Integrated Broadband Antennas for Photonic-enabled Millimetre-wave Transmission

by

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Universidad Carlos III de Madrid

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Abstract

The scope of this doctoral thesis is within the development of advanced wireless communication links operating in the millimetre-wave frequency range, contributing in the development of key requirements, on one hand enhancing the radiated power and enabling the steering of the radiated power. This thesis presents a photonic-based approach to support high data rates and meet the requirements that are expected for next generation systems, defined for beyond 5G (also known as 6G), highly mobile steerable devices.

We present advances in the design and optimization of planar antennas on low-cost laminates integrating high-speed photodiodes in hybrid fashion. We present technique for opto-electronic integration with purpose of maximising the power transfer from photodiodes to the antenna. Both the utilise uni-travelling carrier photodiode (UTC-PD) and p-i-n photodiode (PIN-PD) are utilised for this purpose. With the successful demonstration of single-element UTC-PD and PIN-PD emitters, we conduct further research to realise integrated emitter arrays. In this respect, a photonic-integrated emitter with planar phased array antennas is developed using UTC-PDs, providing radiated mm-Wave power of up to -2 dBm. We experimentally demonstrate the photonics-enabled beamsteering over a frequency range of 60 GHz to 70 GHz, with a beam scanning range of 30°. Moreover, using the phased array module, the wireless data transmission is also achieved together while performing beamsteering. As a result, this is the first of its kind system demonstration employing integrated photonic emitter, forming the basis for large-scale integrated antenna arrays for 5G systems.

Additionally, we also demonstrate photonics-based terahertz source working in the J-band (220 – 325 GHz) frequency. Two modules are developed by integrating a high-speed InGaAs/InGaAsP PIN photodiode in a WR-03 rectangular waveguide WR-3 package for photonic to terahertz conversion. The first device performs direct photonic to THz conversion, providing an output power of -14.4 dBm (36.3 μ W). The second device, is based on a hybrid photonic-electronic technique. By photonic-electronic co-integration, the PIN-PD is coupled to a frequency multiplier chip in a single package, providing the photonic generation and frequency multiplication in the same device. To the best of our knowledge, this is the first realisation of WR-03-integrated emitter based on a PIN photodiode.

Finally, thesis presents the design of ultra-wideband detectors based on Schottky barrier diodes assembled in quasi-optic package. The primary goal is to propose a design for the receivers with large baseband bandwidth, capable of operating across frequency bands in mm-Wave and THz range. A record baseband of 25 GHz is achieved using custom-built detectors. For the proof of concept, experiments are conducted using the custom-built devices capable of transmitting more than 12.5 Gbps of data signals on carrier from mm-Wave to sub-THz.

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CHAPTER 1

Introduction

1.1 Introduction

Communication systems have been evolving over the past decades to meet the data traffic ever-increasing demand providing higher data rates, pushing the limits of the wired and wireless segments of communication networks. This demand has been fuelled by the increasing number of hand-held devices and smart applications, currently projecting an overall mobile traffic rate of 77 exabytes/month by the year 2022 [1], which amounts to a 70% annual growth. The studies suggest that the primary contribution to this year-on-year growth comes from the personal wireless devices, with total number expected to reach 8.7 billion by 2023 [2]. These trends suggest that wireless data will continue to retain the major share of backbone data traffic. The evolution of the wireless systems is highlighted in Figure 1-1, showing a 7-fold increase over a decade. In order to satisfy the current market needs and accommodate the emerging applications such as

Figure 1-1: Evolution trend of data rates for wireless communications [3].

ultra-high definition (UHD) video streaming, autonomous cars and machine learning, wireless systems require significantly new developments. In this respect, two main approaches have been explored; increasing the spectral efficiency (higher order modulation formats) and use of higher carrier frequencies [4]. While the former technique has its technological and practical limitations, the latter has been considered as a viable solution which leads to the need making new spectra available, opening new frequency regions of the spectrum (in the millimetre- and Terahertz-wave range) enabling to reach high data rates without increasing the modulation complexity.

1.1.1 Millimetre and Terahertz Spectra

Millimetre wave (mm-Wave, 30 – 100 GHz) and terahertz (THz, 100 GHz – 10 THz) frequencies have been traditionally used in niche applications such as security imaging and spectroscopy, or high resolution radar. With vast amounts of available bandwidth and large portions of these bands being still unregulated, provide potential over the conventional microwave spectrum to develop high capacity wireless systems. The mm-Wave and THz bands lie in the region between the microwave and infrared waves of the electromagnetic radiation spectrum, as can be seen in Figure 1-2.

Historically, this region of the spectrum has been also referred to as the “THz Gap”, due to its inaccessibility by the lack of efficient and powerful signal sources. This is aggravated by the fact that mm-Wave and THz frequencies experience higher losses in radiated signal, making the deployment of THz communication systems a challenge. The

Figure 1-2: The millimetre- and terahertz-wave in electromagnetic radiation spectrum.

attenuation coefficient is significantly high for frequencies above 30 GHz than at microwave bands, due primarily to free-space path loss (FSPL) and the atmospheric attenuation.

1.1.2 Free-space Path Loss

In a wireless signal transmission link, the free-space path loss (FSPL) is the difference between the emitted and the received power due to the spread of electromagnetic energy over the space as it propagates. The attenuation is quadratically proportional to the link distance d , hence reducing its power density. This propagation loss model can only be valid for a line-of-sight link with fixed antennas. The relation between the transmitted signal power P_{TX} and average received power P_{RX} can be obtained using a Friis formula [5], [6],

$$P_{RX}(dBm) = P_{TX}(dBm) + G_{TX}(dB) + G_{RX}(dB) + 20\log_{10}\left(\frac{4\pi df}{c}\right) \quad (1-1)$$

where G_{TX} and G_{RX} are the gain of the antennas on transmitter and receiving side respectively, f is the signal frequency and c is the speed of light. The Equation 1-1 is only valid in the far-field of the antenna when d is beyond Fraunhofer region [7].

It suggests that the propagation loss is directly proportional to the transmitter signal frequency. The calculated FSPL at various frequencies is shown in Figure 1-3. As can be seen, for a given distance there is a significant amount of power loss when moving from the microwave to THz band.

Figure 1-3: Free-space path loss (FSPL) as a function of wireless transmission distance at four different carrier frequencies for isotropic antennas.

1.1.3 Atmospheric Loss

The atmospheric losses occur due to the absorption of the electromagnetic radiation by the gas molecules and water vapour in the propagation media, the atmosphere. When the signal frequency matches with the resonance frequency of a given molecule species present in the atmosphere, it is absorbed by them.

In addition, we must also take into account that as the frequency increases, the decreasing wavelength of propagating wave matches with the size of water droplets, causing additional increase of the loss due to scattering. For the outdoor wireless communications, it is important to quantify these losses for different conditions to establish reliable links. Figure 1-4 shows the propagation loss for three different atmospheric conditions. These calculations are based on ITU Recommendation ITU-R P.676-12 [8] and ITU-R P.838-3 [9]. Some high intermittent peaks in the curves are the result of absorption by oxygen and water molecules. Some windows with relatively low loss are also shown, covered by standard waveguide frequency bands.

Figure 1-4: Propagation losses for electromagnetic wave over the frequency range of 1 GHz to 600 GHz due to the atmospheric attenuation. The conditions assumed correspond to sea-level as per ITU Recommendations; air pressure of 1013 hPa, temperature of 20°C and water vapour of 0 g/m³ for dry air, 7.5 g/m³ for clear sky and 50 g/m³ for heavy rain [8], [9].

Currently, the mm-Wave frequencies below 60 GHz, with contiguous bandwidth of 3 GHz, are being extensively studied for the 5G New Radio (5G-NR-RF2; n257, n258 and n260) to be use in next generation cellular systems [10], with commercial deployment expect in coming years. The research work in this thesis covers wide range of frequencies, concerning the developments on the sources, detectors and system demonstrations in the E-band (60 – 90 GHz), D-band (110 – 170 GHz) and J-band (220 – 325 GHz).

1.2 Wireless Communications

1.2.1 Photonic mm-Wave/THz Generation

The mm-Wave and THz frequencies have become attractive bands for establishing the high-capacity terrestrial wireless deployment by the recent technological advances [11]. The most promising applications are aimed at providing the wireless back-haul/front-haul for the small-cell networks by implementing a wireless bridge, mm-Wave base stations and short to medium range fixed wireless access (FWA).

The development of high bitrate wireless systems require efficient transmitters and receivers with high output power and sensitivity respectively. Efforts have been made for the generation and detection of continuous-wave (CW) millimetre-wave and terahertz signals by the means of both electronics- and photonics-based technologies [11], with key advances at device level using III-V semiconductor technology. Among the used approaches, the combination of photonics and electronics has shown some of the best results [12], where the transmitter part comprises of photonic devices and electronic detectors perform the reception. We employ the same approach for system demonstrations performed, using the optoelectronic devices presented in PhD thesis.

Figure 1-5 sketches a photonics-enabled wireless links in mm-Wave and THz domain, which highlights the potential of this approach to realize the seamless convergence of RF and fibre-optic links. Such an implementation enjoys the advantages already offered by the optical components and systems such as wide bandwidth, broad tuning range and compatibility with the existing fibre-optic infrastructure. Most importantly, contrary to the RF cables, the low-loss fibre cables allow the transport of high frequency RF signals over large distances, thus achieving antenna-remoting. This allows the centralisation of carrier generation and modulation, thereby reducing the size and complexity of the RF front-end.

Figure 1-5: The system concept for the convergence between wired optical link and an RF wireless link. After [13].

For the generation of RF signals using photonic means, the two main techniques that are commonly used are pulsed signal sources and optical heterodyne. In this thesis we employ the latter method for the experimental work conducted. The optical heterodyning is the most straightforward means of producing RF frequency (f_{RF}) requiring optical mixing of two optical wavelengths (λ_1 and λ_2 , see Figure 1-5) in a high-speed photomixer (photodiode or photoconductor), which generates the sum and difference frequencies at the output. The resultant f_{RF} can be obtained using Equation (1-2). Although flexible, in terms of implementation and frequency tunability, this approach inherits the problems of instability and low phase noise [15] due to the use of free-running lasers. Nonetheless, various techniques have been proposed by the researchers to improve the signal quality using locking techniques [16].

$$f_{RF} = \frac{\Delta\lambda}{\lambda_1\lambda_2}c \quad (1-2)$$

1.2.2 State-of-the-art mm-Wave and THz Links

This section summarized the current state-of-the-art in wireless link demonstrations in the mm-Wave and THz frequency range, comparing both the electronic and photonic technologies for RF carrier wave generation. Table 1-1 includes a summary of most prominent examples. It is worth highlighting that the highest data rates that have been achieved correspond to photonic technologies with the UTC-PD as a source on the

transmitter side. However, recently the PIN-PD was used to establish a link of over 100 Gbps in the 300 GHz band [21]. On the receiver end, the electronics-based detectors (both direct detectors and mixers) have been exclusively used. In the all-electronics approach,

Table 1-1: Summary of wireless link demonstrations with highest performance.

Data Rate (Gbps)	Frequency (GHz)	Distance (m)	Technology (Tx/Rx)	Bit Error Rate, Type	Modulation	Year	Ref.
74.4	92.5	0.9	Photonic Tx, UTC-PD /DBM	10^{-3} , offline	PDM-QPSK (2×2MIMO)	2012	[17]
2× 100	100	0.7	Photonic Tx, PD/DBM	10^{-3} , offline	PDM-QPSK (2×2MIMO)	2013	[18]
10	120	5,800	Photonic Tx UTC-PD+HEMT/ HEMT	$< 10^{-9}$, real-time	ASK	2012	[19]
64	240	850	Electronic HEMT/ HEMT	7.9×10^{-5} , real-time	QPSK	2015	[20]
128	300	0.5	Photonic Tx PIN-PD/ HEMT	2.2×10^{-2} offline	16-QAM	2019	[21]
2x 300	350	2.8	Photonic Tx UTC-PD/ SBD	2.7×10^{-2} , offline	64QAM-OFDM	2020	[22]
10	670	15	Photonic Tx UTC-PD /SBD	$< 10^{-9}$ real-time	ASK	2018	[23]

UTC-PD: Uni-travelling carrier photodiode; PD: photodiode; PIN-PD: *p-i-n* photodiode; DBM; dual-balanced mixer; HEMT: metamorphic high electron mobility transistor; SBD: Schottky barrier diode.

the notable demonstration employs frequency multipliers using GaAs based HEMT transistor technology for transmitter and the receiver [20]. For improving the data rate in a given frequency band, the combination of spatial and polarisation multiplexing techniques have been studied. Recent such examples includes a record link performance of 600 Gbps at 350 GHz using two polarisation orthogonal antennas [22].

1.3 Beamforming

The development of next generation mobile communication networks (e.g. 5G, 6G) is addressing the bandwidth challenge, leaping from currently available 20 MHz at microwave bands below 6 GHz to tens of GHz in the in the low-loss windows at millimetre-wave and terahertz bands. However, as we discussed in sub-sections 1.1.2 and 1.1.3, at these frequency ranges the free-space propagation losses and atmospheric attenuation are of primary concern. This is mainly due to the low RF output power generated by the currently available electronic and photonic THz signal sources, without a clear winner technology. It is a formidable system design challenge to minimise the losses at the generation side to deliver enough power at the receiving end. To address the challenge of boosting the radiated power at high frequencies beamforming is a promising technology to consider [24], taking advantage of the highly directive nature of THz antennas with small physical aperture.

Beamforming is defined as the art to control the spread of the electromagnetic wave in space, increasing the directivity of the beam, and eventually, controlling the radiation direction. In this chapter, we present an overview of mm-Wave and THz antenna array structures enabling dynamic beamforming and beam-steering. These will be needed in next generation small cell wireless access networks as illustrated in Figure 1-6. In a distributed architecture, the antenna sites contain remote radio head (RRH) units that need to be controlled from the base band unit (BBU). The connection between BBU and RRH is established through low-loss fibre-optic front-haul called common public radio interface (CPRI). This allows for the centralisation of resources and processing at central

Figure 1-6: An illustration of next generation communication system utilizing mm-Wave and THz beamforming.

office to reduce the RRH complexity. Complex antenna arrays with beamforming and beamsteering capabilities [26] are advantageous for the directional coverage to the mobile user equipment (UE), especially if the control of radiation direction is in the optical domain. Such a functionality is achieved by the means of phased array antennas that either includes the phase shifters or delay elements. By introducing a progressive phase shift in signals that feed the antennas, the radiated beam can be steered. In CHAPTER 5, we provide a detailed overview about mm-Wave beamsteering and demonstrate a phased array antenna module phased using photonic techniques.

1.4 Thesis Scope

1.4.1 Objectives

The scope of this doctoral thesis is to enhance the characteristics of a photonic-enabled millimetre-wave wireless link aiming to increase the radiated power by means of optimised antennas which enable array configurations in order to support high data rates and meet the requirements for next generation systems. Since the 5G/6G devices are expected to be highly mobile and handy, it is achieved by minimising the size of RF front-end. The target is to design and optimise the planar antennas on low-cost commercial laminates and to integrate them with high-speed photodiodes in hybrid fashion. We present technique for opto-electronic integration with purpose of maximising the power transfer from photodiode to the antenna. Moreover, a photonic-integrated emitter with planar phased array antennas is developed to perform beam-steering that forms the basis for large-scale integrated antenna arrays for 5G/6G systems. This thesis aims to improve the characteristics of millimetre-wave wireless link includes the design of THz detectors with large baseband bandwidth, capable of operating across several frequency bands. For the proof of concept, experiments are conducted using the custom-built devices capable of transmitting more than 12.5 Gbps of data signals in f_{RF} from mm-Wave to sub-THz.

1.4.2 Structure

The rest of the thesis highlights the work conducted, the mm-Wave/THz devices developed and experimental results achieved. The organisation rest of the chapters is as following.

Chapter 2 provides a description of the technologies used for the generation and detection on mm-Wave and THz signals. The focus is given on the two primary devices used in this thesis, namely high-speed photodiodes and Schottky barrier diodes.

Chapter 3 covers the development of antenna-integrated photonic emitters operating in the E-band frequency range (60 – 90 GHz). We demonstrate that highly

directive planar antennas, along with novel integration techniques, can be used to develop signal sources with small footprint. As a result of this study we develop two modules using a different type of high-speed photodiode structure for each source, one using a UTC-PD and a PIN-PD for the other. Moreover, the experiments presented are aimed at measuring the output performance and application of the modules in a wireless communication setup.

In Chapter 4 we describe the development of photodiode-based sources operating in the J-band frequency range. We present two modules developed using high-speed PIN photodiodes, and assembled in WR-3 output waveguide packages. The first of these devices (THz-Emitter-1) is based on a direct optical-to-electrical generation, which a high speed photodiode. The second device (THz-Emitter-2) uses a hybrid photonic-electronic approach, using a PIN-PD followed by an electronic frequency multiplier chip integrated into a single package.

Photonics-enabled mm-Wave beam-steering for wireless communications is presented in Chapter 5. This is achieved by developing an integrated phased antenna array emitter based on the developments in Chapter 3. True time delay steering with a real-time wireless link is demonstrated.

Chapter 6 describes novel ultra-wideband detectors in the mm-Wave and sub-THz range. The detectors are based on low-barrier Schottky diodes and assembled in a quasi-optic package. Two designs are presented, with the aim of increasing the baseband bandwidth for direct detection in a data link. A baseband bandwidth of up to 25 GHz is demonstrated, with the detection performed from 50 GHz to 500 GHz.

Finally, in Chapter 7, we conclude the findings of the thesis and discuss the future work.

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CHAPTER 2

Millimetre and Terahertz Generation and Detection

2.1 High Speed Photodiodes

Photomixers (such as photoconductor and photodiode) are fundamental components for microwave photonic systems [1]. These devices provide the key optical-to-electrical conversion, with two main figures of merit being the generated electrical power and the -3 dB frequency bandwidth. Different detector structures have been developed, involving different types of material substrates (including indirect bandgap as silicon and germanium) to achieve both characteristics simultaneously in one device, which has proven hard to accomplish. When high-speed is desired, it is usual to rely on III-V materials (i.e. InP, GaAs) [2], [3]. While there exist several types of photodetectors [1], majority of telecommunication (1550 nm) detectors are based on pin-junction pin photodiodes (PIN-PD), which we discuss in this section and compare it with another key photodiode structure, the Uni-Traveling-Carrier photodiode (UTC-PD). In this thesis, these two types of photodiodes structures have been utilised to develop millimetre (mm-Wave) and Terahertz (THz) signal sources which will be explained in subsequent chapters.

2.1.1 High Speed PIN Photodiodes

The $p-i-n$ photodiode structure (PIN-PD) introduced in 1950 [4] is shown in Figure 2-1(a), composed of an un-doped intrinsic layer (usually of InGaAs), sandwiched between p^+ -type and n^+ -type semiconductors. The operation principle is the photo-electric effect, by which incident light is absorbed in the intrinsic region, generating an electron-hole pair. The photon energy ($h\nu$) of incident light must be equal or greater than the bandgap energy E_g of the absorbing material. This excites the electron from valence band (E_v) to the conduction band E_c . Due to the presence of internal electric field, the electron and hole travel to right and left, respectively, leading to photo-generate current I_{ph} . The induced I_{ph} results in a decrease in the electric field between the carriers, causing saturation. It can be compensated by applying an external bias voltage V_b . The photodiodes generally require a reverse bias voltage. This creates a strong field in the $p-i-n$ junction that helps in increasing the carrier drift velocity and reducing the transit time.

Figure 2-1: Diagram of: (a) pin photodetector showing photo-generation of an electron-hole pair, and (b) band structure.

In PIN-PD, both the holes and electrons contribute to the transit time limited bandwidth as both carriers drift through depletion region. This can be better seen in the band diagram in Figure 2-1 (b) [5]. Since the effective mass of a hole is larger than that of an electron, with a smaller velocity, the response of a PIN-PD is primarily restricted by the hole velocity [6]. This causes accumulation of holes which results in space-charge effect, decreasing the electric field and carrier velocity resulting in saturation. This can be mitigated by having shorter depletion region, which in turn increases the junction capacitance and limits the bandwidth. Nonetheless, excellent performance is demonstrated by the PIN-PD devices. As example, the 3-dB bandwidth in the excess of 100 GHz with an output RF power close to 0 dBm on chip level can be found in the literature [7], [8].

2.1.2 High Speed UTC Photodiodes

The bandwidth limitations in a *p-i-n* photodiode were addressed by the Uni-Traveling Carrier Photodiode (UTC-PD) structure, in which only one type of carriers circulate across the device. It was first introduced by T. Ishibashi in 1997 [9], engineering the active region of pin-PD to separate the absorption processes from the carrier transport. The band structure, sketch in Figure 2-2 [5], shows a narrow bandgap p-type (undepleted, InGaAs) layer absorbs the light creating electron-hole pair, while a wide bandgap un-doped (depleted, InP) layer is used for carrier collection which is transparent to the incident light at 1550 nm. The electron-hole pairs are generated in the absorption layer. The minority electrons drift towards the depleted collection layer, where they exhibit velocity overshoot, and then into the cathode [6]. A quasi-field in the absorption layer, present due to the bandgap grading [10], accelerates this process, thereby reducing the transport time of the electrons. This is a significant advantage over PIN-PD. A wide bandgap p-type on the anode blocks the electrons from diffusing into it [5]. On the other hand, the majority holes possess fast response within dielectric relaxation time by collective motion. As a result, the electron transport primarily determines the photo-

response of a UTC-PD [5], [6], [10]. Moreover, both the absorption and collection layers can be designed independently here.

Furthermore, the electron diffusion velocity, which is smaller than drift velocity in depleted region, can be boosted by designing a very thin absorption layer without increasing the device capacitance. This provides very large 3-dB bandwidth. This is in contrast with pin-PD, where RC charging time is inversely proportional to the layer thickness. Another advantage a UTC structure offer in comparison to PIN-PD is the higher saturation in photo-generated currents. This is due to the decreased space-charge effect in the depletion layer as a result of high velocity of the minority electrons. This, in turn, allows the UTC-PD to provide increased output RF power.

Since its first introduction, several derivatives of the UTC-PD structure have been reported, including the travelling wave UTC-PD (TW-UTC-PD) [11], [12], the modified UTC-PD (MUTC-PD) [13], [14] and near-ballistic UTC-PD (NBUTC-PD) [15]–[17]. While a 3-dB bandwidth of 170 GHz was reported using the original UTC structure [18], frequencies up to 325 GHz have been achieved with the modified structures (MUTC-PD [17]) reducing the active area dimensions. Figure 2-3 represents the values for the two figures of merit output RF power and -3 dB bandwidth reported recently for various types of high speed photodiode structures operating in continuous-wave signal generation. As

Figure 2-2: Band structure for (a) PIN-PD and (c) UTC-PD. After [5].

Figure 2-3: State-of-the-art of the best continuous-wave output powers achieved from high-speed photodiodes as a function of frequency.

can be seen, the list is dominated by the UTC structure including its variants. In mm-Wave range, the highest power is achieved using a TW-UTC-PD, reaching 10 dB at a high photocurrent of 37 mA in pre-saturation operation.

Increasing the emitter RF power by arranging the photodiodes into arrays has already been reported, achieving an RF power up to 26 dBm at 40 GHz with 4 photodiodes [20]. At higher frequencies, an RF power up to 0.79 dBm of radiated power at 300 GHz was reported by combining the output from two UTCs using an on-chip passive coupler [21]. The use of III-V semiconductors offers the advantage to further scale the power combining photodiodes and high-speed amplifiers, as reported by Umezawa et al. who demonstrated a zero-bias UTC-PD with on-chip InP PHEMT amplifier delivering a record high power of 16+ dBm at 83 GHz [22].

2.2 High-frequency packaging issues

In addition to the photodiode structures, one of the key issues at high frequency is the packaging of the PDs themselves, which turns into a serious challenge. This is another aspect that has been considered in the present thesis, due to its importance in the losses. Figure 2-4 presents the results for some of the best output power obtained from the integrated photonic emitter based on PDs in mm-Wave to THz range. The emitters are classified into three categories using different antenna schemes. For planar antenna integrated emitters, as we can see, the results achieved in this work are among state of the art.

Figure 2-4: State-of-the-art of the best continuous-wave output powers achieved from the integrated photonic emitters using horn antennas ([21],[23],[24],[29]), quasi-optic ([25],[26]) and planar antenna ([27],[28]) employing PIN-PD, UTC-PD and its variants.

2.3 Electronic Sources

2.3.1 Oscillator Diodes

The resonant structures are used to generate high-frequency signals. They rely on negative differential resistance (NDR) to obtain the oscillations. The NDR occurs when the slope of the I - V curve is negative, i.e. $\Delta v/\Delta i < 0$ and such devices can oscillate or amplify the signals. Various devices exist to generate the NDR using different methods, resulting in variation in the performance such as output power, frequency and phase noise.

Resonant tunnelling diode (RTD) is a promising structure, achieving the some of the highest oscillation frequencies in the excess of 1 THz [30], [31]. An RTD is formed by sandwiching a quantum well film between two semiconductors with high band-gap, forming hetero-barrier. For certain level of applied voltage, the electrons tunnel through the barrier when the quantum well resonant state energy level is in alignment with the conduction band edge [32]. The maximum current is observed at this point. Applying further voltage thereon misaligns the energy level of quantum well film and conduction bands and this results in a decrease in the current as the electron mobility drops. In this region, an RTD possess a negative differential conductance (NDC), that can compensate the oscillator's ohmic loss. RTDs can be integrated with the antennas to radiate the generated signal and can be scaled to increase the power. The reported power at 300 GHz is 0.15 mW [33]. In [30], an average power of up to 0.72 mW is obtained by achieving a large array of 89 diode elements.

The other diodes used to generate the NDR are Gunn diode [34], IMPATT diode [35] and tunnelling transit time (TUNNET) diode [36]. A Gunn diode used only one semiconductor material, with GaAs and InP being common, with no pn junction and utilises Gunn-effect [37] to generated NDR. The Gunn oscillators with good noise performance can be realised and used in applications like instrumentation and radar [38]. An IMPATT diode achieves the NDR due to the time delay between the applied voltage and current, with no negative slope in I - V curve. The TUNNET diode, which is the variant

of IMPATT with improved noise performance, uses the tunnelling process. Although decent C. W. power levels are obtained, the noise performance is generally poor for diode oscillators.

2.3.2 Transistor Technologies

The transistors are more attractive devices than diode oscillators, due to their ability to be used as amplifiers [39]. Due to this property, the transistors form majority of the integrated components in the telecoms. Nowadays, they are also used for the generation of mm-Wave and THz frequencies [40], [41]. Various topologies exist such as CMOS, heterojunction bipolar transistor (HBT) and high-electron-mobility transistor (HEMT). For generation of THz frequencies, the metamorphic high-electron-mobility transistor (mHEMT) most preferable design. The cut-off frequency of more than 1 THz is reported [39] for mHEMT. The HBTs have also demonstrated such high frequencies but with inferior phase noise [42].

2.4 Schottky Barrier Diode

In this last section, we introduce a key component in millimetre and terahertz wave detection, the Schottky barrier diode (SBD). It is a semiconductor diode that utilises the non-linear properties of metal-semiconductor contact [43] formed by the junction of a metal (usually gold or titanium, which forms the anode [44]) with a semiconductor (forming the cathode). This junction creates an electrostatic barrier due to the different work functions of metal and semiconductor which is responsible for the rectification properties of an SBD. The conduction is mainly controlled by thermionic emission of majority carriers over the barrier. Therefore, an SBD is a majority carrier device for which the cut-off frequency is not limited by minority carrier effects.

The combination, size and condition of metal and semiconductor primarily determine the electrical properties of Schottky diode. The main factor determining the quality of an SBD as a detecting and mixing device is its cut-off frequency f_c [45], [46]

$$f_c = \frac{1}{2\pi R_s C_j}, \quad 2-1$$

where R_s and C_j are series resistance and junction capacitance, respectively. The value of R_s is determined from $I - V$ characteristics and C_j is extracted from zero-bias operation [45]. For high-frequency operation, it is desirable to have low series resistance and junction capacitance. This necessity implies the semiconductors used should possess high carrier mobility and saturation velocity [44]. For mm-Wave and THz SBDs, gallium arsenide (GaAs) semiconductor has mostly been the material of choice. For lower frequencies, silicon-based SBDs are widely used due to their significantly lower cost. Since the n-type materials offer greater carrier mobility than the p-type ones, they are exclusively used in the formation of Schottky diodes. A general drawing for the construction of SBD is shown in Figure 2-5. It shows a cross-sectional view of Schottky diode used for varistor and a varactor. The structure of the both is essentially the same. However, for a varactor diode, which is used for frequency multipliers requiring high output power, a large epilayer thickness is used [46]. However, this increases the capacitance two to three times than that of a varistor, used for frequency mixing.

Figure 2-5: Diagram showing the cross-sectional view of Schottky barrier diode used as (a) varistor and (b) varactor [44].

For designing the mm-Wave and THz devices based on Schottky diode, such as mixers, rectifiers and multipliers, it is a common approach to use the equivalent circuit. Figure 2-6 shows the principal circuit model of Schottky junction that includes the key parameters used for determining the device performance [43]. The use of such a lumped model for developing a wideband square-law detector will be presented in CHAPTER 6 of this thesis. The C_j mainly results from the depletion region at metal-semiconductor junction. It is an important parameter of SBD, which defines the operation frequency of the diode as determined from equation 2-1. For the f_c in the range of mm-Wave/THz, the C_j of the order of a few pF is required. For the case of zero-bias SBD, C_j is represented as C_{j0} .

Figure 2-6: A principal equivalent circuit model of Schottky barrier diode.

The Schottky diodes, due to their ability to operate in wide range of temperatures from 4 – 300 K [47] (cooled and un-cooled), have recently become some of the most used elements in THz and sub-THz systems, in the sources and detectors. They widely are used in the direct detectors as well as and heterodyne mixing receivers, with the performance comparable to the thermal detectors in respect to responsivity and noise equivalent power (NEP). The ease of integration with other devices and the possibility of bias-free operation (zero-bias SBD) are particularly attractive features for developing compact detectors.

This thesis focusses on the use of SBD-based direct detectors, which can be generally classified as waveguide integrated detectors (WG-SBD) and quasi-optic detectors (QO-SBD). The former has an advantage of higher sensitivity only in a particular frequency band, the latter offers the ultra-wideband detection capability covering frequencies from tens of GHz to 1 THz, albeit with lower sensitivity. While the intrinsic voltage responsivity of the SBD chip can range from 4000 – 30,000 V/W depending on anode diameter and junction capacitance [48], the actual sensitivity of the direct detectors is always limited by the RF coupling losses and significant impedance mismatch between the diode and the antenna. The commercially available WG-SBD from Virginia Diodes [49] offer a typical responsivity and NEP of 2800 V/W and 2 pW/Hz^{1/2} at 100GHz, and 750 V/W and 15 pW/Hz^{1/2} at 1 THz [50]. A high responsivity of 30,000 V/W [51] for a space-based WG-SBD detector at 89 GHz, designed for space applications, was demonstrated using SBD chip from ACST [52]. For QO-SBDs, moderate responsivity in the range of 300 to 1000 V/W have been reported [53], [54]. The lower responsivity is due to the fact that wideband QO design complicates the impedance matching of the diode to the antenna and the output readout connector. Nonetheless, techniques has been presented to increase the responsivity up to 7000 V/W by employing narrow-band resonant antennas[48], [55].

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CHAPTER 3

Photonic-based Millimetre-wave Sources

This chapter is devoted to the development of antenna-integrated photonics-based emitters for high speed wireless communications. The emitters use two different photomixer technologies, UTC-PD and PIN-PD, to generate mm-Wave power in the E-band (60 – 90 GHz) frequency range. We will demonstrate how the planar antennas can be used to develop efficient RF front-end devices that lead to possibilities not achieved otherwise with traditional approaches for photonic emitter devices.

A photodiode (PD) is the most common photomixer device, which has been used in the past years to develop mm-Wave and THz sources. To radiate the generated RF power from the photodiode, two main approaches have been widely used: Quasi-optic approach, in which the PD chip is mounted on an electrically large (and expensive) silicon (Si) lens, or the rectangular waveguide approach, using a launcher to couple the radiation into a rectangular waveguide followed by a horn antenna to collimate the power in the desired direction [1]–[5]. Both approaches aim to maximize the radiation efficiency, with high coupling efficiency and antenna directivity. Due to the relatively large size of these structures, it gets very difficult to integrate customised antennas, scale up the number of emitter elements and realise integrated emitter arrays.

In this context, we have designed and developed compact photomixing emitters operating in the E-band, with performance comparable to that of Si lens and horn antenna, which present a reduced footprint. The modules cover the complete 60 – 90 GHz range

with flat response, providing up to -5 dBm radiated power, which we will demonstrate within a real-time error-free wireless link supporting up to 5 Gbps of bitrate. Moreover, we present a novel scalable planar antenna approach that provides broad bandwidth and high gain, which allows efficient integration with high speed PDs to become an attractive alternative.

This chapter is organised in four sections. The first one covers the design of proposed photonic emitter, along with the details of the antenna and the optoelectronic integration and assembly. Second and third sections cover the characterisation and demonstration of the emitters based on UTC-PD and PIN-PD, respectively.

3.1 Emitter Design

3.1.1 High Directivity Planar Antenna and Passive Circuit

We designed a tapered slot antenna (TSA), with an aperture profile shaped by Fermi-Dirac function [7] which radiates in end-fire direction having a highly directive beam. The design is aimed for low power and point-to-point communications applications. The antenna is implemented on a commercial low-loss substrate Rogers RT/Duroid 5880 ($\epsilon_r = 2.2$) with 130 μm thickness, both key parameters to meet effective thickness constraints [6], avoiding excitation of substrate modes, which distort the radiation pattern and boost the losses.

The geometry of the Fermi TSA is shown in Figure 3-1, in which the profile of the antenna follows the function

$$f(x) = \frac{a}{1 + e^{-b(x-c)}}, \quad 3-1$$

where $a = \lambda_0/2$ (half the aperture width), $b = 2.5/\lambda_0$ and $c = 2\lambda_0$, where λ_0 is RF wavelength in air (the centre frequency f_0 is selected within the E-band, 80 GHz). The three parameters are optimised by performing full-wave simulations to achieve the best directivity. The length L is $4\lambda_0$, the aperture width W_a is equal to λ_0 ($2a$) and the distance d from the tip of the aperture to the edge of antenna d is equal to $\lambda_0/2$. Along the length of antenna are corrugated structures that include periodic slots. This feature enlarges the effective aperture of the antenna, thereby improving the directivity. Other advantages of corrugations are that they help achieve symmetric radiation patterns in E- and H-plane, and reduce the size of a conventional TSA and side-lobe level reduction. This is due to the fact that current flowing along the antenna length encounter larger area. The width W_c and the length L_c of these slots are $0.16\lambda_0$ and $0.11\lambda_0$, respectively, with a spacing s equal to the width (W_c). The slots are made wider than proposed by the authors in [7] to ease the fabrication and achieve desired performance.

Figure 3-1: Layout of high gain tapered slot antenna with CPW-to-slotline transition. The left inset shows the structure of bottom metal layer. Dimensions (unit: mm): $L = 15$, $W_a = 3.75$, $d = 1.85$, $L_c = 0.41$, $W_c = 0.6$, $s = 1.2$. Colours: orange (metal) and green (substrate).

About the contacts, on the left-hand side of the figure, since PD chips commonly used coplanar waveguide (CPW) output pads, the designed antenna features a CPW to slot-line transition (balun). The balun includes an open circuit stub that converts the CPW mode (odd-mode) into slot-line mode (even-mode). This is done by introducing a 180° phase shift in one of the slots of CPW. The stub has a length L_{stub} of $\lambda_g/4$, with λ_g being the guided wavelength at f_0 . The bottom view of the input section of antenna is shown in the left inset of Figure 3-1 that features a smooth transition from grounded CPW to a CPW.

The simulated gain of the TSA over the E-band, shown in Figure 3-2, exhibits a flat gain over the band of interest (71 – 86 GHz), with a 14.5 dBi peak at 85 GHz. The calculated beam profile of the antenna is presented in Figure 3-3 for both, E- and H-planes, at four difference frequencies: 67 GHz, 76 GHz, 81 GHz and 86 GHz. The co- and cross polarisation patterns for both the planes are also shown, observing that it is highly directive, with 3-dB beam-width from 23° to 22° and 40° to 30° in E-plane and H-plane, respectively. The side-lobe level (SLL) is down to -21 dB and -15 dB. Moreover, the E- and H-plane profiles become symmetric as f_0 increases. Considering the fact that the antenna is planar and single element, this performance is a considerable improvement.

Figure 3-2: Simulated gain of TSA in E-band frequency range (60 – 90 GHz).

Figure 3-3: Simulated normalised radiation patterns of the high gain planar TSA at 67 GHz, 76 GHz, 81 GHz and 86 GHz. E-Plane Co-Polarisation (Solid red), E-Plane Cross-Polarisation (dotted orange), H-Plane Co-Polarisation (dashed blue) and H-Plane Cross-Polarisation (dash-dotted green).

In order to validate the performance of designed TSA before integrating it into the photonic emitter, we fabricated a test antenna fed from a 1.85 mm (V) coaxial connector, shown in Figure 3-4(a). The radiation patterns were measured in an anechoic chamber at the maximum frequency of 67 GHz, limited by the equipment available. The measured

results, shown in Figure 3-4(b), give a very good match between the simulated performance expected from design.

(a)

(b)

Figure 3-4: (a) Fabricated TSA with a 1.85 mm (V) coaxial connector and (b) measured radiation patterns at 67 GHz. E-Plane Co-Polarisation (solid red), E-Plane Cross-Polarisation (dotted orange), H-Plane Co-polarisation (dashed blue) and H-Plane Cross-polarisation (dash-dotted green).

3.1.2 Emitter Structure and Assembly

Once that the performance of the antenna design has been verified with the connector coupled sample experiment, we move on to the optoelectronic (O-E) integrated approach, feeding the antenna with a photodiode. In this case, also it becomes critical to take into account the optical and electrical accesses.

After a literature review for previous work on O-E assemblies, we have designed a custom housing to assemble and integrate all the components. The overall 3D structure of the emitter module is shown in Figure 3-5(a), which supports the assembly of the PD chip with the RF board with the antenna as well as the attachment of the optical fibre. The inset in Figure 3-5(a) is a close-up view of the PD chip feed point where the PD is

modelled as a CPW excited by a lumped port with impedance, junction capacitance C_j and series resistance R_s shown in the equivalent circuit, equal to that of active diode. Z_A is the feed point impedance of the antenna circuit is kept at 67Ω instead of PD's 50Ω , due to fabrication constraints, which is expected to cause this mismatch. This model is used to simulate the RF behaviour of the assembly. The chip is mounted on an aluminium nitride (AlN) block, for proper heat dissipation. Here, the photodiode is coupled to the antenna by the means of an electrically conductive silver epoxy, modelled as blocks of silver. This is a different approach to the commonly used wire-bonding technique which reportedly introduces up to 9 dB of coupling loss [8] at E-band frequencies. Figure 3-5(b) shows the final layout of the overall passive circuit, which in addition to the designed antenna, the RF board also includes an RF Bias Tee design including a choke/block filter in the DC line to allow biasing the PD.

Figure 3-5: (a) 3D model of E-band planar antenna-integrated photonic emitter, showing internal structure of housing and fixing of optical fibre. Inset: Structure for PD-antenna integration used for simulate the emitter. (b) Layout of the passive circuit used for emitter assembly that includes biasing network. Inset: RF choke filter in the DC bias line.

The emitter model has been simulated, using the signal access ports as indicated on the insets of Figure 3-5. The achieved results for the numerical S-parameters (S_{11} , S_{21} and S_{31}) are presented in Figure 3-6(a), showing that the insertion loss S_{31} remains below 1 dB over the complete frequency range, indicating good coupling from the PD to the antenna and that the isolation S_{21} between the RF input and DC ports is better than 20 dB. For the sake of comparison, we plot the simulated frequency response for silver epoxy coupling against a standard 25 μm diameter gold wire bond in Figure 3-6(b). The advantage of proposed coupling technique can easily be noticed here.

Figure 3-6: (a) Simulated S-parameters of the E-band emitter. PD to antenna coupling (S_{31} , dash-dotted green), DC port isolation (S_{21} , dashed blue) and reflection at PD port (S_{11} , solid red). (b) Comparison of frequency response for silver epoxy and 25 μm wide gold wire.

3.2 UTC-PD Emitter

The UTC photodiode used in this work was fabricated at III-V Lab, in a previous project coordinated by UC3M [9], [10] (Project iPHOS [11]). A microscopic picture of the chip is shown in Figure 3-7(a), along with the dimensions. The PD has active area of $4 \times 15 \mu\text{m}^2$ and 50Ω CPW output gold pads, fabricated on 120 μm thick indium phosphide (InP) substrate. The device is edge-fed by an optical waveguide. It has a very low R_s of 20Ω and a C_j of 38 fF. For providing an idea about the intrinsic performance of the chip and compare it with our emitter in next sub-section, we include the measured frequency response and output power in Figure 3-7(a) and (b), respectively, for chips with various

dimensions. The $4 \times 15 \mu\text{m}^2$ PD exhibits a 3-dB bandwidth of 78 GHz and saturated output RF power up to 0 dBm.

Figure 3-7: (a) Microscopic image of the $4 \times 15 \mu\text{m}^2$ UTC-PD chip with its physical dimensions. (b) and (c) Measured on-chip frequency response and output power, respectively, for UTC-PD with various active area dimensions. From [9].

After manufacturing all the components for the housing and antenna board, we assembled the module which is presented in Figure 3-8(a). A lensed optical fibre with 5 μm spot size is aligned to the PD optical input waveguide using a multi-axis positioner. Figure 3-8(b) shows the image of UTC-PD chip electrically coupled to the antenna passive circuit by threads of silver-filled epoxy, which was the most demanding task in

the assembly process given the microscopic size of the contact pads. Once the prototype has been developed, we proceed towards its characterisation to establish its performance.

Figure 3-8: (a) Packaged E-band photonic emitter module with protruding TSA and (b) microscopic photograph of the UTC-PD chip coupled to the antenna using a conductive silver epoxy. A lensed fibre can also be seen, coupled from the backside of the module.

3.2.1 Emitter Characterisation

For the characterization, we have measure the absolute radiated power from the assembled module over the entire operating frequency range in the mm-Wave range. The RF frequencies are generated using the optical heterodyne technique, using the experimental setup sketch on Figure 3-9(a). We combine two CW tunable external cavity lasers (TECL, Sacher Lasertechnik) centred around 1550 nm, generating optical wavelengths with a spacing corresponding to the desired RF frequency f_{RF} . One of the ECLs is modulated with a square wave with 12 Hz oscillation frequency, providing the reference for lock-in detection. The optical spectra are monitored on an optical spectrum analyser (OSA) and the actual spectrum can be seen in the inset of the Figure 3-9(a). An erbium doped fibre amplifier (EDFA) amplifies the optical input power P_{opt} , injected into the emitter module through the lensed fibre. We expect more than 3 dB coupling losses at this point from mode mismatch since the fibre has a 5 μm spot size, twice the recommended 2.5 μm. The emitter RF power is radiated into free-space, using a Thomas

Keating calibrated absolute THz power meter (TK) [12] to detect it. The TK meter is an optoacoustic sensor with a large aperture of 30 mm, minimum sensitivity of 1 μW and

Figure 3-9: (a) Block diagram showing the schematic arrangement for characterisation of photonic E-band emitter and (b) photograph of laboratory setup, using a TK power meter and Teflon lenses for collimation.

a calibrated voltage responsivity of 0.5 V/W. An important aspect is that this power meter has to be located at Brewster angle (55.5°) with respect to the E-plane of TSA, to minimise the reflections from the sensor window materials. The photodiode is biased at a voltage V_{bias} of -3 V (reverse bias). A picture of actual laboratory setup is in Figure 3-9(b), where two plano-convex lenses can be observed to collimate the RF power.

Figure 3-10(a) provides the comparative plot for the simulated and measured frequency response in the range of 70 – 86 GHz. We can observe that the peak power reaches up to -9.25 dBm (118.8 μ W) at f_{RF} of 74 GHz, with fluctuations that are within ± 1 dB. The results show stable response of the developed emitter, and a reasonable correlation between the fluctuations on both plots. A peak at 74 GHz in the measured trace relates to the peak at 73 GHz in the simulated one. In order to determine the RF bandwidth, the output power is measured over the whole E-band range at a photocurrent I_{ph} of 10 mA, shown in Figure 3-10(b). The emitter provides a 3-dB bandwidth of almost 30 GHz, which covers the 71 – 76 GHz band and 81 – 86 GHz band, corresponding to a 40% fractional bandwidth which to the best of our knowledge, is the widest reported for an E-band photonic emitter. The sharp drop in the response below 60 GHz comes from the cut-off of the integrated RF choke filter. Comparing the measured radiated power with the on-chip response shown above in Figure 3-7 [10], we can estimate a loss of about 5 dB in the RF coupling. This loss comes mainly from the dielectric losses due to the long length of passive circuit; 10 mm long from feed-point to the antenna and coupling through the conductive epoxy. It must also be mentioned that the PD is not operated at its maximum capacity. The module is expected to yield -6 dBm peak radiated power.

Figure 3-10: (a) Measured and simulated normalized frequency response of photonic transmitter at 5 mA photocurrent and (b) Measured radiated power over complete E-band frequency range at 10 mA photocurrent showing a 3-dB frequency bandwidth of more than 28 GHz.

In addition to the frequency response, it is also important to measure the dependence of radiated RF power P_{rad} with respect to the photodiode DC bias current (I_{DC}), which depends on the optical input power. This is shown in Figure 3-11 at four frequencies values: 71, 75, 81, 86 and 95 GHz. The output power increases linearly on a dB scale of the photocurrent, showing the square law characteristics of UTC-PD; $P_{rad} \propto I_{ph}^2$. No saturation of the output power is seen at any f_{RF} . This is because the I_{ph} is limited to 10 mA to prevent damaging of PD chip, since no thermal control was used at the time of the experiment. In principle, these UTC-PDs start to saturate at current I_{sat} of 16 mA when operated without thermal control [9]. With these characteristics of UTC-PD in consideration, we use the curve fitting method to figure out the maximum achievable P_{rad} based on measured data. The fitted power trace at 71 GHz is given in Figure 3-12. It indicates that the emitter module can provide up to -5 dBm of saturated RF power.

Figure 3-11: Measured radiated power versus DC photocurrent at various carrier frequencies at -3 V bias.

Figure 3-12: Radiated power versus photocurrent measured for f_{RF} of 71 GHz at V_{bias} of -3 V. The data is fitted and extended up to UTC-PD's saturation current of 16 mA.

3.2.2 Wireless Link

Now we turn towards the application demonstration, using the developed module as a transmitter element in a real-time wireless communication link experiment. Figure 3-13(a) shows the block diagram of the link. At the transmitter, we use the optical heterodyne method based on two tunable ECLs, setting the wavelength difference to generate a 75 GHz carrier frequency f_{RF} . An electro-optic amplitude modulator (EOAM) modulates the two optical sub-carriers with a stream of NRZ data coming from a pulse pattern generator (PPG), hence leading to on-off keying (OOK) modulation scheme. The dual sub-carrier modulation method is used here, as it provides better SNR than single sub-carrier modulation [13]. The output from the modulator is boosted with an EDFA and then fed into the emitter lensed fibre. The EDFA allows to control the level of P_{opt} . The V_{bias} in this experiment is kept at -2.5 V. The generated mm-Wave signals are generated and transmitted by the TSA. The wireless path includes two Teflon lenses, to collimate the beam, each of which provides a gain of 14 dBi [14]. On the receiving side, we use an envelope detection based on a Schottky barrier diode based detector. This is a square-law detector that down-converts the modulated data into baseband, where we recover it. The Schottky receiver is another module which has been developed during this thesis, and will be thoroughly explained in CHAPTER 6. A key advantage of using envelope detection is that it is not sensitive to phase noise fluctuation of the lasers (which operate in free-running condition) and eliminates the need of local oscillator (LO) at the receiver. After the recovery, the data signal is boosted by a low-noise amplifier (LNA) and a limiting amplifier (LA) is employed to reshape it. Finally, it is fed into a bit error rate tester (BERT) for the link performance analysis. Figure 3-13(b) shows the actual experimental arrangement in the laboratory, with link distance and antenna gains mentioned.

(a)

(b)

Figure 3-13: Block diagram for the experimental setup for photonics based wireless communications link with OOK modulation in E-band using developed UTC-PD based emitter and (b) photograph of the laboratory setup.

We conducted the data transmission starting from bitrate of 1.5 Gbps, increasing to 3 Gbps and 5 Gbps. The measured performance in terms of bit error rate (BER) characteristics against radiated power is given in Figure 3-14. An error-free transmission ($BER < 10^{-10}$) is successfully achieved at all the bitrates. The required RF power from the emitter is -22.5 dBm (5.6 μ W), -19 dBm (12.5 μ W) and -18 dBm (15.8 μ W) for the bitrates of 1.5, 3 and 5 Gbps, respectively. In Figure 3-15, clean and open eye diagrams for all three cases can be seen. However, a relatively high root mean square (RMS) jitter

was measured at 5 Gbps around 16 ps which prevented to achieve error-free link at higher bitrates. According to the simulations, this was the result of relatively higher group delay in the passive antenna circuit.

The link distance was 0.5 m. A link budget analysis, using the Friis relation, we have estimated the maximum achievable distance considering the measured RF power and gains at each of the stages of the link (Figure 3-13(b)). We estimate a transmission distance of over 4 m can be achieved using the setup described above, which is subject to the sensitivity of the detector. From these results, it can be said that the proposed photonic emitter can find applications in high speed communications systems.

Figure 3-14: Measured BER characteristics of the error-free wireless link at the bitrates of 1.5 Gbps, 3 Gbps and 5 Gbps at f_{RF} of 75 GHz.

Figure 3-15: Demodulated eye diagrams of the error-free wireless link at the bitrates of 1.5 Gbps, 3 Gbps and 5 Gbps at f_{RF} of 75 GHz.

3.3 PIN-PD Emitter

In this section we discuss a second E-band emitter which we have developed, this time using a high-speed PIN-PD as photomixer. The packaged module is shown in Figure 3-16, in which the antenna is the same as described in Section 3.1.1, developed for the UTC-PD based emitter. Here, the PD contacts to the antenna are done with multiple wire bonds, with a gold wire of 25 μm diameter instead of the conductive epoxy. A cleaved-facet optical fibre is butt-coupled to the PD chip, which includes optical spot size converter, and finally pigtailed to the module.

Figure 3-16: Packaged E-band PIN-PD based photonic emitter module with TSA and (b) internal view of the module showing the assembly of the PD chip with the antenna using gold wire-bonds. The optical fibre is attached using butt-coupling method and is pigtailed to the housing.

3.3.1 PIN-PD Frequency Response

The high speed PIN photodiodes are designed and fabricated by the THz Sensors and Systems Group at Fraunhofer HHI [15], where this work was carried out as part of joint collaboration within a COST action short term mission. These PIN-PDs are designed to deliver high output RF power at sub-THz frequencies, thereby are an attractive alternative to UTC-PDs as the band structure of PIN-PDs is less complex. Until today, these devices have only been developed with on-chip integrated antennas, employing a quasi-optic technique (Si lens) for applications in THz sensing.

A picture of the InP-based waveguide-fed PIN-PD chip is shown in Figure 3-17(a). The active part of the chip [16] is fabricated in a mesa structure with p-contact for the signal. No back-side metallisation and anti-reflection coating are used. Various devices are fabricated with different active area dimensions ranging from 28 to 40 $\mu^2\text{m}$. Typical diode parameters for these PD are R_s of 35 Ω and C_j of 18 fF. To establish the intrinsic performance, we measured the frequency response of PIN-PDs using a calibrated setup based on a lightwave component analyser (LCA) and CPW RF probes. The measured relative frequency response is given in Figure 3-17(b) for four different chips. The highest 3-dB bandwidth obtained is 50 GHz for the $4 \times 10 \mu^2\text{m}$ PD, while for the rest; 5×7 , 4×7 and $3 \times 12 \mu^2\text{m}$, it is 40 GHz. These measurements were performed at a DC photocurrent of 5 mA, to make sure the device's optoelectronic response is not saturated. The bias voltage was kept at -2V. It is noteworthy that the equipment used here was limited up to 67 GHz. Since the diode parasitics are very low and comparable to those of UTC-PD, higher bandwidth is expected for PIN-PDs. The relatively high ripples in the curves are believed to be caused by the reflections from the electro-optic modulator used in the LCA. The $4 \times 10 \mu^2\text{m}$ chip is used to develop the emitter presented in this section, as it gives the highest response.

Figure 3-17: (a) Microscopic image of the PIN-PD chip showing optical waveguide and electrical contact pads and (b) measured on-chip frequency response for four PIN-PDs with different active area dimensions.

3.3.2 Emitter Performance

The E-band emitter module was also tested to characterise the output performance. The experimental setup remained the same as in Figure 3-9(a), except that the a pyroelectric thin-film (PTF) sensor (THz20HS, Sensor and Lasertechnik GmbH) is used to detect the power. This PTF detector, calibrated by Physikalisch-Technische Bundesanstalt (PTB), has a sensitivity of 160 V/W, which is more than two orders of

magnitude better than the TK sensor. It is mounted in vertical orientation perpendicular to the antenna axis (E-plane) with a recommended 10° tilt angle to minimise the formation of standing waves. A picture of the setup is given in Figure 3-18. Figure 3-19 shows the measured output power over a large frequency band, ranging from 40 GHz to 120 GHz. The measurement is performed at I_{ph} of 12 mA and V_{bias} of -2 V. While the inset shows the frequency response over the E-band. It can be seen that the power variations are within 3-dB range in the desired band, relative to the peak in the curve at 62 GHz where the power is -5.09 dBm (310 μ W). This corresponds to a total bandwidth of 30 GHz, or a 40% fractional bandwidth, similar to the case of UTC-PD emitter described in previous section. It must be mentioned here that the fluctuations in the output of the EDFA used contributed to the unexpected peaks and dips.

The dependence of radiated power of the emitter module on the P_{opt} , and I_{ph} , is shown in Figure 3-20(a). The measured mm-Wave power at seven different frequencies is given. We can observe that the power increases linearly up to 12 mA DC current for all frequencies. In this regime the photocurrent increases quadratically with higher optical power and the PD is operated in a linear regime. For higher photocurrents, the PD starts to saturate and the RF power decreases. The highest power obtained is -1.39 dBm (726 μ W) at 54 GHz. For 70 GHz, the power peaks to -5.6 dBm (272 μ W). Note that HHI's PIN-PD performs as good as a UTC-PD in terms of output power, when comparing these results with our previous E-band module in section 3.2. The roll-off in the response after 90 GHz results from the frequency cut-off of the TSA and RF choke filter.

Finally, we consider the influence of applied DC bias voltage on the output power of the PIN-PD. The results are shown in Figure 3-20(b) for f_{RF} of 70 GHz. We find that the radiated power decreases by more than 5 dB when V_{bias} is reduced from -2.0 V to -0.5 V, and the PIN-PD saturates at much lower optical input power. This is due to the dependence of the size of the depletion region of the PD on the reverse bias voltage.

Figure 3-18: Picture showing laboratory setup for absolute radiated power measurement for the packaged E-band PIN-PD emitter.

Figure 3-19: Measured radiated power versus frequency at $I_{ph} = 12$ mA and $V_{bias} = -2$ V, showing 3-dB bandwidth of 30 GHz.

(a)

(b)

Figure 3-20: (a) Measured radiated power at various carrier frequencies versus optical input power and photocurrent at $V_{bias} = -2$ V and (b) radiated power at 70 GHz versus photocurrent at V_{bias} of -2 V to -0.5 V.

3.4 Summary

A new concept for the development of photonics based wireless emitters has been presented, allowing simple and scalable optoelectronic integration approach. We demonstrated two integrated emitter modules for E-band (60 – 90 GHz) photonics-enabled systems featuring a wideband and high gain planar antenna that meets the requirements of compact yet directional radiation for mm-Wave and THz sources. The modules integrate two different state of the art high speed photodiodes, UTC-PD and PIN-PD, an E-Band (60 – 90 GHz). Good agreement was obtained for the simulated and measured frequency responses, validating the technique presented. As shown by the experiments, the two emitters developed provide peak RF power of -5 dBm (250 μ W) and -1.39 dBm (726 μ W) in case of UTC-PD and PIN-PD, respectively. Then, a 3-dB RF bandwidth of 30 GHz was demonstrated for both modules, which is possibly the highest performance delivered by an E-band photonic emitter, as per our knowledge. Finally, to validate the applicability of such a design, a real-time OOK based wireless link was established using the using the UTC-PD emitter, delivering at least 5 Gbps bitrate.

In conclusion, it is possible to dispense the rectangular waveguide and horn antennas and exploit the new planar antenna techniques having equally high directivity and low-cost. Integrating such antennas with photomixing sources presents opens doors to the investigate the photonics-based phased array antennas in the emitters for beamforming and beamsteering, which is an essential feature for 5G systems and beyond. Such a module, based on the developments presented here, is demonstrated in CHAPTER 5.

3.5 References

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CHAPTER 4

Photonics-based Terahertz Sources

The potential of high speed photodiodes in developing multi-Gbps wireless communications in mm-Wave range was demonstrated in CHAPTER 3. Our research has also explored the development of photonics-based RF sources operating in the terahertz range of the spectrum, which is discussed in the present chapter.

Photonic techniques are of particular interest to THz-wave generation for wireless communications given the bandwidth and broad tunability of photomixers and lasers [1], [2]. The main type of photodiode employed for establishing wireless links have been high-speed UTC-PD reaching into 100 Gbps utilising both the packaged modules and standalone PD chips [1] – [4]. At THz frequencies, one of the main challenges for the device performance is the interconnections and packaging due to signal loss [5]. THz photomixers are mainly integrated in quasi-optic packages [6], [7], using broadband planar antennas (i.e. bowtie, log-spiral and log-periodic) which are monolithically fabricated with the photodiodes. A silicon lens is commonly used to extract the radiation out of the high permittivity semiconductor substrates, leaving little room for impedance matching between the two elements and for developing array configurations. In contrast, a rectangular waveguide is more convenient than the quasi-optic approach because of the possibility to integrate other components such as amplifiers and custom antennas. Moreover, the inductive loading can easily be performed to compensate the capacitive reactance of the PD chip and improve the output power. There have been only a few THz UTC-PD modules available in a rectangular waveguide package, first developed by

Nippon Telegraph and Telephone (NTT) in Japan, with commercially available modules with WR-3 [8] and a WR-1.5 [9] output waveguides.

In this chapter, we present fully integrated WR-3 emitters working in J-band (220 – 325 GHz) based on high power PIN photodiodes. Two modules have been designed, developed and packaged following two different approaches to reach into the THz range. The first module is based on a purely photonic approach, while the second one is based on photonic-electronic co-integration combining a PIN-PD with an electronic frequency multiplier in a single package. The primary goal and outcome of the work presented in this chapter is the first ever successful development of a WR-3-integrated PIN-PD emitter module.

First, the structure of the modules is presented. This includes the design of the components involved and their packaging. This is followed by the simulated performance. Finally, a complete characterisation is performed, together with measurement of integration losses, and output performance established.

4.1 Terahertz WR-3 Integrated PIN-PD Emitters

4.1.1 Design, Simulation and Packaging

The block diagrams of the two target modules to reach into the THz range are presented in Figure 4-1. Each of the modules is referred to as THz-Emitter-1 for the fully photonic approach, and THz-Emitter-2 for the photonic-electronic co-integration. The photonic carrier generation involves tunable laser sources (TLS), consisting of two ECLs with frequency spacing f_0 . The first device, shown in Figure 4-1(a), features two main components: (a) A PIN-PD chip, and (b) a WR-3 waveguide transition (or coupler). The optical signal is down-converted into the THz range by the PD chip, directly generating the THz signal which is then guided to the hollow waveguide by means of the coupler, and radiated through a horn antenna.

In the second emitter module, shown in Figure 4-1(b), we propose for the first time the introduction of an additional component, an electronic frequency multiplier following the PIN-PD, leading to a hybrid photonic-electronic signal source. Here, the 100 GHz signal generated at the PD is fed into the Schottky-diode Tripler ($\times 3$ multiplication factor) generating a 300 GHz signal is then fed into the WR-3 waveguide. The PD behaves as a photonic local oscillator (LO) for the multiplier. The novel feature for THz-Emitter-2 is that for the first time a THz photomixer source is co-integrated with

(a)

(b)

Figure 4-1: Block diagram of photonics-based THz emitters with WR-3 rectangular waveguide output fed by photonic signal source (a) THz-Emitter-1 for direct RF generation from a PIN-PD and (b) Emitter-2 integrating a PIN-PD, as a photonic LO for 100 GHz, and a tripler chip, generating 300 GHz signal, integrated in single package.

an electronic one, in a single package. While the previous such demonstrations in literature include the cascading of packaged commercial UTC-PD and Schottky tripler devices [10]. There are two fundamental advantages of the new approach presented: First, the PDs working in the THz range have a roll-off in their output, since they are being operated beyond their 3-dB bandwidth. By incorporating a multiplier chip, the output frequency can be increased. Secondly, this allows to integrate a chain of amplifiers (both the low-noise and power) at the LO frequency which are commercially available and at much lower cost than in the THz range. Finally, the high power handling capability of Schottky diodes (hundreds of mW) comes as an added advantage to this hybrid photonic-electronic scheme. We believe this approach can outperform an all-photonic one, while retaining the compatibility with photonic networks.

The design and assembly approach for these modules is based on the experience that we gained from the development of the mm-Wave modules presented in the previous chapter, using in this case a rectangular waveguide coupling approach shown in Figure 4-2(a). The modules use a PIN-PD chip, with $5 \times 7 \mu^2\text{m}$ active area and 50Ω coplanar waveguide (CPW) output. This structure was presented earlier in sub-section 3.3.1 of this thesis, and was selected for the THz range approach since it delivers the highest power at this frequency range. The coupler circuit, shown in Figure 4-2(b), is a grounded-CPW to WR-3 transition that converts the QTEM mode into TE_{10} dominant mode. This is achieved by the means of a $\lambda_g/4$ long monopole that radiates vertically into the waveguide, with λ_g being the guided wavelength in the substrate. A 4-mm long WR-3 waveguide section leads to the module output which has a standard UG-387/U waveguide flange. The coupler circuit is fabricated on a commercial low-loss substrate Rogers/Duroid 5880 with a thickness of $130 \mu\text{m}$ and has an input impedance of 60Ω , including again an RF choke filter to enable biasing the PD which also helps to compensate the PD junction capacitance with the RF choke inductance. The packaged designed also allows to attach an optical fibre. The overall dimensions of the module are $50 \times 43 \times 15 \text{ mm}^3$. The package length is kept long, to accommodate the fibre assembly (Figure 4-2(a)).

Figure 4-2: (a) A general overview of 3D model of WR-3 waveguide-integrated PIN-PD THz emitter showing assembly details and internal structure. Inset: Structure for PD-to-WR-3 coupler integration. (b) Layout of the CPW-to-WR-3 transition circuit, along with an RF choke filter and holes for alignment and waveguide flange.

The designed passive circuit is simulated and optimised using a 3D electromagnetic wave solver. Figure 4-3 shows the simulated frequency response of the CPW to WR-3 coupler over the frequency range from 250 GHz to 320 GHz. A 3-dB bandwidth of 50 GHz is achieved. From the insertion loss (S21) graph we can notice a

minimum of 2.5 dB loss across the band that results primarily from the relatively higher substrate losses (0.2 dB/mm) at these frequencies. The length of the coupler section was kept to 10 mm to ease the packaging complexity and facilitate the optical fibre attachment in the module, which resulted in higher propagation losses. In addition, the reflection coefficient (S11) curve shows a low level of reflections.

Figure 4-3: Simulated S-parameters of CPW-to-WR-3 transition used for the PIN-PD THz emitter modules. (S11 – Reflection coefficient), (S21 – Insertion loss). The inset on right side shows the actual circuit used for simulate the performance.

The final packaged devices, THz-Emitter-1 and THz-Emitter-2, are shown in Figure 4-4(a) and Figure 4-4(b), respectively with an single-mode polarisation maintaining (PM) optical fibre pigtailed. The metal housing and circuit boards were manufactured at commercial facilities. The insets of the figures provide the internal views, showing the assembly details. The PIN-PD chip, $0.5 \times 2 \text{ mm}^2$, is connected with the passive circuit using two-wire bonds i.e. two wires for each contact pad. This proved to be a major drawback in the assembly, leading to high coupling loss that will be shown and quantified in the next sub-section. The initial idea was to use ribbon bonding technique that provides better RF coupling response compared to the thin wires, which could not be realised due to technical issues. The optoelectronic assembly and fibre

coupling were carried out in the packaging laboratory of Fraunhofer HHI. A feed-through filter is attached into the housing to the signal line that allows applying DC voltage.

The complete circuit layout of THz-Emitter-2 can be seen in the inset of Figure 4-4(b). The additional elements here include two RF choke filters (100 GHz and 300 GHz), an open-circuit coupled-line capacitor to provide different bias voltages V_{bias} to the active chips and a low-pass filter to prevent the leakage of 300 GHz signal into the LO path. A frequency multiplier chip, consisting of two anti-parallel varactor SBDs ($R_s = 7.3\Omega$, $C_j = 5.14$ fF), is flip-chip mounted using a silver-filled epoxy.

Figure 4-4: (a) Photograph of the packaged THz-Emitter-1 with a single mode optical PM fibre pigtailed. The insets below show the internal views of the module showing PIN-PD chip coupled to the WR-3 transition using two-wire bonds. (b) Packaged THz-Emitter-2 based on PIN-PD and frequency tripler units. The inset shows the passive circuit with an anti-parallel SBD tripler chip is mounted.

4.1.2 Characterisation

The measurements were undertaken in the THz Sensors and Systems laboratory at Fraunhofer HHI. The current-voltage (I-V) characteristics are obtained first to determine the working status of the modules after the customised packaging. The measured I-V curves at different optical input power levels P_{opt} are given in Figure 4-5.

Figure 4-5: Measured I-V curves at several optical input power levels for THz-Emitter-1 with a DC responsivity of 0.3 A/W and THz-Emitter-2. Very low current is obtained for the latter module due to misalignment of PM fibre post-attachment.

THz-Emitter-1 shows a DC responsivity (ratio of I_{ph} and P_{opt}) of 0.3 A/W at P_{opt} of 30 mW and V_{bias} of -2 V, indicating good optical fibre coupling. This is not the case for the second device, unfortunately, that shows an extremely low responsivity of 0.027 A/W. This severe degradation results from the misalignment of butt-coupled fibre after completing the packaging. Hence, the performance of THz-Emitter-2 could not be established and no comparison between the two approaches was possible.

The THz-Emitter-1 was the only one which could be further tested, to characterise the output power and frequency response. Figure 4-6(a) shows the laboratory setup for free-space absolute power measurement. The module is fed by an optical heterodyne optical signal, using two 1550 nm free-running ECL so that the generated frequency f_0 is equal to the difference frequency of the two optical wavelengths. A 20 Hz square wave signal is applied on the PIN-PD bias voltage for lock-in detection. As shown in the figure, a WR-3 horn antenna (with 25 dBi gain) is used to radiate the THz power, using a PTF sensor THz20HS (Sensor and Lasertechnik GmbH) [11] absolute power meter calibrated by PTB [12] to detect the generated power. The output voltage is measured using a lock-in amplifier. Since the aperture of the THz sensor is larger than that of horn antenna, both devices are placed face-to-face, including a 10° tilt to avoid the standing waves.

Figure 4-6(b) shows the measured output power spectrum of the module in the frequency range of 150 GHz to 350 GHz. The power is measured at I_{ph} of 8 mA and 12 mA, with V_{bias} of -1.2 V showing a peak power of -14.4 dBm (36.3 μ W) at 200 GHz. This is about the same power generated by the quasi-optic emitter module [13] using the same PD chip and using a monolithically integrated bowtie antenna. The subsequent gradual drop in the power is noticed, attributed mainly to the strong loss due to the 400 μ m long bond wires. However, relatively stable output, with variations in 3-dB limits, is obtained from 260 GHz to 290 GHz which was the actual target frequency range of the design. A power drop around 250 GHz comes from the cut-off of the choke filter, confirmed by the simulated response in Figure 4-3.

(a)

(b)

Figure 4-6: (a) Photograph of the experimental setup used to characterise the module THz-Emitter-1 using a PTF sensor. Measured frequency response of THz-Emitter-1 at $I_{ph} = 8$ mA and 12 mA, and $V_{bias} = -1.2$ V.

In order to quantify the RF coupling loss, we measured the frequency response of the wire-bonds. Two CPW lines are fabricated on an aluminium nitride (AlN) substrate, assembled in back-to-back and connected using two-wire bonds replicating the emitter sub-assembly, shown in the inset of Figure 4-7. The S-parameters were measured using a 110 GHz vector network analyser, showing the results for the insertion loss in Figure 4-7. The measured coupling loss reaches 8 dB at 100 GHz, and if we extrapolate to 300 GHz it increases to -24 dB. Considering this, we can safely expect that the output power of our emitter can be improved by 15 dB in the J-band when opting for flip-chip [14] or ribbon bond integration technique. In such a case, the THz emitter presented here has the potential to provide output performance comparable to that of the only existing WR-3 photomixer module [8].

Figure 4-7: Insertion loss of two-wire bonds connecting two CPWs on AlN substrate, measured up to 110 GHz. The response is extrapolated up to 300 GHz. Inset: Picture of the wire-bonded CPW assembly.

Finally, we investigate the influence of the optical power and bias voltage on the output power behaviour of THz-Emitter-1 by sweeping the parameters and measuring the radiated power. The P_{opt} and V_{bias} are varied from 20 mW to 60 mW and -0.5 V to -2 V, respectively and power measured using the setup shown above. Figure 4-8 shows the relative intensity of the measured power at f_0 of 200, 230, 280 and 330 GHz. It can be

seen that for all the frequencies, the RF power increases with the optical input power with no saturation visible. However, it depends strongly on the bias voltage, which is significantly noticed for higher frequencies where lower V_{bias} yields more power. This test gives the insight into the optimum conditions to use the emitter in practical applications.

Figure 4-8: Relative output power of THz-Emitter-1 as a function of optical input power and bias voltage at four difference frequencies.

4.2 Summary

In this chapter, two integrated THz photonic emitter modules were presented using different approaches. The first module directly generates the THz frequencies from a high-speed PIN-PD and the second one uses a novel true optoelectronic method by integrating photonic and electronic chips in one package. We demonstrated that by using PIN-PD as a 100 GHz LO for a frequency tripler chip based on Schottky diodes, generating 300 GHz signal. This approach eliminates the need of employing the costly THz amplifiers and can play an important role in the development of photonics-compatible high power THz sources. Due to the de-coupling of fibre in the packaged module, no experimental results were obtained for the thesis.

The first emitter module, however, was successfully tested using a pyroelectric power meter providing -14.4 dBm of radiated power. The high coupling loss of about more than 20 dB at 300 GHz resulted due to the use of wire-bonds. Nevertheless, the work presented here demonstrates the promising performance of PIN-PDs that can be compared with state of the art UTC-PDs. The main outcome of this work is the development of first ever WR-3 integrated PIN-PD emitter module.

4.3 References

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CHAPTER 5

Development of Photonic-enabled Phased Array Antennas

5.1 Introduction

Single-element antenna-integrated and rectangular waveguide based mm-Wave and THz emitters have been developed, to optimized the radiated power in the millimetre and terahertz wave ranges. In this chapter, we present the next natural step to develop a photonic-enabled phased array emitter leading to achieving beam-steering capability at the millimetre-wave range with application to broadband wireless communications.

In future beyond 5G cellular networks (also known as 6G), front/backhaul wireless access requires the transport of RF and baseband signals over optical fibre to the remote antenna sites to relax the complexity of radio-access unit (RAU) architecture at the RF front-end. In such application scenarios, beam-steering is a required functionality to mitigate the relatively higher path losses at mm-Wave and THz frequencies by providing direction-oriented coverage in point-to-multipoint systems and compensate for antenna misalignment in point-to-point links.

Photonics-based antenna-integrated emitters have been investigated and provided link performance of more than 100 Gbps [1], developing 2×2 photomixing emitter arrays in multiple input multiple output (MIMO) configuration at both the mm-Wave and THz frequency bands utilising commercial PD modules and antennas [2] - [4]. As we have discussed in CHAPTER 3, the common approaches to increase the radiation power is to either mount the photomixers on electrically large collimating lenses to increase the

received power or to couple them to bulky horn radiators. Such antenna structures can hardly be scaled into phased arrays and be produced at low cost. Nonetheless, different techniques have been reported in order to achieve beamsteering using the said antennas. These include from mechanically moving the photomixer position with respect to the lens focal point [5] to the development of linear horn arrays for phased array systems in W-band [7], [8] and J-band [9] frequencies. In the former solution, it was shown that a lens provides very limited steering capability when used in a phased array configuration [6] while in the latter, requires a complex split-block fabrication process. Needless to mention the extent of complexity and cost for 2D horn arrays.

Electronic beam-steering techniques on the other hand include phase shifters that are designed around a single frequency, and provide very limited bandwidth at mm-Wave/THz bands. Consequently, the presence of other frequencies in a wideband signal results in the beam-squint effect [10]; steering of the beam away from desired direction. Realising integrated emitter arrays for power scaling and beam-steering applications is hence a necessity and a challenge.

Optical beam-steering can be achieved using the True time delay (TTD) technique, which provides an effective alternative to the phase shifters. With TTD, all the spectral components present in a signal are delayed by equal amount of time ensuring that the radiated beams across whole bandwidth are steered by the same angle. A general system concept for TDD based beam-steering is sketched in Figure 5-1, which is based on creating as many copies of the signal as elements in the array, each of which has a programable time delay unit. Different attempts have been reported [7], [8], [11], demonstrating photonic TTD beam-steering with discrete off the shelf components.

Our approach has been to develop an integrated broadband photonics-based phased-array antenna (PAA) emitter based on the highly directive planar antennas developed for the single emitter modules presented before, operating in the 50 GHz – 120 GHz frequency range employing high-speed UTC-PDs, which were fabricated in

Figure 5-1: System concept for photonics-enabled true time delay based mm-Wave beamsteering.

linear array InP bars. By employing photonic TDD technique, mm-Wave squint-free beam-steering is demonstrated over the frequency range from 60 GHz – 70 GHz.

The rest of the chapter is organised as follows. Section 5.2 below presents the design details of the photonic emitter array module, together with the simulated performance of antennas. In sub-section 5.2, we cover the experiments conducted. The frequency response is measured by performing free-space radiated power measurements. Then in section 0 we demonstrate the beamsteering, using photonic TTD by performing radiation pattern measurements. Finally, a real-time multi-Gbps wireless link is established using the PAA module performing data transmission with beamsteering.

5.2 Photonic Emitter Array

5.2.1 Emitter Array Design

In the traditional method of packaging the photomixer emitters, there is little room for integrating custom antennas to achieve the desired performance and functionality. This limit is mainly imposed due to the nature of lens and horn antenna. We overcome this problem by developing a completely customised emitter design leading to simpler integration of PAAs. The developed integrated PAA emitter prototype is shown in Figure 5-2(a), packaged in a custom housing that consists of the fabricated passive circuit with

two antennas and bias network (bias-T), and a pair of UTC-PD chips fabricated in an InP bar at III-V Lab. A view of the passive circuit is shown in Figure 5-2(b). The coupling between the chips and the antennas is achieved using a silver-filled conductive epoxy, eliminating the inductance of thin wires at mm-Wave/THz frequencies. The pitch of passive optical waveguides (POW) in the chip is $500\text{ }\mu\text{m}$, using a 2×1 lensed fibre array fixed on a glass V-groove chip to couple the light into the UTC-PD pair, which have an active area of $4\times 15\text{ }\mu\text{m}^2$. Further details were presented in Section 3.2 of the thesis.

The antenna forms an important element of the mm-Wave/THz sources, increasing its complexity when aiming at realising phased arrays. We developed two

Figure 5-2: (a) The packaged photonic PAA emitter prototype with antenna and photodiode chips, inset: microscopic view of 2×1 UTC-PD array bonded to the antenna excited by lensed fibre array and (b) the layout of the designed passive circuit board including 2×1 tapered slot antenna array.

corrugated TSA antennas, based on the single-element TSA we presented in CHAPTER 3 [10], in 2×1 configuration with grounded CPW input. The overall structure of the antennas is shown in Figure 5-2(b) fabricated on a $130 \mu\text{m}$ thick commercial low-loss substrate Rogers/RT Duroid 5880 ($\epsilon_r=2.2$). The spacing d between antennas was kept to one wavelength λ_0 (4.2 mm) at central frequency f_0 of 70 GHz . To reduce the coupling complexity of fibre array, the number of antenna elements was limited to only two. However, good performance of beam-steering was achieved.

The simulated S-parameters of TSA array are given in Figure 5-3 over a wide frequency range from 50 GHz to 120 GHz , when excited simultaneously. The reflections (return loss) are less than -10 dB . A fence of thru vias help to keep the cross coupling better than -17 dB in most of the simulated frequency range. As can be seen, the antennas provide wideband performance, which is only limited by the RF choke.

Figure 5-3: Simulated S-parameters of the 2×1 TSA array.

The simulated gain of TSA array is better than 14 dBi over the frequency range of $70 - 100 \text{ GHz}$, shown in Figure 5-4(a), with a peak value of 15.9 dBi at 86 GHz . The simulated radiation patterns (gain) at frequencies of 75 GHz , 85 GHz and 100 GHz are plotted in Figure 5-4(b) for the E- and H- planes. The 3-dB beam-width and maximum

side lobe level (SLL) in E-plane are 18° and -17 dB, respectively. While in the H-plane, both the parameters increase to the 38° and -10 dB, respectively. The asymmetric beam profile here is due to the shorter width of single element as compared with the single TSA. There is no scaling of number of elements in the H-plane. Looking at the radiation performance, the advantage of the designed planar antenna is evident.

Figure 5-4: Simulated radiation performance of the 2×1 TSA array showing (a) gain vs. frequency and (b) radiation patterns in E-plane and H-planes at 75 GHz (red), 85 GHz (blue) and 100 GHz (green).

5.2.2 Radiated Power Performance

The performance of the dual antenna array was characterised by first measuring the responsivity of the integrated photodiode array varying the injected optical power P_{opt} and recording the DC photocurrent I_{ph} . The test is performed when accessing the PDs with a single fibre as well as using the 2×1 lensed fibre array mounted on a glass V-groove chip with 500 μm pitch. As in previous experiments, we expect around 3-dB optical coupling loss due to the 5 μm spot size diameter of the lensed fibre, which is double the recommended spot size of 2.5 μm . This experiment allowed us to assess the alignment tolerances for fibre array, which is one of the main challenges in optoelectronic assembly and packaging. The comparative responsivity is shown in Figure 5-5, where the insets show the photograph of the fibre-coupling in both cases. No bias voltage V_{bias} is applied to UTC-PDs. As can be seen, the responsivity curves are linear for both the cases. For single fibre, a responsivity of 0.2 A/W is achieved, being reduced to 0.063 A/W for the fibre array. This is believed to be caused by the inaccuracy in the pitch of V-groove chip. Nonetheless, both chips perform equally when excited individually.

Figure 5-5: Comparison of DC responsivity for the photonic emitter array at V_{bias} of 0V when excited using a single lensed fibre and dual-channel fibre. The insets on right show the microscopic pictures of fibre coupling to the photodiode chips.

The following characterisation test is to measure the generated RF power and bandwidth of the arrayed module. The experimental setup is sketched in Figure 5-6, showing that as in previous experiments, an optical heterodyne optical source is used to feed the UTC-PDs consisting on two free-running distributed feedback (DFB) lasers (Toptica Photonics), generating λ_1 and λ_2 , at wavelength of $\sim 1.5 \mu\text{m}$. The DFBs are tuned to generate the beating frequency f_{RF} in the range from 50 GHz to 120 GHz range. The DFB lasers' output is amplified by an EDFA and passed through an optical chopper to modulate the optical sub-carriers at a reference frequency of 12 Hz for lock-in detection. The generated and radiated mm-Wave power travels over a distance of 35 cm, using 90° off-axis parabolic mirror (focal length, FL=50 mm) and a 100 mm wide Teflon lens (FL=150 mm) to couple the radiation into an absolute THz power meter (TK sensor) at the receiving end. The power P_{rad} is calculated using (5-1), where V_0 is the output voltage of lock-in amplifier and R_{TK} being the absolute responsivity of the TK power sensor (0.5 V/W).

Due to the insufficient optical coupling for simultaneous excitation of two PDs (Figure 5-5), we measure the P_{rad} individually for each of the two antennas. An optimum V_{bias} of -3 V is applied to the UTC-PDs. Figure 5-7 shows the mm-Wave power, against the DC photocurrent on a logarithmic scale, measured at four different f_{RF} of 60 GHz, 75 GHz, 85 GHz and 100 GHz. The power radiated by Antenna-1 and Antenna-2 is plotted by the red and blue curves, respectively, while the black curve is the combination of both. We can observe a linear increase in the P_{rad} with respect to I_{ph} , due to the square-law behaviour of the photomixers. A correlation can be seen between the output power from both the antennas. However, minor fluctuations at lower I_{ph} are noticed, caused by fluctuations in the TK meter as the P_{rad} is close to the sensor's noise floor i.e. $1 \mu\text{W}$. The emitter provides the highest power at f_{RF} of 75 GHz reaching up to -9.7 dBm ($107 \mu\text{W}$) per element in the array. This is followed by 60 GHz; -10.2 dBm ($95.5 \mu\text{W}$). For higher frequencies, a degradation in the output can be seen, due to the decrease in RF isolation of bias-T. It is important to notice that no saturation of P_{rad} is experienced up to $10.5 I_{ph}$.

(a)

(b)

Figure 5-6: (a) Schematic diagram showing the experimental arrangement to characterise the integrated UTC-PD emitter array module using optical heterodyne and (b) laboratory setup for showing the lensed fibre array coupled to the emitter, transmission path and Thomas Keating THz sensor.

$$P_{rad} = \frac{V_0}{R_{TK}} \quad 5-1$$

This shows the high linearity of UTC-PDs used. In case of simultaneously exciting both the emitters in the array, a combined power of nearly -6.7 dBm (214 μ W) at 75 GHz (black curve) is achieved at I_{ph} of 10.5 mA.

Figure 5-7: Detected radiated power for antenna-integrated UTC-PD emitter array versus DC photocurrent at carrier frequencies of 60 GHz, 75 GHz, 85 GHz and 100 GHz. The bias voltage is -3 V.

The next step is to measure the frequency response of the emitter array module, shown in Figure 5-8 from 50 GHz to 120 GHz with a step size of 1 GHz step. The operating conditions for the PDs, I_{ph} and V_{bias} , are kept at 5 mA and -3 V respectively. It can be seen that both the emitter elements, Antenna-1 and Antenna-2, provide a broadband response, with a -3 dB bandwidth of about 45 GHz ranging from carrier frequency of 55 GHz to 100 GHz. In other terms, this translates to a fractional bandwidth of about 58%. To the best of our knowledge, it could be the widest 3-dB bandwidth achieved for a planar antenna-integrated photonic emitter, demonstrating the advantages of the our approach. Moreover, the response curves of both the antennas correlate very well, in terms of power variations, validating the free-space power measurement technique used. A gradual roll-off in the response is observed beyond 110 GHz, owing to the frequency cut-off of the TSA and bias-T. In Figure 5-9 we estimate the radiated power, P_{rad} , at 70 GHz, combining the emission for the two antennas. The curve fitted data is extended up to 16 mA, the saturation point I_{sat} of UTC-PD (see section 3.2.1) which allows us to estimate about -2 dBm of peak power can be obtained from the emitter array.

Figure 5-8: Relative frequency response of photonic emitter array with simultaneous excitation of UTC-PD measured at $I_{ph} = 5 \text{ mA}$ and $V_{bias} = -3 \text{ V}$ in the frequency range of 53 to 120 GHz. The 3-dB bandwidth is 45 GHz.

Figure 5-9: Combined P_{rad} versus I_{ph} for Antenna-1 and Antenna-2 measured for f_{RF} of 70 GHz at V_{bias} of -3 V. The plot is fitted and extended up to UTC-PD's I_{sat} of 16 mA.

5.3 Photonic Phased Array Antenna

5.3.1 Photonic TTD-enabled Beamsteering

Once the radiation performance of the UTC-PD emitter array has been established, we now demonstrate its beam-steering functionality, which was the primary motivation. The photonic TTD technique is exploited for this purpose. The experimental setup used to measure the radiation pattern is sketched in Figure 5-10(a), which shows again the optical heterodyne system to generate the RF frequency using a pair of narrow linewidth tunable ECLs that are modulated at 30 Hz for lock-in detection. This modulation takes place before the optical signal is split into two copies for each of the photodiodes. Different phase shift is applied to each copy by means of mechanically-tunable optical delay lines (ODL). The use of identical optical components help keeping the path lengths as close as possible to avoid the need for path length corrections. The time delay for the n^{th} antenna element in the array is τ_n . To steer the radiated beam by an angle θ from the boresight, we can use the expression:

$$\tau_n = n \frac{d \sin \theta}{c},$$

(a)

(b)

Figure 5-10: (a) Schematic diagram of the experimental arrangement for photonics-enabled beamsteering and (b) picture showing the laboratory setup for radiation pattern measurement in the E-plane.

where d is the spacing between the antennas and c is the speed of light. The optical signals are then fed into the PAA emitter, converted into RF by photomixing and radiated into free-space. Finally, a Schottky barrier diode detector detects the mm-Wave signal and the signal intensity is measured with a lock-in amplifier corresponding to a peak radiated power of less than -20 dBm. The whole setup is mounted on a rotation stage, as shown in Figure 5-10(b), to measure the radiation patterns and demonstrate the beam-steering. A hexapod allows to effectively align the 2×1 fibre array into the UTC-PDs, which are biased to fix I_{ph} to 3 mA with V_{bias} at -3V. The distance from the PAA to the Schottky receiver is set to 30 cm.

We validate the beamsteering behaviour of the PAA module by measuring the radiation patterns in far field in the azimuth (E-) plane in the range of -70° to $+70^\circ$. The angular resolution is set to 3° . The steering is performed at three different f_{RF} of 60, 65 and 70 GHz. Figure 5-11 shows the measured beam patterns for the pointing angles of boresight (0°) and $\pm 10^\circ$ achieved at 60 GHz and 70 GHz. For comparison, the simulated radiation patterns for the corresponding angles are also given. Note that SLL of less than -10 dB is maintained while performing the steering indicating the possibility of further scanning of the beam. This is confirmed by the measurements at 65 GHz, given in Figure 5-12. Here, the antenna radiation is further steered to an angle of 15° from the boresight. As a result, a continuous beam scanning of more than 20° and 30° can be achieved with the PAA, depending on the carrier frequency. The beam-steering also results, in the main lobe, a power reduction of 2 dB to 4 dB as the SLL increases due to appearance of the grating lobes. The amount of time delay in the two antenna elements was set to 0 ps and 2.4 ps for 10° , 2.4 ps and 0 ps for -10° , 0 ps and 3.6 ps for 15° and, 3.6 ps and 0 ps for -15° . The measurements performed in a benchtop environment with no RF absorbers in the surroundings. Hence, the contribution of reflection and scattering in the given data is expected, resulting in slight distortion in radiation patterns.

Figure 5-11: Beamsteering radiation patterns in the E-plane, simulated and measured, at the carrier frequencies of 60 GHz and 70 GHz for three different pointing angles. A total steering range of 20° is obtained.

Figure 5-12: Beamsteering radiation patterns in the E-plane, simulated and measured, at the carrier frequencies of 65 GHz for three different pointing angles. A total steering range of 30° is obtained.

5.3.2 Wireless Link with Beamsteering

Our last experiment with the PAA is aimed at establishing a real-time multi-Gbps wireless communications link which incorporates the ability to steer the radiated beam. The analogue beam-steering technique and the PAA emitter module presented in previous section are exploited for this purpose.

We use an amplitude modulation based experimental setup, that can be seen in Figure 5-13(a), with a picture of implemented setup shown in Figure 5-13(b). This arrangement is adapted from the one used for radiation pattern measurements (Figure 5-10(a)). The RF carrier frequency is fixed at 70 GHz. The output of EOAM, optical sub-carriers modulated by NRZ data of 2^{15} -1 word length provided by a PPG, splits into two paths and feeds the PAA emitter after passing through ODLs. The time delays, τ_n , are adjusted to provide the pointing angles of 0° , -10° and 10° for the emitted beam. Here, the transmission distance is reduced to 12 cm from 30 cm, due to reduced optical coupling with simultaneous excitation (see Figure 5-5). While the power available from the EDFA allowed to reach a maximum I_{ph} of 5 mA only. The bias voltage for UTC-PDs is kept unchanged at -3 V. In case of transmission with the beam steered away from the boresight, the transmitter is counter-rotated instead of the receiver due to practical limitations in the experimental arrangement. On the receiving end, a wideband zero-bias QO-SBD detector works as an envelope detector, down converting the data from the RF back into the baseband. The receiver output is amplified by an LNA with 30-dB gain and then fed into the BERT test equipment.

The wireless link experiment is performed at bitrates of 1.5 Gbps and 4 Gbps for the three beam angles, requiring a total of six measurements. For this purpose, we measured the radiated power P_{rad} beforehand at 70 GHz for simultaneous excitation of UTC-PD array using the setup discussed in Figure 5-6. As I_{ph} was limited to 5 mA, the maximum RF power that could be transmitted for data transmission was up to 27 μ W (-15.7 dBm) – about 12 dB less than emitter's peak output. It is important to mention that I_{ph} from each of the emitters was maintained identical throughout the measurements to

(a)

(b)

Figure 5-13: (a) Schematic diagram of the experimental setup for wireless communication link at f_{RF} of 70 GHz with beamsteering using photonic phased array antenna emitter and (b) a picture of implemented laboratory setup.

transmit equal amount of power from the antennas. The recorded BER responses are shown in Figure 5-14, plotted as a function of radiated power.

It can be observed wireless link with beamsteering is successfully established with the PAA emitter, as can be expected from the single emitter experiment in sub-section 3.2.2. For the case of 1.5 Gbps bitrate, an error-free ($BER < 10^{-10}$) transmission is achieved for all three beam angles. We can notice that as the antenna radiation steers

away from the boresight (0°), the required P_{rad} increases to maintain the BER. This power penalty is the cost to be paid for having introduced beam-steering, resulting from the drop in main lobe intensity and is consistent with the measured radiation patterns in Figure 5-11. The error-free link at the steering angles of 0° , 10° and -10° is achieved for P_{rad} of about -19.2 dBm, -18.2 dBm and -17.1 dBm, respectively. A similar trend is also noticed for of 4 Gbps curves. Here, the BER reaches error-free threshold for 0° at -15.5 dBm emitted power, while at 10° and -10° link response was limited by the available optical power. Nonetheless, BER rates well below the FEC limits are achieved for all cases. Taking into account the available RF power budget (-2 dBm), we believe longer transmission can easily be achieved at higher bitrates. The eye diagrams of demodulated received signals are shown in Figure 5-15.

Figure 5-14: BER as a function of radiated power P_{rad} for OOK-based wireless link with photonic TDD beamsteering angles of 0° , -10° and 10° at 70 GHz carrier frequency. The bitrates demonstrated are 1.5 Gbps (dashed curves) and 4 Gbps (solid curves).

The patterns were recorded for the error-free conditions at P_{rad} mentioned above showing open eyes for all six data transmission cases, with an increase in degradation at beam angles away from 0° . This effect is clearly noticeable for 4 Gbps.

Figure 5-15: Demodulated eye diagrams of OOK-based wireless link with photonic TDD beamsteering angles of 0° , 10° and -10° at 70 GHz carrier frequency for the bitrates of 1.5 Gbps and 4 Gbps.

5.4 Summary

An integrated photonics-based phased array antenna emitter for beamsteering was presented using linear two-element antenna and UTC-PDs, working in the frequency range of 50 GHz to 110 GHz. By designing broadband, low-profile and high gain planar antennas, the scaling into array was achieved and it became possible to develop the photomixer emitter without using Si lens and horn antenna. An overall 3-dB bandwidth of 45 GHz was obtained, while the experiments show that the emitter array can achieve a radiated power of up to -2 dBm. By employing the optical true time delay technique, the squint-free mm-Wave beamsteering was demonstrated, with scanning capability of at least 30°. Due to the large bandwidth of optical components, the problem of limited bandwidth of frequency-specific mm-Wave/THz phase shifters can be solved.

This development shows the potential of realising large arrays and 2-dimensional steering for microwave photonic systems. Wireless data transmission with beamsteering was successfully demonstrated using the developed PAA emitter module. The real-time error-free link provided bitrate of up to 4 Gbps with beam scanning of angles $\pm 10^\circ$.

5.5 References

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CHAPTER 6

mm-Wave and sub-THz Schottky Barrier Diode Detectors

6.1 Introduction

Up to this point, this thesis has mainly been dealing with the antenna-integrated photonics-based emitters for broadband wireless data transmission system. In such systems, in addition to the emitter for the transmitter side, needs a detector on the receiver end. In this chapter, we will now discuss our efforts to develop a suitable broadband receiver paired with the transmitters developed in the previous chapters.

Among the available technologies to detect mm-Wave and THz signals, Schottky barrier diodes (SBD) are widely considered for developing receiver devices [1]. These provide high sensitivity and fast switching speeds (up to 2 THz) at room temperature. During the last decade, researchers have reported several SBD-based detectors with promising results following two main approaches for their integration; rectangular waveguide (WR-SBD) and quasi-optic (QO-SBD) coupled. The WR-based receivers have always provided the highest sensitivity and baseband bandwidth[2] – [4], having the drawback that their operating frequency range is limited by the rectangular waveguide cut-off frequency. On the other hand, wideband-QO-SBD based detectors [5] – [8] provide the widest operating frequency range at the cost of lower detection sensitivity (due to large mismatch between the antenna and the chip impedances [7]) and higher noise equivalent power (NEP, 1.1 pW/ $\sqrt{\text{Hz}}$ for WR-SBD versus 12-30 pW/ $\sqrt{\text{Hz}}$ for the QO-SBD). These disadvantages can be address with the resonant-QO-SBD option in

which resonant narrowband antennas are used [6], providing greater control over the characteristic impedance to match it with the SBD, albeit limiting the operating frequency range. It has been reported that resonant-QO-SBD devices have achieved up to 19 GHz baseband bandwidths[9], while for wideband-QO-SBD the reported baseband bandwidths range from a few MHz to 5 GHz owing to the difficulty of designing antenna to the output connector interface. In these detectors, there is a delicate trade-off between sensitivity, RF bandwidth and baseband bandwidth. A novel type of receiver component has been recently developed, the Fermi-level managed barrier (FMB) diode, demonstrating a wideband detector integrating a bowtie antenna and a trans-impedance amplifier (TIA) [10], which achieved 11.3 GHz baseband bandwidth.

In order to achieve a detector that can be paired with our transmitters, we developed a wideband-QO-SBD detector with the aim to improved its performance in an incoherent real-time mm-Wave/THz wireless communication link. The device operating frequency range must cover from the mm-Wave E-band (71 to 86 GHz) and the sub-THz range (170 to 250 GHz), with a baseband bandwidth of 25 GHz. To the best of our knowledge, this would be the highest for a detector of this kind. With such detector, we develop a wireless link demonstrating data speeds up to 12.5 Gbps (limited by the available equipment and components), capable of operating at different carrier frequencies. To compare the performance of new design with a previous one, two devices with different antenna topologies and contact pads structure are developed and tested on the same setup.

This chapter is arranged in three sections. The design procedure, including antenna design, impedance matching network and device performance characterisation are detailed in Section 6.2. In Section 6.3, photonics-based ultra-wide carrier frequency tunable wireless link is established and demonstrated. Afterwards, in the Section 6.4, the developed detector is utilised in a wireless link to demonstrate a peak bitrate of 12.5 Gbps over two different waveguide frequency bands. Finally, a third QO-SBD with best baseband bandwidth, 25 GHz, is presented in Section 6.5.

6.2 QO-SBD Detector Design and Characterisation

6.2.1 Detector Design and Assembly

The desired performance for the detector is that its RF frequency range must cover the mm-Wave E-band (71 to 86 GHz) and the sub-THz range (170 to 250 GHz), with a baseband bandwidth of 25 GHz. These demanding characteristics require a careful design of the individual components before fabrication and assembly. The design will use the commonly employed planar antenna to couple the received RF power to the SBD chip. For the first detector developed in this thesis (Detector-1), we designed a linearly polarised bowtie antenna to be fabricated on a 0.5 mm thick high resistivity silicon (HR-Si) substrate, which has a relative permittivity (ϵ_r) of 11.7, yielding a 74Ω characteristic impedance of the antenna. Figure 6-1(a) shows the layout of the antenna, optimised after performing 3D electromagnetic (3D-EM) simulations to cover the frequency range from 30 GHz to 1 THz. The baseband received signal from the SBD chip is routed using a 74Ω coplanar strip (CPS) line. For a second detector (Detector-2), we used a log-spiral antenna, shown in Figure 6-1(b), from a previous relevant work conducted at UC3M [11].

Figure 6-1: Antenna layouts including contact lines used for QO-SBD. (a) Bowtie antenna for Detector-1 and (b) Log-spiral antenna for Detector-2.

In order to increase the directivity of the receiver, the antenna chip is mounted on HQ-Si hyper-hemispherical lens, as shown in Figure 6-2(a), where the size of the antenna

is exaggerated to give the right perspective view of QO antenna. The lens has a diameter d of 20 mm and an extended slab with a length L of 3.5 mm, shown in Figure 6-2(b). These dimensions were chosen after simulations and optimization to maximum the directivity and achieve Gaussian radiation beam. To reduce the discontinuity at antenna-lens interface and eliminate the excitation of higher order substrate modes, both the lens and the antenna use of same substrate material. In such structure, most of the radiated power is directed towards the dielectric lens due to the Silicon high ϵ_r . The ratio of the power emitted into the lens and the air (the two dielectric half-spaces) is given by the relation $\sqrt{\epsilon_{Si}/\epsilon_{air}}$ [12].

Figure 6-2: (a) 3D perspective of bowtie antenna on a HR-Si substrate placed over a hyper-hemispherical Si lens, with antenna feed point in inset and (b) side view of Si lens with dimension. $d = 20$ mm and $L = 3.5$ mm.

The radiation characteristics of the lens-mounted antenna are obtained by performing simulations using an in-house built programme, which combines the full wave simulation with ray tracing methods and traditional physical optics. Figure 6-3

details the achieved far-field performance for the bowtie antenna at carrier frequency of 70 GHz. In Figure 6-3(a), we plot the normalised intensity of E-field distributed over the surface of dielectric lens, effectively showing the radiation beam. The lens properly collimates the power and increases the directivity, while the ripples across the lens cross-section are the side-lobes. Figure 6-3(b) shows the 2D radiation pattern in the E-plane ($\Phi = 0^\circ$) and H-plane ($\Phi = 90^\circ$). The half-power beam-width (HPBW) is of the order of 10° and side-lobes in both planes well below -20 dB. Finally, Figure 6-3(c) shows the

Figure 6-3: Simulated radiation patterns of bowtie antenna at 70 GHz mounted on a hyper-hemispherical Si lens. (a) Intensity of E-field distributed over the lens surface, (b) 2D radiation pattern in E-plane ($\Phi = 0^\circ$, solid blue) and H-plane ($\Phi = 90^\circ$, dashed red) and (c) directivity versus frequency.

overall directivity of the lens antenna for the range of frequencies from 60 GHz to 110 GHz, which indicated that the directivity increases with frequency, since the lens becomes electrically larger with respect to the decreasing wavelength.

Considering the integration of the antenna with the diode, this is a critical step to avoid excess losses because unlike rectangular waveguide based detectors, no amplifiers can be introduced between the two components in QO receivers. The interface between the antenna and the diode can be represented by an equivalent circuit, as shown in Figure 6-4, in which the incoming RF signal is considered as a signal source, and the antenna establishes a voltage divider with the SBD load impedance. In such situation, to maximize the power transfer to the SBD, we must achieve impedance matching between the SBD and the antenna. This is usually referring to as matching factor (M-Factor) [13], which is in essence a ratio between the power coupled to the active chip (P_{SBD}) and the overall power available at the antenna (source, P_A) and can be represented as

$$M - Factor = \frac{P_{SBD}}{P_A} = \frac{4R_A R_{SBD}}{(R_A + R_{SBD})^2 + (X_A + X_{SBD})^2} \quad (6-1)$$

where, the load impedance is $Z_{SBD} = R_{SBD} + jX_{SBD}$ and the source impedance is $Z_A = R_A + jX_A$, show by lumped elements in Figure 6-4. The maximum power transfer condition is that $Z_A = Z_{SBD}^*$, in which case the M-Factor equals unity. Such a matching can be easily performed in the case of narrowband detectors, since the input impedance of resonant antennas (both the resistance and reactance) can be controlled by design parameters. However, the wideband antennas reduce such a possibility due to the difficulty to control the impedance over extremely broad frequency range. Nonetheless, it is important to perform the antenna-SBD co-simulations to determine the M-Factor and choose the diode with right parameters.

The diodes used in this work are zero-bias low-barrier SBD, manufactured by ACST GmbH [14]. The chip is transferred on to a 5 μm thick gallium arsenide (GaAs) substrate for assembly. The low Schottky barrier height results in very low differential

resistance R_d , which eases the matching with antennas and other $50\ \Omega$ components. In addition, low junction capacitance C_j helps achieve RC-limited bandwidth of more than 1 THz. Table 6-1 summarizes the parameters for ACST 3DSFx type diodes which have a 3750 V/W intrinsic voltage responsivity. These parameters are used to calculate the M-Factor for the detector. The bowtie antenna is excited by a discrete port, as shown in the inset of Figure 6-2(a), with a lumped element at the feed point that emulate the SBD lumped parameters; junction capacitance C_j and differential resistance R_d .

The dielectric lens-mounted bowtie antenna is simulated over wide frequency range of 50 GHz to 300 GHz. We performed the simulation for all the four diode types to study the performance, to do the comparison and select the right diode chip. In Figure 6-5 we plot the achieved the M-Factor over a wide frequency band, ranging from 50 GHz to 300 GHz. It can easily be notice that the highest matching achieved is for diode 3DSF30, owing to the low value of R_d . The M-Factor peaks to 0.48 at 50 GHz, and then drops subsequently as the carrier frequency increases. Similar drop is also noticed for other diodes. This is because the antenna feed point impedance becomes more capacitive with respect to increasing frequency. The M-Factor starts to rise at 200 GHz, reaching 0.4 at 300 GHz. The lowest performance is observed for 3DSF5 for which less than 10% of the power can be coupled to the diode, due to very high R_d . However, it is more stable as compared to other diodes. Since our application is mainly at mm-Wave frequencies, we select 3DSF30 diode chip for our detectors.

Figure 6-4: Equivalent circuit of the QO-SBD for impedance matching between the diode and the antenna.

Table 6-1: Parameters of ACST's Low-Barrier Schottky Diodes [14].

Diode Type	Junction Capacitance C_j (pF)	Differential Resistance R_d (Ω)
3DSF5	5	2500
3DSF10	10	1000
3DSF20	20	500
3DSF30	30	300

Figure 6-5: Simulated M-factor of dielectric lens-mounted bowtie antenna for different low-barrier SBDs available at ACST GmbH.

6.2.2 Baseband Matching Network

Once the proper SBD is selected, we now consider the baseband circuit. Usually, the designers of THz detectors are focused on RF performance of the device in order to improve the sensitivity and NEP. However, for high speed data communication applications, it is also important to optimise the baseband performance. This is even more significant especially in the case of wideband QO detectors.

We designed a wideband impedance matching network on a 0.5 mm thick commercial RT/Duroid 5880 low dielectric substrate. Figure 6-6 shows the complete structure of the device assembly of Detector-1. For this module, we have implemented the baseband with a mode-matching circuit that converts the balanced CPS mode of the antenna output into the unbalanced coplanar waveguide mode with $50\ \Omega$ impedance. This provides a smooth transition, hence increasing the matching and bandwidth for extracting down-converted currents. Here, a video amplifier can be easily added to boost the received data signal. The circuit is optimised with a 3D-EM simulator.

Figure 6-6: Overall 3D schematic of the QO-SBD detector with bowtie antenna, integrated with a baseband matching board. The board implements a CPS-to-CPW transition to extract the baseband currents from the antenna to the output coax connector, providing impedance matching between the two. The inset shows the antenna chip placed at the centre of the board. Materials: Gold metal (yellow), HR-Si substrate, lens (gray), and Duroid 5880 substrate (maroon).

For validation purposes we fabricated a back-to-back configuration of the baseband transition, shown in Figure 6-7(a). In Figure 6-7(b) we present the simulated and measured S-parameters from 1 KHz to 30 GHz. The measured insertion loss (IL, S_{21}) shows its -3 dB cut-off frequency around 20 GHz. Since the back-to-back sample incorporates two of such transitions, this means that we have achieved 40 GHz bandwidth for a single transition. The return loss (RL, S_{11}) is fairly low here. The bandwidth mainly

depends upon the length of CPS line and quality of wire bonds. Longer CPS section leads to phase mismatch at higher frequencies and reduction in bandwidth. Nonetheless, this makes a perfect circuit for our detector and the bandwidth achieved is enough to satisfy the requirements.

(a)

(b)

Figure 6-7: (a) Fabricated and connectorised back-to-back baseband impedance matching board and (b) simulated (black) versus measured (red) S-parameters.

After manufacturing all the required components, the final assembly of the two SBD detectors was carried by ACST. Figure 6-8 shows the photographs of the final packaged modules, including the bowtie and log-spiral antennas with flip-chip mounted diodes. The Si lenses used here have no anti-reflection coating, so we expect about 1.5

dB power loss for the incoming wave due to the reflections at air-dielectric boundary. A K-connector is mounted on the device for the output.

Figure 6-8: Photographs of wideband QO-SBD detectors. (a) Packaged and connectorised module, (b) internal view showing the details of device assembly structure along with outer dimensions, (c) microscopic image of bowtie antenna chip for Detector-1 with 3DSF30 diode chip mounted at the feed point and (d) log-spiral antenna chip for Detector-2 with 3DSF30 chip.

6.2.3 Responsivity and Baseband Bandwidth Measurements

We now evaluate the performance characteristics of QO-SBDs by first measuring the frequency response over the entire RF frequency range. The setup is shown in Figure 6-9, where the transmitted signal is generated by a commercial THz system from TOPTICA Photonics AG, which includes a PIN photodiode based THz emitter to generate carrier frequencies, f_{RF} . The absolute THz power falling at the detector could not be measured at the time of this experiment. Therefore, we establish the relative voltage sensitivity of the detector measuring the output voltage V_0 at the lock-in amplifier.

Figure 6-10 shows the measured normalised response for both detectors, on a logarithmic scale from 50 GHz to 800 GHz frequency span with step size of 1 GHz. Moreover, the figure also shows the measured noise floor of the detector used to calculate the dynamic range (DR), an important parameter that reveals the actual measurement bandwidth of the detector. The DR, shown by the dashed curve in Fig. 1(a), is calculated as the ratio of maximum signal amplitude E_{max} and the root-mean-square of detector noise N_{rms} , and is given as

$$DR = \frac{E_{max}}{N_{rms}} . \quad (6-2)$$

Figure 6-9: Block diagram of the experimental setup for voltage responsivity and baseband bandwidth measurement.

We find that the maximum sensitivity obtained is at 75 GHz, with a DR of 20700 (86 dB) for Detector-1 and 10000 (80 dB) for Detector-1, which occurs as a result of local peak in the output power of the Tx. From there onwards, it decreases almost linearly until 350 GHz. While a moderate sensitivity is achieved up to 300 GHz with DR of 45.48 (16.57 dB) for Detector-1, the signal of Detector-2 meets the noise floor. The overall

(a)

(b)

Figure 6-10: Measured performance characteristics of the QO-SBD envelope detectors showing relative voltage responsivity, noise floor and calculated dynamic range. (a) Detector-1 with bowtie and (b) Detector-2 with spiral antenna.

bandwidth for first device, whose overall sensitivity is better than the second one, goes up to 500 GHz. While intrinsic cut-off frequency of the Schottky diode used in this work is well beyond 1 THz [14], the primary cause limiting the total bandwidth of developed device here is the exponential roll-off in the output power of the emitter used.

Next, we evaluated the baseband performance of detector. The experimental setup used here is same as in earlier case, setting the optical wavelength spacing at the carrier frequency f_{RF} where the sensitivity peaks (75 GHz) and introducing a sine wave modulation with an optical modulator driven by a CW frequency synthesiser. The signal at the SBD is monitored with an electrical spectrum analyser, tracing the received amplitude as the frequency of the sine wave is varied from few kHz to 20 GHz. Figure 6-11 shows the measured baseband signal bandwidth, normalised to the lowest frequency point in the graph. As can be seen, for Detector-1 the 3-dB bandwidth is 10.5 GHz and that of 6-dB is 13 GHz. While for Detector-2, it is 7.5 GHz and 8 GHz, respectively. Again, we see that bowtie antenna-based device has better performance, primarily due to the optimised CPS lines for the output. This is expected to be the bandwidth across the whole RF range in Fig. 1(a), since the baseband spectral response of an envelope detectors is independent of the carrier frequency. It must be mentioned that a 10 GHz modulator, used in this experiment, and thin bond wires lead to a drop in the response that can be observed in the plot after 10 GHz. Higher bandwidth for Detector-1 is expected otherwise.

(a)

(b)

Figure 6-11: Measured baseband signal bandwidth for QO-SBD detectors at f_{RF} of 75 GHz. (a) Detector-1 with bowtie and (b) Detector-2 with log-spiral antenna.

6.3 Wireless Communications with Ultra-wide Frequency Tuning

After successful development and characterization of the detector, we have introduced this device in our wireless data transmission test-bed as a receiver to demonstrate a flexible ultra-wide frequency tunable wireless communication system utilising QO emitter and receiver modules. The experiments feature a 1.5 Gbps wireless link for data and high definition television (HDTV) transmission over six different mm-Wave frequencies across different bands, showing the results at 60, 78, 100, 140 and 180 GHz carrier waves generated and detected using the same emitter-receiver pair. This range covers a total of 120 GHz range and spans across three continuous waveguide bands, from the E-band (60 – 90 GHz), F-band (90 – 140 GHz) and up to the D-band (110 – 170 GHz).

The first sub-section covers the details of experiment and results achieved. Afterwards, we perform a link-budget analysis to determine the transmission range of such a wide-band system.

6.3.1 Ultra-wide Frequency-tunable Wireless Link

The overall arrangement of the experiment and a picture of actual setup are shown in Figure 6-12. On the transmitter side, the optical carrier generation is achieved by optical heterodyning comprising two tunable external cavity lasers (ECL, Sacher Lasertechnik). This provides two wavelengths λ_1 and λ_2 spaced by the required frequency f_{RF} . The two optical tones are modulated with data by an electro-optic amplitude modulator (EOAM, Thorlabs LN81S-FC). We use double side-band modulation, since it provides better signal to noise ratio (SNR) [15]. The digital data stream is provided by a pulsed pattern generator (PPG) with a word length of 2^{31-1} . The output from the modulator is boosted using an erbium-doped fibre amplifier (EDFA). To generate and radiate the mm-Wave signals, we use a state-of-the-art high-speed PIN photodiode (PIN-PD) developed by Fraunhofer HHI [16], with which this work was a collaboration. The PD features a monolithically integrated extended bow-tie antenna mounted on a 10 mm wide

Si lens, and operates from 50 GHz to 2 THz. The continuous wave output power of the module is up to 300 μ W and 3 μ W at 300 GHz and 1 THz, respectively. The bias voltage is kept -1.5 V, while the optical power P_{opt} injected is varied from 2 mW to 30 mW. For the real-time HDTV transmission experiment, the setup remains the same. A computer provides the HD video (1920 x 1080) signal, which is fed into the EOAM after converting into non-return-to-zero serial data using an HDMI to SDI converter.

(a)

(b)

Figure 6-12: Block diagram of the photonics based real-time wireless data transmission link at 1.5 Gbps using single quasi-optic emitter and Schottky detector pair at carrier frequencies, f_{RF} , of 60 GHz, 78 GHz, 100 GHz, 120 GHz, 140 GHz and 180 GHz and (b) experimental arrangement showing the modules and transmission path.

On the receiving side, the key element is the QO-SBD (Detector-1, Rx) used as an envelope detector. The demodulated signal is amplified by a low noise amplifier (LNA, DC ~ 1.5 GHz) with a 30 dB gain. A low pass filter (LPF) removes any harmonics generated by EOAM due to possible fluctuations in DC bias. After reshaping the signal with a limiting amplifier (LA, 32 GHz), it finally goes into a bit error-rate tester (BERT). For HDTV transmission experiment, the video signal is recovered through an SDI-to-HDMI converter and fed into a screen. The link distance is 0.9 m, limited by the available space. The transmission path consists of two plano-convex lenses with 50 mm diameter to collimate the radiation, providing 14 dBi of gain each [18].

For the transmission of 1.5 Gbps data and HDTV signal, the f_{RF} is swept from 60 GHz to 180 GHz and bit error rate (BER) recorded. Figure 6-13 plots the series of BER characteristics for versus P_{opt} . It can be seen that error-free ($BER < 10^{-10}$) link is achieved across the whole 120 GHz frequency range. As the f_{RF} increases, a moderate increment in P_{opt} is required to maintain the link error-free, despite the fact that the antenna directivity of Tx and Rx increases. This is mainly due to the power roll-off in the emitter. The 180 GHz curve appears to be an exception, requiring two times the input power of 140 GHz. The reason in this case is the frequency dependence of the detector and non-uniform antenna radiation patterns of the Tx module, resulting in less power being collimated as the beam profile becomes multi-lobe. In Figure 6-14 we can see clean and open eye diagrams at all the f_{RF} points. Figure 6-15 is the picture of HDTV experiment. While changing the f_{RF} during transmission, no interruptions were observed.

Figure 6-13: BER characteristics for the ultra-wide frequency tunable real-time 1.5 Gbps wireless link at six carrier frequencies, f_{RF} , with a span of 120 GHz.

Figure 6-14: Eye diagrams of demodulated data at 1.5 Gbps real-time wireless link for f_{RF} of 60 GHz, 78 GHz, 100 GHz, 120 GHz, 140 GHz and 180 GHz.

Figure 6-15: Picture of the real-time wireless HDTV video transmission. The signal, obtained from a source (right), is transmitted and displayed on a screen (left) after detection.

6.3.2 Link Budget

In order to investigate the maximum link distance that can be achieved, we have performed a link budget analysis. For this, we have first measured the receiver sensitivity at the different carrier frequencies f_{RF} for the error-free transmission conditions as well as for pre-FEC BER (10^{-3}) condition. We achieve this by measuring the radiated RF power at the receiver location using a calibrated absolute power meter (TK-meter), while keeping the same levels of input power into the Tx and other conditions (such as bias voltage, link distance, etc). The PIN-PD emitter module bias voltage is superimposed with a 12 Hz square modulation for providing an electrical chopping signal for lock-in detection. The received power is measured at both BER cases and will be used to calculate the link distance. We also measured the available peak power at the emitter, carefully placing the TK power meter at its output, avoiding creating a standing wave with the back-reflections. Table 6-2 summarises all the measured power levels, as well as the simulated antenna directivity for both the Tx and Rx. Since the minimum sensitivity of TK is $1 \mu\text{W}$, the lower power levels given in the table are estimated by subtracting the path loss from the peak emitter power at corresponding P_{opt} .

Table 6-2: Measured receiver sensitivities for error-free and pre-FEC BER cases and simulated antenna directivities.

		Carrier Frequency, f_{RF} (GHz)					
		60	78	100	120	140	180
Receiver Sensitivity (dBm)	BER < 10^{-10} (Error-free)	-30	-29.2	-28.98	-20.9	-15.9	-20.08
	BER = 10^{-3} (FEC)	-36.98	-35.2	-34.25	-26.57	-24.6	-25.22
Antenna directivity (dBi)	Transmitter	29	30	31.1	32	33.2	34.5
	Receiver	34	35	37	38	39	41

With this data in hand, the theoretical link distance for the proposed system can now be determined using by using Friis transmission formula:

$$P_r = P_t + G_t + G_r + 20 \log_{10} \left(\frac{c}{4\pi r f_{RF}} \right) \quad (6-3)$$

where P_t and P_r are the transmitter power and receiver sensitivity, G_t and G_r are the simulated directivity of Tx and Rx antennas, respectively, as given in the table above, c is the speed of light and r is the transmission distance. We would like to highlight that the values for the antenna gain parameters include the 14 dBi measured gain of the Teflon lenses that are included, as observed in picture of setup in Figure 6-12. In Figure 6-16 plots for each of the six f_{RF} the required transmitter power versus the link distance. A dashed line shows the maximum power available from the PIN-PD emitter. We find that the error-free link maximum distance is 21 m, achieved at 60 GHz (higher QO-SBD sensitivity and emitter power), while for the FEC case, this increases to more than 30 m. At 78 GHz and 100 GHz, decent ranges of 11.5 m and 7.5 m, respectively, which can be increased to 30 m when forward error correction algorithms are used. Shorter distances are observed for rest of the three f_{RF} , particularly due to the frequency dependence of Tx and Rx response. In general, the overall link distance can be extended by twice by opting for pre-FEC BER. The model used here does not consider the atmospheric attenuation.

Figure 6-16: Theoretically calculated required transmitter power over the link distance to achieve error-free ($\text{BER} < 10^{-10}$, solid line) and pre-FEC ($\text{BER} = 10^{-3}$, dotted line) wireless transmission at six carrier frequencies, f_{RF} , from 60 GHz to 180 GHz. The measured maximum available Tx power is also indicated (dashed line).

6.4 Multi-frequency High Speed Wireless Link

In this section, we demonstrate the ultra-wideband receiver performing error-free wireless link up to 12.5 Gbps at two different carrier frequencies: 75 GHz (f_{RF1} , E-band) and 120 GHz (f_{RF2} , F-band). Figure 6-17(a) shows the experimental setup for the real-time multi-Gbps and multi-frequency wireless link. The optical carrier generation remains same as in previous section (Figure 6-12(a)). To generate and transmit the RF signals, we use two different commercial photodiodes (PD): (a) a u2t XPDV4120 PIN-PD, connected to WR-12 horn antenna and, (b) an NTT IOD-PMF-10001 uni-traveling carrier photodiode (UTC-PD), connected to WR-12 and WR-08 horn antennas.

Figure 6-17: (a) Block diagram of experimental configuration used to test the QO-SBD envelope detector in a photonics-based real-time error-free wireless communications link setup at transmission frequencies of 75 GHz and 120 GHz, photographs of actual setup for (b) 75 GHz and (c) 120 GHz.

The demodulated signals from QO-SBD are amplified, shaped by an LA and then fed into a BERT for the analyses. The link distance is set to 0.5 m, limited by the laboratory arrangement. Figure 6-17(b) and (c) show the pictures of laboratory setups for 75 GHz and 120 GHz, respectively.

The link performance in terms of BER vs. photocurrent (I_{ph}) for f_{RF1} and f_{RF2} are shown in Figure 6-18 (a) and (b), respectively, at data rates of 5 Gbps and 12.5 Gbps. As can be noted, the error-free transmission ($BER < 10^{-10}$) is achieved for both the frequencies and maximum available bitrate. This validates the measured baseband bandwidth of the detector. Moreover, the FEC-limit BER of 10^{-3} is achieved at a much lower I_{ph} compared with error-free transmission, indicating that a lower radiated power is required. One interesting point that must be mentioned here is that for 120 GHz frequency, a larger separation is observed between the two BER curves when compared to the 75 GHz case. This shows the frequency dependence of the receiver sensitivity, which is 10 dB lower in the latter case, implying that increase in the transmit power is needed to achieve error-free transmission for higher frequencies and data speeds. The insets of Figure 6-18 show the eye diagrams for the case of error-free transmission at 12.5 Gbps at both f_{RF} . The eyes are wide open and clear. While the measured RMS jitter for both the transmission experiments was between 7 to 8.5 ps. These factors indicate that further increase in data rate is possible by increasing the bandwidth of the EOAM and PPG used.

Figure 6-18: Wireless link BER characteristics for bitrates of 5 Gbps and 12.5 Gbps using (a) PIN-PD at f_{RF1} (75 GHz) and (b) UTC-PD at f_{RF2} (120 GHz). Eye diagrams of demodulated signal at 12.5 Gbps for $BER < 10^{-10}$ are in the inset.

6.5 Detector-3: Improving Baseband Response

Based on the above results, we carried out an additional iteration of the SBD receiver, developing a third detector (Detector-3) with improved design and performance. The new design is based on Detector-1 (Bow-Tie antenna type), for which the baseband bandwidth was limited to 10 GHz when the impedance matching network was designed to provide a baseband bandwidth in the excess of 30 GHz.

For Detector-1 we found by 3D-EM simulations that the bandwidth limitation occurred from the 90° angular transition between the bowtie antenna and the CPS contact lines, which is more critical at higher baseband frequencies. To overcome this problem, we designed the extended bowtie antenna shown in Figure 6-19(a). Unlike the linear or exponential taper of the bowtie arm, the extended bowtie features a non-tapered extension of short tapered arms. Thus, the demodulated high frequency currents experience a relatively smooth transition from the antenna to the CPS contacts. In addition, we address the losses caused by the wire-bond between the antenna chip and matching board by performing a three-wire interconnection. Figure 6-19(b) shows the picture of the detector assembly, highlighting the wire-bonds.

Figure 6-19: (a) Microscopic image of extended bowtie antenna chip for Detector-3, with 3DSF30 diode chip mounted at the feed point and (b) a picture of device sub-assembly.

We characterised Detector-3 to measure the baseband response, using the setup described in sub-section 6.2.3 Here, a rectangular waveguide-integrated PIN-PD with WR-12 horn antenna is employed as the source. The resulting response is plotted in Figure 6-20, up to frequency of 30 GHz. For comparison, the simulated baseband loss response is also given. Again, as in the case of previous detectors, this measurement is taken at carrier frequency of 75 GHz. The measured 3-dB bandwidth of 25 GHz is obtained for the modified detector. The simulated curve, however, shows that the actual bandwidth could be much higher than 30 GHz. The limit is believed to be caused by non-ideal bonded wires.

Figure 6-20: Baseband signal bandwidth for Detector-3 at f_{RF} of 75 GHz, showing the comparison between simulated and measured response.

Finally, we provide a performance comparison of the proposed detector against a commercially available SBD module. Figure 6-21 shows the measured baseband characteristics for Detector-1, Detector-3 and a WR-10 integrated SBD [19]. All the measurements were performed using the same source and test setup. The acquired data are also corrected for the roll-off in the frequency response of PIN-PD module, as well as

the loss of the RF coaxial cables used. We can see that the new detector provides more than twice the bandwidth of previous receiver. Detector-3 also offers a broader response than the commercial SBD, confirming the significant advantage of the SBD-to-output connector matching technique presented. The obtained results suggest that the develop receiver can support bitrates of over 25 Gbps in an OOK-based data link in various mm-Wave and THz frequency bands.

Figure 6-21: Baseband bandwidth comparison for the custom-made QO-SBD receivers, Detector-1 and Detector-3, and a commercial WR-10 SBD detector.

6.6 Summary

In this chapter, we focused on the development of envelope detectors based on high speed low-barrier Schottky diode technology, specifically for the application of high speed wireless communications. An overview of wideband QO-SBD detectors was given and key features were identified, along with the main issues. A new design was proposed to solve the inherent low baseband problem and complete characterization carrier out.

The first section was devoted to the procedures for designing, simulating and developing of wideband SBD detectors. In order to improve the baseband bandwidth, we proposed an impedance matching technique. This involved the designing of a bowtie antenna with optimised contact pads, in order to provide a smooth transition from the SBD chip to the 50Ω ohm output connector. To compare its performance, we developed a second detector with a log-spiral antenna with traditional contact structure. Two detectors were manufactured and packaged. The device characterisation included the measurement of device responsivity in mm-Wave to THz range, as well as the baseband performance. As expected, a significant improvement was recorded, with 10.5 GHz of bandwidth that turned out be better than the log-spiral design. Moreover, a third detector was developed to further improve the performance. A record baseband bandwidth of 25 GHz was demonstrated, surpassing a commercial SBD receiver. This could be the highest bandwidth reported for QO-type detectors.

In the second and third section, we experimentally verified the detector performance. This included the first time demonstration of ultra-wide frequency tunable wireless communication link, with a continuous 120 GHz tuning range, using only a single pair of transmitter and receiver. Then, the detector was successfully used in a multi-frequency high speed error-free link with bitrate up to 12.5 Gbps. To the best of our knowledge, this is among the highest speed achieved by a wideband QO detector.

6.7 References

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CHAPTER 7

Conclusions and Outlook

7.1 Conclusions

Future wireless communication systems (known as Beyond 5G, or 6G networks) are expected to operate within the mm-Wave and/or THz frequency bands of the electromagnetic spectrum. The availability of efficient and small-footprint transmitters and receivers are key components to enable it, including a much needed convergence of RF links with existing optical network. Photonics-enabled RF devices are essential in this regard. The aim of this research work was to develop the antenna-integrated mm-Wave and THz sources and detectors with broad bandwidth, low production cost and small size. The research activities conducted in this thesis were mainly within the context of a Marie Skłodowska-Curie Innovative Training Network (MCSA-ITN) FiWiN5G (Fiber-Wireless Integrated Networks for 5th Generation delivery) project which ran from January 2015 to December 2018.

For the sources, we studied the use of planar antennas for the integration with photomixer chips (UTC-PD and PIN-PD) fabricated on III-V semiconductor substrate. It was demonstrated that how the high directivity and low-profile planar antennas can be appealing alternative to the electrically large and costly lens and horn radiators with comparable radiation performance. It was also shown that such planar antennas can be made broadband, meeting the system requirements. Low-profile antennas, in combination with the simple optoelectronic integration, provided the added advantage of device scalability in the integrated modules. This was proven by the development of a phased antenna array emitter, offering the advantage of power scaling and most importantly the

functionality of beam-scanning in the E-band. The optical components such as delay lines and phase shifter are especially advantageous in providing ultra-broadband and squint-free beam-steering in the THz range.

The work related to the sources was extended to the THz frequency range. Here, rectangular waveguide integrated emitters in the J-band frequency were developed using high-speed and high-power p-i-n photodiode chips. The aim was to demonstrate that the THz PIN-PDs can potentially provide comparable performance to UTC-PD devices. The limitations in optoelectronic integration technique, however, resulted in significant amount of RF power loss. Moreover, a second THz emitter based on a PIN-PD and a Schottky multiplier diode was designed. With this, we presented the unique study for the photonic-electronic co-integration in a single THz module. For future THz systems, such a strategy can be beneficial by combining the advantages of both technologies. Most importantly, this allows to integrate a low-cost mm-Wave LNA to generate THz wave. The major outcome of this work is the first ever development of a WR-03 integrated PIN-PD emitter module.

The scope of this thesis also included the development of and demonstration of wideband and compact detectors. Three different designs of QO-SBD detectors have been presented, with special attention to the baseband bandwidth, essential for high data rate links. A new antenna with contact structure and impedance matching network allowed to increase the 3-dB baseband response up to 25 GHz, which to the best of our knowledge is the widest bandwidth achieved for a wideband QO-type detector operating over such a wide RF frequency range spanning from the E-band into the sub-THz range. On the other hand, the ultra-broad RF response of the presented detector allows for truly leveraging the broadband nature of photonic sources. At the end, a wireless link with ultra-wide tunability of RF carrier frequency was also established in this respect.

7.2 Future Work

In the work presented in this thesis, one of the primary factors limiting the performance of the mm-Wave and THz signal sources was the RF losses in the optoelectronic integration. Moreover, the use of traditional commercial substrates, with high losses, is not the optimum solution for the high frequency devices. Therefore, it is necessary to come-up with alternative methodologies to facilitate the THz antennas and integration of photonic and electronic components. We propose some future research tracks that could potentially address the above mentioned limitations. Initial studies and experimental work is already performed in this context relying on hybrid integration, which is probably the integration technology of the future, enabling to mix different substrate materials. Hybrid integration of low-loss and low-permittivity polyimide substrates is a key technology for efficient RF interposer boards. We are currently designing THz antennas on a polyimide substrate in collaboration with Fraunhofer HHI Hybrid Integration Group using their PolyBoard platform.

In addition, we are also exploring the use of the polyimide¹ ($\epsilon_r = 2.7$, $\tan\delta = 0.002$) is an optical material that can be used as interposer to facilitate both the optical and RF interconnections in the assembly of microwave photonic front-end devices.

¹ Moritz Kleinert, David de Felipe, Crispin Zawadzki, Walter Brinker, Jung Han Choi, Philipp Reinke, Magnus Happach, Simon Nellen, Martin Möhrle, Heinz-Gunter Bach, Norbert Keil, Martin Schell, “Photonic integrated devices and functions on hybrid polymer platform,” Proceedings of SPIE 10098, Physics and Simulation of Optoelectronic Devices XXV, 100981A, 2017.

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