dependent spreading resistance.
Figure 4.20. Fabricated 1 μm thick folded dipole antenna (N=5) structure on GaAs substrate to achieve conjugate matching condition for maximum detector sensitivity.

Figure 4.21. Observed interference pattern on antenna layer after etching down the polyimide layer to approximately 1 μm thickness.
For the planar FDAs, a 1 μm thick Ti/Au/Ti/Au (20 nm /1000 nm/20 nm/3 nm) antenna layer was deposited on the GaAs substrate (exposed by the isolation etch) using conventional photolithography and lift-off, as shown in Fig. 4.20. This antenna thickness is chosen to be well above the skin depth of Au at the operational frequency (200 and 585 GHz in this case) and minimizes the high frequency losses. The top Ti layer was used to observe the polyimide thickness using optical microscope. Since the Ti layer can easily be oxidized in ambient atmosphere, the 3 nm Au layer on top of Ti serves as a cap layer to prevent Ti oxidization for improved metal interconnects. Following the antenna fabrication, polyimide layers were spin coated, quasi-planarized by thermal processing, and cured at 250° C in inert gas (N₂) atmosphere. The cured polyimide was uniformly thinned using reactive ion etching (RIE) using a mixture of CF₄ and O₂ until colorful interference patterns are observed on the antenna layer, as shown in Fig. 4.21. The appearance of interference patterns represents approximately 1 μm thick polyimide insulation layer above the antenna, as illustrated in Fig. 4.16(g). Via holes through the polyimide for interconnects between the LPF and bottom-layer metallization were patterned using photolithography. Underdeveloped photoresist was used to achieve a gradual, over-cut profile so that when the via pattern was transferred to the polyimide by O₂/CF₄ RIE, a sloped via sidewall profile would result, as illustrated in Fig. 4.22 (a) and (b). The SEM images of the overcut sidewall profile of polyimide are shown in Fig. 4.23 (a) and (b) for the development of DC pad on GaAs substrate and LPF on polyimide respectively.
Figure 4.22. Schematic of developing overcut profile for metal interconnects: (a) polyimide layer is patterned using under developed photoresist (b) the photolithographic patterns are transferred to the polyimide dielectric layer.

Figure 4.23. SEM image of the overcut polyimide for interconnects between (a) LPF and DC pad and (b) FDA and LPF metallization. Overcut profile for the polyimide pattern was achieved using under developed photoresist.
Figure 4.24. SEM image of the exposed metal contact (cathode contact and antenna) for the airbridge development.

Figure 4.25. SEM image of the HBD detector after submicron-scale airbridge fabrication on polyimide sacrificial layer.
RIE etch of polyimide simultaneously exposed the top of the antenna and cathode contact layers for airbridge and LPF development and the colorful interference patterns on the antennas disappeared immediately once the top of metal layer is exposed. Fig. 4.24 shows the SEM picture of exposed cathode contact prior to airbridge development. The submicron-scale airbridge to the cathode contact was fabricated on cured polyimide layer (Fig. 4.16(i)). The airbridges were defined using electron-beam lithography in a bilayer MMA/PMMA resist stack and a 500 nm thick Ti/Au finger metal was deposited on polyimide sacrificial layer. SEM picture the HBD detector after airbridge fabrication on polyimide is shown in Fig. 4.25. After airbridge fabrication, the 1 µm thick Ti/Au LPF and DC pad on polyimide dielectric layer and GaAs substrate respectively were fabricated using conventional photolithography and lift-off. Finally, the airbridge was released from the sacrificial polyimide layer using isotropic O₂ plasma etch.

Figure 4.26. SEM picture for a submicron-scale HBD successfully integrated to antenna circuits with a submicron-scale airbridge (after release) [117].
Figure 4.27. Optical micrograph of the fabricated 200 GHz single detector with an HBD at the feed point of a FDA with co-planar strip line low-pass filter structure and DC pad.

Figure 4.28. SEM images of a fully-integrated single-element Sb-HBD detector for operation at 200 GHz with its passive components.
Fig. 4.26 shows the SEM picture of a typical HBD integrated with submicron-scale air-bridge finger after airbridge release etch. Optical micrograph of the fabricated 200 GHz single detector with an HBD at the feed point of a FDA with co-planar strip line low-pass filter structure and DC pad is shown in Fig. 4.27. Fig. 4.28 shows the SEM image of the HBD detector with its components for operation at 200 GHz. This process is scalable for fabricating large numbers of identical HBD detectors.

In addition to developing a repeatable fabrication process, TLM patterns have been fabricated on InAs cathode and anode layer of device heterostructures, using the same fabrication process techniques as for the devices for measuring contact resistance. For a typical Ti/Au contact on n+ InAs, specific contact resistances of $1.1 \times 10^{-7} \, \Omega \cdot \text{cm}^2$ and $5 \times 10^{-8} \, \Omega \cdot \text{cm}^2$ were obtained for the cathode and anode contacts respectively from the TLM patterns measurements. Since the TLM patterns on the bottom n+InAs layer (anode contact) was fabricated on newly exposed n+InAs layer, the specific contact resistance of the anode was significantly lower than the contact resistance of cathode.

4.4 Characterization results

4.4.1 DC Characterization

The DC characterization of the submicron HBD detectors with and active areas of 0.35×0.35 $\mu$m$^2$, 0.7×0.7 $\mu$m$^2$ and 3×3 $\mu$m$^2$ were performed in a DC probe station for voltage sweep from -0.1 V to 0.1 V. The typical measured I-V characteristic of the integrated HBD with an active area of 0.35×0.35 $\mu$m$^2$ is shown in Fig. 4.29. The curvature
coefficient, \( \gamma \), which is an area independent parameter and a measure of sensitivity, was calculated from the \( I-V \) characteristics using the expression:

\[
\gamma = \frac{\partial^2 I}{\partial V^2} \frac{\partial I}{\partial V} .
\] (4.13)

For the device characteristic shown in Fig. 4.28, a record-high curvature coefficient of -58 V\(^{-1} \) was measured at zero bias condition [117]. Since the detector sensitivity is directly proportional to the curvature coefficient (\( \gamma \)) of diode, the integrated detectors are expected to exhibit very high sensitivity under conjugate matching condition. For the 0.7×0.7 μm\(^2 \) and 3×3 μm\(^2 \) HBD detectors, a typical zero bias curvature coefficient of -40 V\(^{-1} \) was obtained.

Figure 4.29. Measured DC and curvature coefficient of submicron-scale HBD detector. A record high zero bias curvature coefficient of -58 V\(^{-1} \) is measured for HBD with an active area of 0.35×0.35 um\(^2 \) [117].
4.4.2 Quasi-optical characterization

For quasi-optical characterization, the integrated detector was mounted on a test fixture as shown in Fig. 4.30(a) and (b). The HBD detector chip was connected to a co-planar waveguide (CPW) interconnect submount on a high-resistivity silicon substrate by wirebonding. The detector chip was aligned to the back of a silicon lens (R=5 mm, L=2 mm) for measuring detector sensitivity, NEP, and radiation patterns. No anti-reflection coating was used on the silicon lens.

The schematic of the measurement configuration and measurement setup for sensitivity testing is shown in Fig. 4.31 and 4.32. For this measurement, the incident power was provided through a diagonal horn antenna; input frequencies from 150 GHz to 210 GHz were generated using an amplifier-multiplier chain (AMC) based source. A 9-dB directional coupler was placed in between the output of the multiplier chain and horn.
antenna; the through port of the coupler was connected to an Erickson power meter to monitor the source output power level. The polarizations of the horn antenna and FDA were aligned, with E- and H-plane polarizations in the horizontal and vertical directions, respectively, to couple the incident power into the HBD detector using off-axis parabolic mirrors (17.5 cm focal length). A chopper was used to modulate the incident power, and the output of the HBD detector was measured using a lock-in-amplifier. The input power to the detector was maintained in the range of 1 to 2 $\mu$W in order to operate the detector in the square-law region.

![Schematic of the measurement setup for quasi-optical measurements of HBD detectors](image)

Figure 4.31. Schematic of the measurement setup for quasi-optical measurements of HBD detectors [114].
Figure 4.32. Measurement setup in the laboratory for quasi-optical measurements of HBD detectors (i.e. radiation patterns and sensitivity).

Figure 4.33: Measured response of 3×3 μm² active area HBD at for a frequency range from 190 to 200 GHz.
For a prototype demonstration, Fig. 4.33 shows the measured response of the 3×3 μm² HBD detector for a frequency range from 190 GHz to 210 GHz. The initial sensitivity measurement of the detector was performed by comparing the detector’s response with a VDI Schottky diode’s response (with known sensitivity placed at the same distance from the source) and a sensitivity of ~100 V/W was estimated at 200 GHz. Since the device active area is large, there is substantial mismatch between the antenna and HBD, resulting in lower responsivity. Fig. 4.34 (a) and (b) shows the measured radiation patterns (both E- and H-plane) at 200 GHz. The 3-dB beam widths of the FDA in the E- and H-plane were found to be ~10° and ~6° respectively. The measured E-plane radiation pattern is slightly broader than the pattern in the H-plane, as expected. Ray-tracing simulation results of lens-coupled FDAs in the E- and H-planes radiation patterns are included for comparison; as can be seen, reasonably good agreement has been achieved.
TABLE 4.2

OPTICAL LOSSES IN SENSITIVITY MEASUREMENT

<table>
<thead>
<tr>
<th>LOSSES</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Lens reflection loss</td>
<td>1.6 dB</td>
</tr>
<tr>
<td>Mirror reflection loss</td>
<td>0.5-1 dB</td>
</tr>
<tr>
<td>Coupling loss</td>
<td>1.5 dB</td>
</tr>
<tr>
<td>Misalignment loss</td>
<td>2 dB</td>
</tr>
</tbody>
</table>

Figure 4.35. Measured sensitivity and NEP of zero bias HBD detector in the frequency range from 150 GHz to 190 GHz [114].

The quasi-optical characterization of 0.7×0.7 μm² HBD detector were performed using similar measurement setup as shown in Fig. 4.31. Fig. 4.35 shows the measured sensitivity of the HBD detector for a frequency range from 150 GHz to 190 GHz at room
temperature. The measurement result shows that the HBD detector has a peak measured sensitivity of ~2400 V/W at 170 GHz, with a moderately narrow bandwidth of approximately 13 GHz, after correcting for measurement system losses such as mirror reflectivity, alignment, and measurement system coupling losses. Table 4.2 tabulates each of these loss components. The Gaussian coupling efficiencies of the horn antenna and LC-FDA are ~84% and ~85% respectively [96,122], leading to an aggregate coupling loss within the measurement system of approximated 1.5 dB. The reflection losses of the parabolic mirrors also contribute to incident power loss, which is estimated to be between 0.25 to 0.5 dB/mirror [123]. Finally, misalignment between the FDA and the optical axis of the silicon lens leads to position-dependent loss (e.g. ~3 dB loss for x/R=0.15 [124]). For the precision of assembly used here, we estimate this loss to be ~2 dB. In addition, the lens used did not have an antireflection coating applied. The lens reflection loss is estimated to be approximately 1.6 dB. If this loss is accounted for, the detector sensitivity can be estimated to be approximately 3500 V/W. Thus, the projected sensitivity of the FDA-coupled HBD detector is modestly lower than the expected HBD detector sensitivity of 3,790 V/W obtained from simulation, likely due to additional uncompensated losses in the measurement system (e.g. path loss) and modest discrepancies in the impedance matching conditions. The NEP of the HBD detector was estimated from the measured sensitivity and junction resistance. Fig. 4.3 shows the NEP of the lens-coupled detector for a frequency range from 150 GHz to 190 GHz. The result shows that a minimum NEP$_{\text{min}}$ of approximately 2.14 pW/√Hz (after accounting for system
losses) has been achieved with at 170 GHz and it increases rapidly at the both sides of resonance frequency. If lens reflections are accounted for, the $\text{NEP}_{\text{min}}$ is projected to reach 1.48 pW/$\sqrt{\text{Hz}}$. The results show that the HBD detector has superior sensitivities at resonance frequency.

The far-field co-polarization radiation patterns of the lens-coupled FDA in the E- and H-planes have been measured at 170 GHz. A network analyzer frequency extension module was used to provide the incident power, and the detector was mounted on a rotation stage. The measured radiation patterns for the both E- and H-planes are shown in Fig. 4.36(a) and (b) respectively. The measured radiation patterns show nearly Gaussian-shape main beams in both planes, as expected for a lens-coupled antenna. Ray tracing simulations for the E- and H-planes are included for comparisons. The simulation results show that 3-dB beam widths of 12° and 8° were obtained in the E- and H-planes respectively and reasonably good agreement has been achieved with measurement results.

The measurement results demonstrate that approximate impedance matching was achieved at 170 GHz and the performance of the HBD detector is comparable to or superior to existing room temperature Schottky-based detectors. Since the results are consistent with the modeling results, the results also suggest that monolithically integrated HBD detectors by employing smaller devices can be developed and demonstrated at millimeter-wave and terahertz frequencies (which can be designed on the basis of similar analysis approach) with higher sensitivities and minimum NEPs.
Figure 4.36. Measured and calculated radiation patterns of LC-FDA at 170 GHz in the (a) E-plane (b) H-plane [114].
4.5 Summary

In this chapter, we have presented the design, integration, fabrication and measurement results of submicron-scale airbridge integrated Sb-based HBDs for terahertz sensing and imaging applications. The 200 and 585 GHz detectors are designed based on the analysis of submicron-scale HBDs and planar FDAs. The effect of parasitic capacitance and frequency dependent spreading resistance have been studied to achieve intrinsic performance of HBD detectors. Simulation and device modeling results show that by using submicron-scale airbridges and diode mesa-to-anode spacing, the parasitic effects could be minimized. A novel fabrication process has been developed to fabricate the integrated detector using minimum number of fabrication steps. To demonstrate the effectiveness of the reported integration and fabrication process, the DC and RF characteristics of the fabricated HBD detectors have been measured, evaluated and discussed.
CHAPTER 5:
TERAHERTZ FOCAL PLANE ARRAYS: DESIGN, FABRICATION AND CHARACTERIZATION

In this chapter, the design, fabrication and characterization of two-dimensional (2D) FPAs based on HBD detectors are presented for THz imaging applications. In order to expand the single element design into full 2D THz FPAs, the off-axis radiation patterns of lens-coupled FPAs have been analyzed using the ray tracing techniques. The angular resolution of the FPAs has been calculated for varying element spacing of FPAs. The mutual coupling between the two adjacent FDAs has been studied using full-wave simulations. The designed 2D FPAs have been developed using a similar integration and fabrication process reported in chapter 4. The results have demonstrated the potential to achieve room-temperature, high-performance and large-scale FPAs for THz imaging applications.

5.1 Terahertz imaging focal plane array design

To develop two-dimensional THz FPAs, the “reverse-microscope” optical configuration is utilized, as discussed in chapter 3. Fig 5.1 shows the quasi-optical configuration of a FPA where the array chip is attached to the backside of the imaging lens. Using this imaging configuration, in addition to eliminating the trapped surface wave losses, the cross talk between the adjacent imaging elements can be reduced, resulting in high resolution THz imaging system. In the next few subsections, the analysis of off-axis
radiation patterns, angular resolution, and mutual coupling of the lens coupled-focal plane array (LC-FPAs) are presented.

![Diagram of reverse microscope configuration](image)

Figure 5.1. Reverse microscope configuration to construct THz imaging array proposed by D.B. Rutledge et al.

5.1.1 Off-Axis radiation patterns

In chapter 3, the far-field radiation patterns of the LC-FDAs have been calculated for single element HBD detector on the basis of ray-tracing analysis. In order to evaluate the performance of the 2-D THz FPAs, the far-field off-axis radiation patterns of the LC-FDAs have been studied for a 5×5 FPA using an expanded version of the ray tracing analysis reported in [125]. The geometry of the off-axis ray tracing analysis is shown in Fig 5.2 where the 2D array lay in the x-y plane and z is the center axis. The surface normal is
Figure 5.2. The optics geometry for the off-axis radiation patterns calculation [126].

defined to be,

\[ \hat{n} = \cos \theta \cos \varphi \hat{x} + \cos \theta \sin \varphi \hat{y} + \sin \theta \hat{z}. \]  \hspace{1cm} (5.1)

The ray path inside the lens to be,

\[ \mathbf{v} = (x_s - dx)\hat{x} + (y_s - dy)\hat{y} + (z_s - dz)\hat{z}, \]  \hspace{1cm} (5.2)

\[ \hat{v} = \frac{v}{v'}, \]  \hspace{1cm} (5.3)

where \( x_s, y_s \) and \( z_s \) are the surface coordinates and \( dx, dy \) and \( dz \) are the feed location coordinates. The basis vectors for parallel and perpendicular plane of incidence are
The perpendicular and parallel electric fields can be obtained by

\[ E_{\varphi d} = E_d \cdot \hat{P}_\perp, \]  
\[ E_{\varphi d} = E_d \cdot \hat{P}_\parallel, \]

where \( E_d \) is the electric field inside the lens surface. The field outside the lens surface was derived to be,

\[ E_x = E_{\varphi d} T_\perp (\hat{x} \cdot \hat{P}_\perp) + E_{\theta d} T_\parallel \times [(\hat{x} \cdot \hat{P}_\parallel) \cos(\psi_{tr} - \psi_i) - (\hat{x} \cdot \hat{v}_\parallel) \sin(\psi_{tr} - \psi_i)], \]
\[ E_y = E_{\varphi d} T_\perp (\hat{y} \cdot \hat{P}_\perp) + E_{\theta d} T_\parallel \times [(\hat{y} \cdot \hat{P}_\parallel) \cos(\psi_{tr} - \psi_i) - (\hat{y} \cdot \hat{v}_\parallel) \sin(\psi_{tr} - \psi_i)], \]
\[ E_z = E_{\varphi d} T_\perp (\hat{z} \cdot \hat{P}_\perp) + E_{\theta d} T_\parallel \times [(\hat{z} \cdot \hat{P}_\parallel) \cos(\psi_{tr} - \psi_i) - (\hat{z} \cdot \hat{v}_\parallel) \sin(\psi_{tr} - \psi_i)], \]

\[ E = E_x \hat{x} + E_y \hat{y} + E_z \hat{z}, \]

where \( \psi_i \) and \( \psi_{tr} \) are the angle of incidence and transmission, \( T_\perp \) and \( T_\parallel \) are the perpendicular and parallel transmission coefficients. The propagation path in the air can be found by:
The field components can be obtained by:

\[ s_x = v_x \cos(\psi_{tr} - \psi_i) + (\hat{x} \cdot \hat{P}_||) \sin(\psi_{tr} - \psi_i), \]  

\[ s_y = v_y \cos(\psi_{tr} - \psi_i) + (\hat{y} \cdot \hat{P}_||) \sin(\psi_{tr} - \psi_i), \]  

\[ s_z = v_z \cos(\psi_{tr} - \psi_i) + (\hat{z} \cdot \hat{P}_||) \sin(\psi_{tr} - \psi_i). \]  

The field components can be obtained by:

\[ H = \hat{s} \times E. \]  

Once the field components are calculated, the far-field radiation patterns for the off-axis pixel can be calculated using equations 3.10 to 3.15. The far-field off-axis radiation patterns of the LC-FDAs \((R=6 \text{ mm}, L=2.2 \text{ mm})\) have been studied for 200 GHz and 585 GHz FPAs. Fig. 5.3 (a) and (b) show the calculated off-axis radiation patterns of a 200 GHz 5×5 FPA in the E- and H-planes [4]. The array has an antenna spacing of \(d=0.6\lambda_d\) in the E-plane where \(\lambda_d\) is the guided wavelength, as shown in Fig. 5.3(a). The E-plane radiation pattern of the FPA element has a 3-dB beam width of \(\theta_{3,\text{dB}} \approx 8.2^\circ\) with side-lobe levels less than -15 dB. The angular spacing between the adjacent beams is \(\Delta\theta \approx 8.7^\circ\) with a crossover power level of -3 dB. In the H-plane, as shown in Fig. 5.3(b), the element pattern has \(\theta_{3,\text{dB}} \approx 6^\circ, \Delta\theta \approx 8^\circ\) and a crossover power level of -4 dB for \(d=0.5\lambda_d\). The far-field off-axis radiation patterns of 585 GHz FPA (4×4) have also been studied for varying element spacing. Fig. 5.4 (a) and (b) show the off-axis radiation patterns in the E- and H-planes respectively for \(d=0.6\lambda_d\). The E-plane pattern has a 3-dB beam width of \(\theta_{3,\text{dB}} \approx 4.8^\circ\) and a crossover power level of -3 dB. In the H-plane, the element pattern has \(\theta_{3,\text{dB}} \approx 3.3^\circ\), and a -8 dB crossover power level for the same element spacing.
Figure 5.3. Off-axis simulation results: radiation patterns of a LC-FDA array (5×5) at 200 GHz calculated using ray-tracing technique for element spacing 0.6λₐ and 0.5λₐ in the (a) E-plane and (b) H-plane respectively [96].
Figure 5.4. Off-axis simulation results: radiation patterns of a LC-FDA array (5×5) at 585 GHz calculated using ray-tracing technique for element spacing $0.6\lambda_d$ in the (a) E-plane and (b) H-plane.
5.1.2 Angular resolution

The angular resolution, which is limited by single pixel size (FDA’s dimension) and determined by diffraction-limited main beam, have been calculated at 200 GHz for element spacings varying from $0.6\lambda_d$ to $2.2\lambda_d$ and $0.5\lambda_d$ to $2.0\lambda_d$ in the E- and H-planes respectively. Fig. 5.5 (a) and (b) show that the angular resolution increases linearly with decreasing element spacing in both the E- and H-planes. The calculated highest angular imaging resolution in the E- and H-planes are $8.7^\circ$ and $6.8^\circ$ for element spacings of $0.6\lambda_d$ and $0.4\lambda_d$ (limited by the FDA dimensions) respectively. However, the imaging contrast may become worse with increased angular resolution especially when crossover power level becomes less than 3 dB (see Fig. 5.3 (a) and (b)). Thus, a trade-off between imaging resolution and contrast must be made by selecting appropriate element spacing in the two-dimensional THz FPA.

5.1.3 Mutual coupling of array elements

As discussed in the previous subsection, by decreasing element spacing between the adjacent antennas in the FPA, higher imaging resolution could be achieved. However, the mutual coupling between two adjacent antennas, which quantify the level of cross-talks between the adjacent antennas, cannot be ignored. The mutual coupling is typically undesirable because received power by one antenna is affected by a nearby antenna. As a result, the coupling effect significantly degrades the antenna efficiency and radiation patterns.
Figure 5.5. Angular resolution of the LC-FPA array varying with element spacing along (a) E-plane and (b) H-plane at 200 GHz [96].
The effects of mutual coupling between the neighboring antennas have been studied from using HFSS for various antenna spacing \((d)\) in the E- and H- planes at 200 GHz, as shown in Fig. 5.6 (a) and (b). The two identical FDAs were designed on semi-infinite GaAs substrate and their mutual coupling were estimated from the simulated s-parameters. A mutual coupling of \(-28\) dB (S21 of a two-antenna array simulated using HFSS, see Fig. 5.6 insets) has been obtained for \(d=0.52\ \lambda_d\) (limit is 0.50 \(\lambda_d\) due to the antenna dimension) in the E-plane array (see inset of Fig. 6(a)). For \(d\) larger than 0.52 \(\lambda_d\), the mutual coupling decreases quickly with increasing element spacing. The same trend has also been obtained in the H-plane with a mutual coupling of \(~-20\) dB for \(d = 0.35\ \lambda_d\). The mutual coupling of the antenna array at 585 GHz has also been investigated and similar trend was observed. As shown in Fig 5.7 (a) and (b), a mutual coupling of \(-30\) dB has been obtained for \(d=0.5\ \lambda_d\) in the E plane and \(-22\) dB is obtained for \(d=0.25\ \lambda_d\) in the H-plane. We consider mutual coupling of less than \(-20\) dB (i.e. coupling coefficient less than 0.01) for the FPA design which is good enough for many practical applications.

5.2 Fabrication of THz focal plane array

To fabricate the designed FPAs, we have used similar integration and fabrication process discussed in chapter 4. The FPAs were fabricated for element spacings of \(0.6\lambda_d, 0.8\lambda_d, 1.0\lambda_d,\) and \(1.2\lambda_d\) in both the E-and H- planes. The cathode contact, diode junction, mesa isolation, anode layer and antenna structure were developed using the same fabrication steps as discussed in chapter 4. After the antenna development, the polyimide
Figure 5.6. Simulated mutual coupling of two folded dipole antennas along (a) E-plane and (b) H-plane at 200 GHz.
Figure 5.7. Simulated mutual coupling of two folded dipole antennas along (a) E-plane and (b) H-plane at 585 GHz.
layer was spin coated, quasi-planarized, cured, and etched in RIE. The submicron-scale airbridge contacts, LPF, and DC pad were also developed using same lithography steps as discussed in chapter 4. Fig 5.8 and 5.9 show the fabricated 200 GHz and 585 GHz focal-plane array (6×2) for element spacing $1.2\lambda_d$ and $1.0\lambda_d$ in the E- and H-planes respectively. Fig. 5.10 shows fabricated 585 GHz focal-plane array (6×4) for element spacing $1.2\lambda_d$ and $1.0\lambda_d$ in the E- and H-planes respectively. After fabrication, the FPAs were diced for DC and radiation patterns measurements.

Figure 5.8. Fabricated 200 GHz focal-plane array (6×2) for element spacing $1.2\lambda_d$ and $1.0\lambda_d$ in the E- and H-planes respectively.
Figure 5.9. Fabricated 585 GHz focal-plane array (6×2) for element spacing $1.2\lambda_d$ and $1.0\lambda_d$ in the E- and H-planes respectively.

Figure 5.10. Fabricated 585 GHz focal-plane array (6×4) for element spacing $1.2\lambda_d$ and $1.0\lambda_d$ in the E- and H-planes respectively.
5.3 Initial characterization of THz FPAs

Fig. 5.11 shows the I-V characteristics of four HBDs (0.7μm×0.7μm active area) in a 200 GHz 2×2 FPA. All FPA elements exhibit approximately identical I-V characteristics (and curvature coefficient) demonstrating that a repeatable and successful fabrication process has been developed. Since the sensitivity of FPA elements is directly proportional to the curvature coefficient of HBDs at zero bias, all FPA elements are expected to exhibit similar sensitivity.

To evaluate the array performance, the 2×2 detector was first bonded to a silicon substrate using glue, where an ultrasonic wire-bonding tool was used to electrically connect both the grounds and center conductors of CPW pad with the DC contact pad of HBD detector. The array was mounted to a quasi-optical block for radiation pattern measurements. Two SMA connectors are utilized to extract the output DC signals. Fig. 5.12 shows the measurement setup for radiation patterns measurements. To measure the off-axis radiation patterns of the array, DC signals from the two neighboring elements were measured. A VDI 190-210 GHz solid-state source was employed to provide THz incident power. The responses of the two elements in the E- and H-planes are normalized and plotted, as shown in Fig. 5.13(a) and (b), where two distinct peaks were observed in the both planes.
Figure 5.11. I-V characteristics of a 200 GHz FPA (2×2) elements. All FPA elements exhibit approximately identical characteristics [121].

Figure 5.12. Measurement setup for the quasi-optical characterization of a 2×2 array.
Figure 5.13. Normalized radiation patterns of the 2×2 FPA at 200 GHz (a) E-plane (b) H-plane.

5.4 Summary

In this chapter, the design, fabrication and characterization 2D focal plane arrays at 200 GHz and 585 GHz for THz imaging applications are described. In order to expand the single element design into full 2-D THz FPAs, the off-axis radiation patterns have been analyzed using the ray tracing technique. The angular resolution of the FPA has been calculated for varying element spacing. The mutual coupling between two adjacent FDAs has been studied using full-wave simulation. The designed 2D FPAs have been fabricated using a similar integration process reported in chapter 4. The results have demonstrated the potential to achieve room-temperature, high-performance and large-scale FPAs for THz imaging applications.
6.1 Summary of this report

In this report, design, simulation, fabrication, and characterization results are presented for developing advanced THz detectors and FPAs based on HBDs. To accurately design HBD based THz detectors and FPAs, the performance of Sb-HBDs was evaluated based on their nonlinear lumped element circuit model. The simulation results show that a maximum detector sensitivity of 21,000 V/W at 200 GHz and 9,500 V/W at 585 GHz, respectively, can be potentially achieved using 0.16 μm², and 0.1 μm² active area HBDs respectively. The impedances for submicron-scale HBDs were studied to achieve conjugate impedance matching condition for maximum detector sensitivity. In addition, HBDs’ NEP, dynamic range, and cutoff frequencies were studied to assess their performance at THz frequencies.

In order to develop THz HBD detectors and FPAs, lens-coupled folded dipole antennas (LC-FDAs) are proposed to maximize antenna directivity and minimize surface wave losses. The performance of the LC-FDAs was analyzed on the basis of even-odd mode analysis (EOA). The embedding impedances of FDAs were studied for different antenna geometries; the results show that a wide range of impedance values (for both real and imaginary parts) can be obtained for impedance matching with HBDs. The LC-FDA’s far-field radiation patterns, antenna directivity and Gaussian coupling efficiency for
different lens structures were studied using ray-tracing techniques and a good agreement between the calculation and experimental results has been obtained.

On the basis of analysis results, FDAs were designed for conjugate impedance matching with HBDs at 200 GHz, and 585 GHz operational frequencies and maximum detector responsivities of 21,000 V/W and 9,500 V/W were obtained respectively. In order to develop integrated HBD detectors and FPAs, a unique fabrication technique was developed using submicron-scale airbridges. The developed process offers the potential to achieve high accuracy and reproducibility for fabricating integrated THz detectors and FPAs. For prototype demonstration, an HBD with 0.5μm² device was used. To demonstrate the effectiveness of the reported integration and fabrication process, DC and RF characterizations were performed. The measurements show that a peak detector sensitivity of approximately 2400 V/W and a NEP_min of 2.14 pW/√Hz were obtained at 170 GHz without applying an anti-reflection coating on the silicon lens. If an antireflection coating was used, a sensitivity of approximately 3500 V/W and a NEP_min of 1.48 pW/√Hz are projected. The radiation patterns of the quasi-optical detector in both the E- and H-planes were measured and good agreement was achieved between the simulation and measurement. The performance of this detector can be further improved by scaling the HBD device active area.

In order to expand the single element design into full 2D THz FPAs, the off-axis radiation patterns of the FDA mounted on an extended hemispherical silicon lens have been analyzed using the ray tracing technique. The angular resolution and mutual
coupling of FPAs were studied for varying element spacing. The designed 2D FPAs were developed using a similar fabrication process to that reported in Chapter 4. The results have demonstrated the potential to achieve room-temperature, high-performance and large-scale FPAs for THz imaging applications.

A summary of high performance HBD detectors and FPAs developed by Notre Dame (narrow band, lens-antenna coupled and integrated diodes), HRL (waveguide coupled, flip-chip mounted) and Traycer (broadband, lens-antenna coupled and integrated diodes) are summarized in Fig. 6.1. A further comparison of the performance of HBD-based THz detectors/FPAs with other unamplified room-temperature detectors/FPAs is presented in Table 6.1. As can be seen, the HBD offers compelling performance metrics compared to the competing device alternatives.

![Graph comparing HBD responsivity](image)

Figure 6.1 Comparison of experimentally-measured HBD responsivity reported by several groups (points) and simulated (solid line) maximum detector responsivity of HBD detectors.
### Table 6.1

**COMPARISON OF UNAMPLIFIED ROOM TEMPERATURE THZ FPA TECHNOLOGIES**

<table>
<thead>
<tr>
<th>Detector Technology</th>
<th>Detector type</th>
<th>Band</th>
<th>Sensitivity (V/W) (theory)</th>
<th>Sensitivity (V/W) (measured)</th>
<th>NEP (pW/Hz)</th>
<th>NEΔT</th>
<th>Ref</th>
</tr>
</thead>
<tbody>
<tr>
<td>HBD (0.1 μm²)</td>
<td>CPW feed</td>
<td>0.07-0.11 THz</td>
<td>50,000 at 0.094 THz</td>
<td>50,000 at 0.094 THz</td>
<td>0.18</td>
<td>0.5K 15 fps†</td>
<td>[58]</td>
</tr>
<tr>
<td>Schottky Diode</td>
<td>LC-Sinusous</td>
<td>0.15-0.45 THz</td>
<td>-</td>
<td>700 at 0.3 THz</td>
<td>1.5</td>
<td>-</td>
<td>[126]</td>
</tr>
<tr>
<td>MOSFET</td>
<td>LC-Dipole</td>
<td>0.75-1 THz</td>
<td>-</td>
<td>1100 at 0.9 THz</td>
<td>50</td>
<td>-</td>
<td>[69]</td>
</tr>
<tr>
<td>Graphene-FET</td>
<td>LC-Bow-tie</td>
<td>0.4 THz</td>
<td>-</td>
<td>74 at 0.4 THz</td>
<td>130</td>
<td>-</td>
<td>[127]</td>
</tr>
<tr>
<td>Bolometer (RT)</td>
<td>LC-Spiral</td>
<td>0.1-2 THz</td>
<td>-</td>
<td>70 at 0.4 THz</td>
<td>50</td>
<td>-</td>
<td>[128]</td>
</tr>
<tr>
<td></td>
<td>LC-Bow-tie</td>
<td>0.08-0.12 THz</td>
<td>-</td>
<td>400 at 0.1 THz</td>
<td>69</td>
<td>-</td>
<td>[129]</td>
</tr>
<tr>
<td>EMSM detector</td>
<td>0.4 THz</td>
<td>2169 at 0.4 THz</td>
<td>14.9</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>[130]</td>
</tr>
</tbody>
</table>

† Theory

### 6.2 Proposed future work

A prototype demonstration of 170 GHz lens-coupled HBD detector has been carried out in this research, but much additional work to developing THz imaging systems could be pursued. Demonstration of 2-D imaging arrays is the next step following demonstration of a single antenna-detector pixel with promising performance based on the analysis results presented in Chapter 5. Fig. 6.1 shows a 4×4 FDA 2D array for a prototype demonstration of the THz imaging system. In this imaging system, all 16 detectors share one ground, and the DC signals will be processed by an A/D converter and digital circuits before displaying a reconstructed image on a screen or monitor. In addition to narrowband HBD detectors, broadband antennas can be employed for developing THz...
imaging FPAs. However, broadband antenna designs are typically challenging to realize and require impedance matching networks [131].

Different array architectures can be employed to develop large-scale imaging arrays [132]. For example, these arrays can be developed based on a spider web array architecture to contain more pixels and maximize imaging resolution. Fabrication of larger format FPAs are challenging because a larger number of identical HBD detectors need to be fabricated with very high device yield and better device/antenna uniformity. One way to develop large format FPAs is to use spider web array architecture where small arrays (15 elements) can be fabricated separately and large format arrays can be realized by joining them together using quilt packaging technique, as shown in Fig 6.2. Fig 6.3 shows the overall diagram of the proposed room temperature, portable THz camera consisting of HBD detector array (either brick or spider), IC chips and LCD screen, as shown in Fig 6.4. HBD based THz imaging systems are expected to be light-weight, low-cost and portable and affordable to people inside and outside of the scientific/engineering community.

Figure 6.2. A prototype room temperature THz imaging system consist of a FDA array (4x4, brick spacing) and Sb-HBDs.
Figure 6.3. Spider web array architecture for HBD based room temperature THz camera [132].

Figure 6.4. The overall diagram of the proposed room temperature, portable THz camera.


APPENDIX A:

RECIPE FOR INTEGRATED HBD DETECTORS AND FPAS FABRICATION

1) Alignment mark and top contact metal deposition
   i. Surface Treatment
      a. Descum (PVA) 5 min (pressure 300 mTorr, power 600 mW O2 flow 80 sccm)
      b. Acetone clean 5min
      c. IPA clean 5min
      d. IPA rinse and N₂ blow dry
      e. HCl clean (20 sec) and DI rinse
      g. Dehydration bake 180C, 2min
   ii. Electron beam lithography
      a. MMA 1000 nm, spin speed: 4000 rpm
      b. PMMA 150 nm, spin speed: 4000 rpm
      c. EBL Dose 500 μC/cm²
      d. Development in MIBK:IPA:MEK (1.5 : 3 : 1) for 40 sec, Rinse in IPA for 30 sec
      e. Descum in PVA 1 min
   iii. Contact metal deposition
      a. Pre-deposition surface treatment 1:2 diluted HCl -15 sec
      b. Metal deposition: FC1800#1, 20 nm Ti/ 460 nm Au/ 20 nm Ti,
         Deposition rate: Ti-1 A/s/ Au-3 A/s/ Ti-1 A/s
      iv. Lift-off: warm acetone 15 min, eyedropper pipe rinse

2) Diode mesa etching
   i. Solvent clean and dehydration bake
      a. Acetone clean 5min
      b. IPA clean 5min
      c. Dehydration bake 110C 5 min
   ii. Photolithography
      a. AZ5214E spin speed: 3000 rpm
      b. Soft bake: 110C 30sec
      c. Primary exposure: 0.5 s
      d. Develop time: 35s AZ300 DI rinse
      e. Native oxide removal- 1:2 diluted HCl -10 sec, DI rinse
   iii. Cathode etching: Citric acid (CA):H₂O₂ 1:2 Etch Time: 80 sec DI rinse
   iv. Anode etching: NH₄OH:H₂O 1:5 Etch Time: 120 sec DI rinse
3. Isolation mesa etching
   i. Solvent clean and dehydration bake
   iii. Electron beam lithography
       a. PMMA 500 nm
       b. EBL Dose 500 $\mu$C/cm$^2$
       c. Development in MEK:IPA:MIBK for 35 sec, rinse in IPA for 30 sec
   iv. Photolithography and RIE etch
       a. AZ5214E 3000 rpm
       b. Soft bake: 110C 30sec
       c. Primary exposure: 0.5 s,
       d. Develop time: 35s AZ300
       e. RIE Etch of PMMA (50 mW Pressure 200 mTorr O2 Flow 32 sccm)
           Etch time 7 min
       f. Flood Exposure 16 sec (Karl Suss)
       g. Develop time: 35s AZ300
       h. Post development bake to improve resist adhesion (110 C for 20 sec)
   v. Mesa etching:
       a. CA:H2O2 1:2 240 sec;
           DI rinse

3) Anode contact metal deposition
   i. Solvent clean and dehydration bake
   ii. Electron beam lithography
       a. MMA 1300 nm
       b. PMMA 150nm
       c. EBL Dose 500 $\mu$C/cm$^2$
       d. Development in MEK:IPA:MIBK for 40 sec, rinse in IPA for 30 sec
       e. Descum (PVA) 1 min
   iii. Contact metal deposition
       a. Pre-deposition surface treatment 1:2 diluted HCl -15 sec
       b. Metal deposition: FC1800#1, 20 nm Ti/ 230 nm Au
           Deposition rate:: Ti-1 A/s  Au-3 A/s
   iv. Lift-off: warm acetone 15 min, eyedropper pipe rinse

4) Antenna layer fabrication
   i. Solvent clean and dehydration bake
   ii. Photolithography
       a. AZ5214E spin speed: 2000 rpm
       b. Soft bake: 110C 30sec
       c. Primary exposure: 0.1 sec,
       d. Post bake: 110 C 40 sec
e. Flood exposure: 16 sec  

f. Develop time: 35s AZ300  

g. Descum (PVA): 3 min  

iii. Antenna metal deposition  

a. Pre-deposition surface treatment 1:2 diluted HCl -15 sec  

b. Metal deposition: FC1800#1, 20 nm Ti/1100 nm Au/20 nm Ti/2 nm Au  

   Deposition rate: Ti-1 A/s, Au-3 A/s/Ti 1 A/s/Au 0.5 A/s  

iv. Lift-off : warm acetone 15 min, squeeze bottle rinse or eyedropper pipe rinse  

6) Polyimide coating and planarization  

i. Solvent clean and dehydration bake  

ii. (optional) Adhesion promoter (AP3000) 5000rpm 30sec  

   Polyimide 2610 spin speed: 5000rpm  

iii. Multiple coating, Intermediate bake 180C 60 sec  

iv. Hard Cure: 240C for 1 hour in Lindberg Blue Furnace under inert N2 flow condition. Load/unload temp 150 C. Temp ramp 10 C/min  

7) Polyimide blanket etching  

   RIE recipe for polyimide etch  

   Gas flow: 32sccmO2 / 8 sccmCF4,  

   Pressure: 200mT,  

   DC bias: 60V,  

   Power: 50W  

   etch rate: 200 nm/min.  

   Inspect the sample under optical microscope and watch carefully. Proceed the remaining time until colorful interferences appear on antenna for a polyimide thickness of approximately 1000 nm.  

8) Polyimide patterning for via hole  

i. Solvent clean and dehydration bake;  

ii. Photolithography and RIE etch  

   a. AZ5214E spin speed: 2000 rpm  

   b. Soft bake: 110C 30sec  

   c. Primary exposure: 0.5 s,  

   d. Develop time: 25s AZ300 (under develop)  

iii. Polyimide via etching(RIE): 4 min  

   (Inspect with optical microscope and watch carefully and proceed the remaining time until colorful interference patterns disappear on antenna. Finally inspect in SEM)  

9) Airbridge fabrication  

i. Solvent clean and dehydration bake  

ii. Electron beam lithography  

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a. MMA 1000 nm  
b. PMMA 150 nm  
c. EBL Dose 500 $\mu$C/cm$^2$  
d. Development in MEK:IPA:MIBK for 40 sec, rinse in IPA for 30 sec  
e. Descum (PVA) 1 min  

iii. Metal deposition  
a. Pre-deposition surface treatment 1:2 diluted HCl - 15 sec  
b. Metal deposition: FC1800#1, 20 nm Ti/ 500 nm Au,  
   Deposition rate: Ti - 1 A/s/ Au - 3 A/s  

iv. Lift-off : warm acetone 15 min, squeeze bottle rinse or eyedropper pipe rinse  

11) Residual polyimide removal and cleaning substrate for filter and DC pad fabrication  
i. Solvent Clean and dehydration bake;  
ii. Photolithography  
   a. AZ5214E spin speed: 2000 rpm  
   b. Primary exposure: 0.5 s,  
   c. Develop time: 25s AZ300 (under develop)  
iii. O2 plasma etching in RIE: 15 min  
   Gas flow: 32sccm O2  
   Pressure: 200mT  
   DC bias: 60V  
   Power: 50W  

10) Airbridge release from polyimide  
i. Solvent clean and dehydration bake;  
ii. Photolithography  
   a. AZ5214E spin speed: 2000 rpm  
   b. Primary exposure: 0.5 s,  
   c. Develop time: 45s AZ300  
iii. O2 Plasma etching in PVA:  min  
   pressure 200 mTorr,  
   power 600 W  
   O2 flow 80 sccm  

Inspect the sample under SEM until airbridges fully released from polyimide.  

12) Filter and DC pad Fabrication  
i. Solvent clean and dehydration bake  
ii. Photolithography  
iii. 20 nm substrate (GaAs) etch  
v. Photolithography  
vi. Metal deposition
a. Pre-deposition surface treatment 1:2 diluted HCl - 15 sec
b. Metal deposition: FC1800#1, 20nm Ti/ 800 nm Au
   Deposition rate: Ti-1 A/s Au-3 A/s
vii. Lift-off: warm acetone 15 min, squeeze bottle rinse or eyedropper pipe rinse

13) Detector dicing and mounting
   i. Spin coat the sample with unbaked photoresist before dicing
      Resist spin speed: 2000 rpm
      Cutting speed 2 mm/sec
      Spindle speed 18000 rpm
   ii. Mount the diced HBD detectors on high resistivity silicon using glue
   iii. Connect the DC pad and CPW line using wire bonding tool (gold wire)
APPENDIX B:

THERMAL STABILITY OF POLYIMIDE AND HETEROSTRUCTURE BACKWARD DIODES

Although submicron-scale HBD detectors were demonstrated using 0.7×0.7 μm² active device area HBDs, it was very challenging to develop HBDs with smaller device with great performance. It was found that the polyimide layer became slightly unstable during the process of airbridge and LPF development that resulted in device failure. Fig B.1 (a) and (b) show the FIB cross-sectional image of HBD device before airbridge and after LPF development respectively. The results show that the polyimide sacrificial layer was affected by the thermal step of lithography process for airbridge and LPF development and the thermal instability of polyimide affects airbridge interconnect. Fig. B.1. (c) shows the effect of thermal step of photolithography on the performance of HBDs. The result shows that the device characteristics degrade after LPF development due to increased interconnect resistance. Although micron scale devices can survive against this instability (detectors larger than 0.5 μm²), it is very hard to fabricate submicron-scale HBDs with 0.4×0.4 μm² and 0.3×0.3 μm² areas.
Figure B.1. Cross-sectional view of HBD after before (a) airbridge and (b) after LPF development. (c) I-V characteristics of HBD detector before and after LPF development. The result show that HBDs are affected by thermally unstable polyimide.
Figure B.2. Effect of thermal processing on the performance of HBDs. HBDs are affected by thermal curing at temperatures beyond 250°C.

In order to solve this problem, the polyimide layer needs to be fully cured at higher temperatures for developing airbridges and LPFs. However, it was found that HBDs are not stable at higher temperatures. Thermal stability of a typical HBD with an active area of 0.5 μm² was studied at three different temperatures: 240°C, 250°C and 270°C. The dc characterization results show that submicron-scale HBDs are stable for temperatures below 250°C and their performance degrades at temperatures beyond 250°C. Fig B.2 shows the I-V characteristics of the HBD before and after thermal processing where the HBD was kept in the furnace for 60 and 90 minutes at 270°C. The result shows that the device characteristic degrades with curing time at elevated temperature. This was caused by migration of Ti metal of contact layer through a solid-state diffusion to the tunnel junction assisted by higher temperatures. To solve this problem, the polyimide layer has
to be cured at lower temperature (i.e. 200°C-220°C) and a variable frequency microwave curing could be used for this purpose. This curing temperature is safe for HBDs. Alternatively, titanium-tungsten diffusion barrier layer can be used in the contact layer to prevent Ti diffusion into the tunnel junction. Using this contact scheme, thermal stability of ohmic contact up to 450°C can be achieved and polyimide could be fully cured at elevated temperatures. The updated process will improve the performance and yield rate for submicron-scale HBD detectors significantly.
APPENDIX C:

MECHANICAL DRAWING OF QUASI-OPTICAL BLOCK

Figure C.1. Back end of the quasi-optical block (a) front view (b) back view
Figure C.2. Front end of the quasi-optical block (a) back view (b) front view (c) top view (d) side view.