

Physical Layer Design of Terahertz Band 5G Wireless Communications

by

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Physical Layer Design of Terahertz Band 5G Wireless Communications

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This is to certify that I have examined this copy of a master's thesis by

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*To my lovely mother Farzana Khalid for all the sacrifices she made in her life to
educate me and make me a stronger man.*

To my lovely sister Shehla Khalid for being so sweet and loving.

ABSTRACT

The emerging technology Terahertz Band (0.3 - 10 THz) communication is envisioned to accommodate high speed wireless communication. Large bandwidth makes it a good candidate for 5G mobile networks as it can alleviate the problem for spectrum congestion. In this thesis, we propose the first ever physical layer design of communication in THz Band not only for single-input single-output (SISO) based networks but also multiple-input multiple-output (MIMO). We first introduce the testbed system and present its use for carrying fundamental experiments in THz Band. Path loss and phase delay measurements from 260 GHz to 400 GHz for different distances, angles of arrival and objects acting as reflectors are then discussed. We have shown links reaching speeds of terabits per second that confirms THz Band as an excellent candidate for 5G wireless networks. Furthermore, we statistically analyze the THz Band channel and determine the exponent of a path loss exponent model with log-normal shadowing. Channel coherence bandwidth is also determined using the power delay profile information. We also analyze different modulation schemes based on there energy consumption, data speed and bit-error rate. The analysis helps us determine the most energy-efficient modulation scheme for THz Band, which is very important as energy-consumption is a challenging problem for high speed wireless data transmission. Moreover, we propose a THz Band transceiver design that can be used for network demonstration and can be extended for protocol testing. Furthermore, we present the first ever experimental results of a THz Band line-of-sight (LOS) 2×2 MIMO channel using the principles of diffraction limited optics. Finally, we look at a some networking scenarios in full-dimension multiple-input multiple-output (FD-MIMO) based THz Band indoor wireless networks to determine the number of nodes that can be connected to a base station as a function of the antenna characteristics.

ÖZETÇE

Gelişmekte olan teknoloji Terahertz Band (0,3 - 10 THz) iletişiminin yüksek hızlı kablosuz iletişimi kapsamı öngörülüyor. Geniş bant genişliği, spektrum tıkanıklığı sorununu hafifletebildiği için 5G mobil ağlar için iyi bir aday yapar. Bu tezde, sadece tek girişli tek çıkışlı (SISO) tabanlı ağlar için değil, aynı zamanda çoklu girişli çoklu çıkış (MIMO) için THz Bandı'ndaki iletişimin ilk fiziki katmanı tasarımı öneriyoruz. Önce deney düzeneğini tanıtmak ve THz Bantında temel deneyleri gerçekleştirmek için kullanımını sunmaktayız. Farklı uzaklıklar, varış açıları ve reflektörler olarak etki eden nesneler için 260 GHz ila 400 GHz arasındaki yol kaybı ve faz gecikmesi ölçümleri anlatılmaktadır. THz Bant'ı 5G kablosuz ağlar için mükemmel bir aday olarak onaylayan saniyede terabit/saniye'lere ulaşan bağlantılar gösterdik. Ayrıca, istatistiksel olarak THz Bant kanalını analiz edip, log-normal gölgelendirme ile bir yol kaybı üstel modelinin üssünü belirledik. Kanal tutarlılık bant genişliği de güç gecikme profili bilgisi kullanılarak belirlenir. Ayrıca, enerji tüketimi, veri hızı ve bit-hata oranı temelinde farklı modülasyon şemalarını analiz ederiz. Analiz, enerji tüketimi yüksek hızlı kablosuz veri iletimi için zor bir sorundur. Dahası, şebeke tanıtımı için kullanılabilen ve protokol testi için genişletilebilen bir THz Bant alıcı-verici tasarımı öneriyoruz. Ayrıca, difraksiyon sınırlı optik prensiplerini kullanarak THz Bant görme hattı (LOS) 2×2 MIMO kanalının ilk deneysel sonuçlarını sunmaktayız. Son olarak, antenin bir fonksiyonu olarak bir baz istasyonuna bağlanabilecek düğüm sayısını belirlemek için tam boyutlu çoklu girişli çoklu çıkışlı (FD-MIMO) tabanlı THz Bandı kapalı kablosuz ağlarda bazı ağ senaryolarına bakıyoruz özellikleri.

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NOMENCLATURE

ADC	analog-to-digital converter
ASE	area spectral efficiency
AWGN	additive white Gaussian noise
balun	balanced to unbalanced
C/I	carrier to intermodulation ratio
CDF	cumulative distribution function
DAC	digital-to-analog
dB	decibels
dBc	decibels relative to the carrier
dBm	decibels relative to one milliwatt
DHBT	double heterojunction bipolar transistor
DSB	Double Sideband
FD-MIMO	full-dimension multiple-input multiple-output
FSPL	free space path loss
GHz	Gigahertz
HDMI	high-definition multimedia interface

HPBW	half-power beam width
Hz	hertz
IDFT	inverse discrete Fourier transform
IF	intermediate frequency
IF	intermediated frequency
IFBW	intermediate frequency bandwidth
IMD	intermodulation distortion
InP	Indium Phosphide
IoT	Internet-of-Things
KHz	kilohertz
LNA	low noise amplifier
LO	Local Oscillator
LOS	line-of-sight
LTE	Long-Term Evolution
MFSK	M-ary Frequency shift keying
mHEMT	metamorphic high electron mobility transistor
MHz	megahertz
MIMO	multiple-input multiple-output

MQAM	M-ary Quadrature amplitude modulation
NF	noise figure
NLOS	non-line-of-sight
PA	power amplifier
PAPR	peak-to-average power ratio
PB	petabyte
PLL	phase lock loop
QoS	quality of service
QPSK	Quadrature phase shift keying
RF	Radio frequency
RMS	Root mean square
Rx	Receiver
SHM	subharmonic mixer
SINR	Signal-to-interference-plus-noise-ratio
SISO	single-input single-output
SNR	signal to noise ratio
SSB	Single-Sideband
Tbps	terabits per second

TDM Time Division Multiplexing

THz Terahertz

Tx Transmitter

UHD ultra-high-definition

VNA Vector Network Analyzer

WPAN wireless personal area networks

YIG Yttrium iron garnet

Chapter 1

INTRODUCTION

Consumers of wireless communications are growing exponentially and demands for ubiquitous access to information and entertainment content at higher data rates are being intensified. According to a recent survey conducted by Ericsson, the mobile subscription has reached 7.5 billion and it is anticipated that by 2020, it will increase by 23%. As smartphones constitute the majority of subscribers utilizing wireless broadband services, further development of smartphone applications will increase data usage such that the worldwide data usage per smartphone is anticipated to reach 25000 petabytes (PB) until 2020 [Ericsson, 2016]. This huge increase will require very high data transfer rates. The current wireless systems support data throughput up to few hundred megabits per second, which is quite low in comparison to the wired communication standards such as Gigabit Ethernet and high-definition multimedia interface (HDMI) and anticipated future requirements. Furthermore, wireless data throughput is approaching the capacity of the current allocated spectrum and hence, to support the increasing demand in wireless communications, higher frequency bands must be exploited. Among the high frequency candidates, current research is more focused on 60 GHz ISM band which can provide data throughput up to few gigabits per second due to its limited bandwidth. To fulfill the anticipated requirements, wireless technology must shift to an even higher spectrum where the bandwidth can reach from few to several GHz. In this context, Terahertz (THz) Band is the most suitable candidate as it can provide contiguous bandwidth of several GHz or even up to THz and thus it must be included in communication spectrum of 5G wireless networks [Akyildiz et al., 2014, Khalid and Akan, 2016b, Yilmaz and Akan, 2015a, Huang and

Wang, 2011, Yilmaz and Akan, 2015b].

In this thesis, we propose physical layer design of THz band wireless communications by conducting detailed experiments and in-depth analysis. We first present a single-input single-output (SISO) THz Band channel measurement setup. Using this equipment we determine the channel characteristics of the wideband THz Band channel [Khalid and Akan, 2016b]. Moreover, we determine the path loss exponent and channel coherence bandwidth. Furthermore, we determine the most energy-efficient modulation scheme for terahertz band communication [Khalid et al., 2017b]. We also propose a SISO THz Band transceiver design [Khalid et al., 2017a]. In addition, we discuss the possible usage of a multiple-input multiple-output (MIMO) channel based principles of diffraction limited optics [Khalid and Akan, 2016a]. Finally, we propose a concept of guard width in a full-dimension multiple-input multiple-output (FD-MIMO) system, to provide optimum spacing between the adjacent beam widths in a THz Band network. The whole work confirms the success of THz Band wireless communications and affirms the idea of including it in the communication spectrum of 5G wireless networks.

1.1 Wireless Communication

One of the biggest success stories of applied engineering is wireless communications for at least last 30 years. The industry has made phenomenal progress and is expected to grow exponentially in the coming years. Impact of wireless communication is not only phenomenal on the stock market but also the society. Companies that were totally unknown before 30 years are now very popular throughout the world due to there ubiquitous wireless products, and in some countries the wireless communications industry even dominates the whole economy. Previously, wireless industry was considered merely as cellular telephony providing possibility of talking anytime anywhere. In recent times, due to the development of smart phones the industry now provides wireless access to information via the Internet. This has made data transmission a very important segment of the industry. This have led significant change in

working habits and mobility of workers - answering emails in a restaurant has become a usual practice. But besides these activities, a plethora of previously less obvious applications have been developed, and are vastly influencing our lives. These are Internet-of-Things (IoT) that includes smart sensors monitor factories, RFIDs, UHD video streaming, to name a few.

1.2 5G

The first generation of wireless communications was initiated a few decades back. Although a big success, it introduced only voice communication systems. Over the last years, the industry has evolved through second, third and fourth generation of wireless communications. Digital modulation, frequency reuse, Internet and advancement in physical layer technologies are the main constituents of this evolution. Moreover, with the increase of smart phones and wireless sensors, the fourth generation Long-Term Evolution (LTE) networks have become ubiquitous.

Demands for ubiquitous access to information and entertainment at higher data rates are increasing. Consumers are expecting wireless communication to reach the throughputs of wired communications. This demand suggests that wireless data rates reaching terabits per second (Tbps) speed must be realized in 5G wireless networks [Huang and Wang, 2011], [Daniels and Heath, 2007]. While current development in 60 GHz spectrum is expected to increase data rates up to 6 Gbits/s [IEEE, 2009], the available spectrum around this frequency is limited to 9 GHz and cannot satisfy the increasing demands of higher data rates [Ericsson, 2015]. To support such high throughputs of Tbps, enormous bandwidth is required, and hence, researchers around the world are exploring the Terahertz (THz) Band.

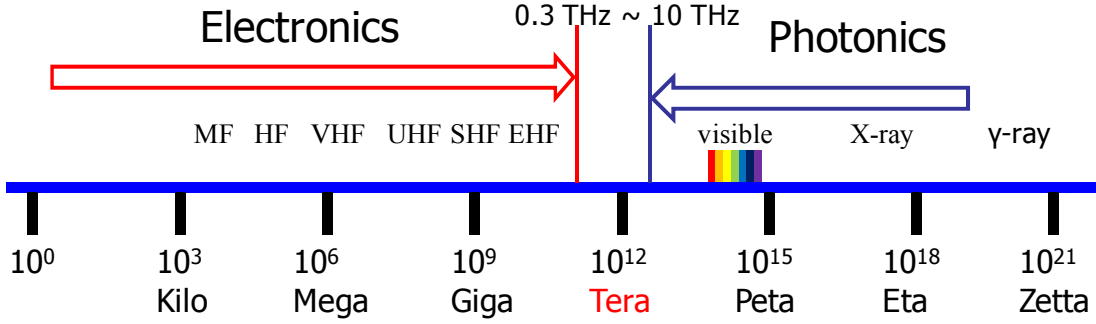


Figure 1.1: Electromagnetic spectrum showing THz Band gap between the photonics and electronics region.

1.3 Terahertz Band Overview

1.3.1 Terahertz Regime

Terahertz Band is a unique band in electromagnetic spectrum which is a transition region between the realm of microwave electronics and photonics as shown in Fig. 1.1. Due to awaiting standardization, slightly different frequency ranges of the spectrum are included in the band by different authors. Some authors from photonics society place 0.3 THz to 10 THz in this band [Williams, 2007]. While others from microwave society typically consider the range to be as 0.3 THz to 3 THz [Siegel, 2002]. Some also consider frequencies from 0.1 THz to 30 THz as part of Terahertz regime [Dhillon et al., 2017, Tonouchi, 2007]. In this thesis, we consider 0.3 THz to 10 THz as terahertz band mainly explore the lower frequency range around 0.3 THz which is synonymously termed as low-terahertz or submillimeter-wave [Lubecke et al., 1998].

1.3.2 Terahertz Gap

The unique band terahertz gap in electromagnetic spectrum is a transition region with interesting applications. The unique characteristics of this band makes it a very good candidate for high speed communication. However, the technology is not yet mature and requires a lot of research work. The radiation can be generated

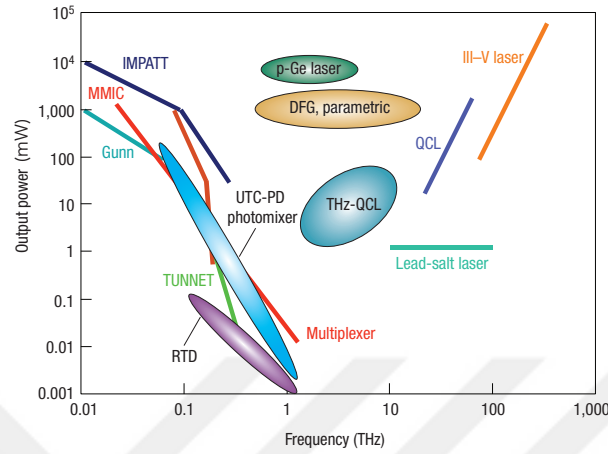


Figure 1.2: Reported THz Band sources [Tonouchi, 2007].

using two approaches: photonics or electronics. In the top-down approach, photonics technology is limited to milliwatt range around 2 THz frequencies while around 0.5 THz, the radiation is around to 0.1 milliwatts. Moreover, the wavelength of THz Band limits a high photon energy generation ($E = h\nu$) from simply a photonics source. On the other hand, in the bottom-up approach, the performance of electronics THz Band source and detector is limited by the active device performance due to the cut-off frequency of transistors. Due to these reasons, hardware is not as advanced as high frequency electronics or photonics and currently the current bottleneck of THz Band communications. This technological barrier is generally known as the 'THz gap' [Samoska, 2011, Huang et al., 2008]. Although an obsolete graph, Fig. 1.2 shows the THz Band sources and clearly shows the lack of sources creating that has created the gap in the electromagnetic spectrum.

1.4 Terahertz Band Applications

The unprecedented development of THz Band technologies has revealed its potential to support several 'real world' applications. For example, THz radiation is being used for material optimization of newly developed solar cells. Its imaging aspect can be a turning point for future airport security scanners. Other than its use in spectroscopy,

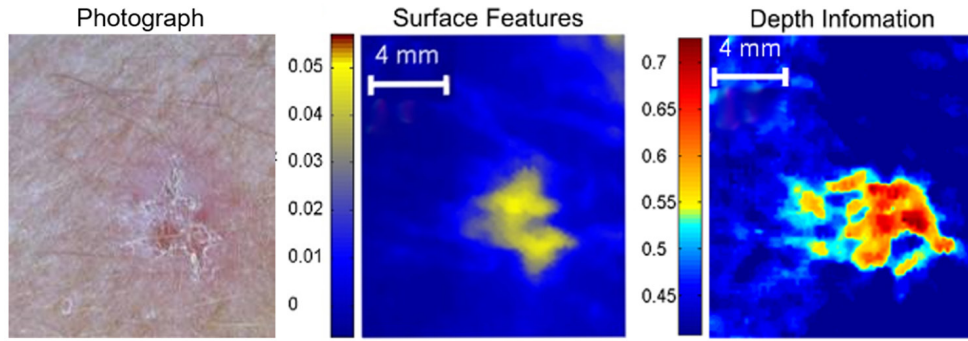


Figure 1.3: A vivo cancer image. Collected using a flatbed reflection, time domain, imaging system (TeraView Ltd, Cambridge, UK). The simple photograph (left) only show surface feature compared with the (center) terahertz image formed by plotting the reflected terahertz pulse. Depth image (right) reveals features invisible to naked eye [Dhillon et al., 2017].

semiconductor nano-structures, microscopy, biological and medical applications, non-destructive testing, space technologies etc, its use in communications is very beneficial as it can fulfill the soaring demands of high data rates [Dhillon et al., 2017]. Some examples are shown in Fig. 1.3 and 1.4. Therefore, it is important to explore the capabilities of THz technology for the next-generation wireless communications.

1.4.1 Terahertz Band Communication

The vast available spectrum in Terahertz band can fulfill the soaring demands of high speed communications. It can easily push the capacity limitations of current wireless networks. The enabling of high speed wireless communications will open a plethora of new applications and will allow many long-awaited wireless services such as wireless Ultra-high definition video links, terabit wireless local and personal area networks, high speed chip to chip communications. While the frequency regions above and below this band have been extensively explored, the THz Band is still the band in the electromagnetic spectrum that is least explored for communications.

Thz Band communication allows an unparalleled interconnections between the

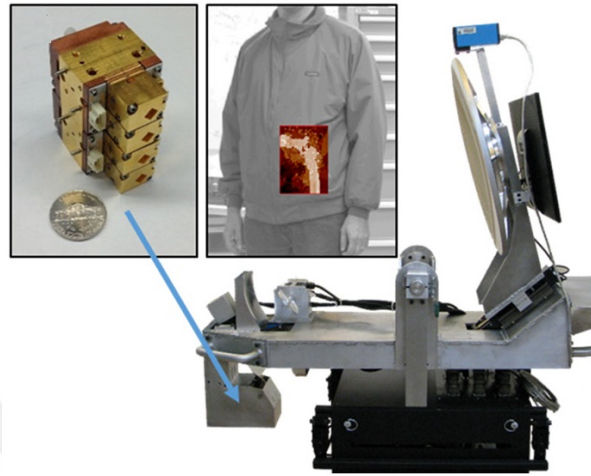


Figure 1.4: A snapshot taken from JPL's imaging radar at 8.3 Hz frame rate video revealing a handgun even inside the pocket of an insulated rain jacket. This was obtained using a four-element, 340 GHz heterodyne transceiver front end [Dhillon et al., 2017].

ultra high speed wired and wireless without compromising on speeds. This development will allow bandwidth intensive applications to run smoothly on not only static laptop user but also mobile tablet users. Few examples of such application are, high definition holographic video conferencing, ultra-high definition wireless TV link and ultra-high speed wireless LAN connection in laptops as shown in Fig. 1.5. While chip to chip communication is shown in Fig. 1.6.

1.5 Atmospheric attenuation in Terahertz Band

Compared to lower frequency range, the attenuation of the THz Band signal is quite high due to the oscillations of water molecules and atmospheric gases available in the air caused by the electromagnetic waves. Thus the amount of attenuation is strongly dependent upon the resonance and concentration of the gas molecules in air. Therefore, the THz Band communication link must operate in an atmospheric window with least amount of absorption. As shown in Fig. 1.7, these windows are

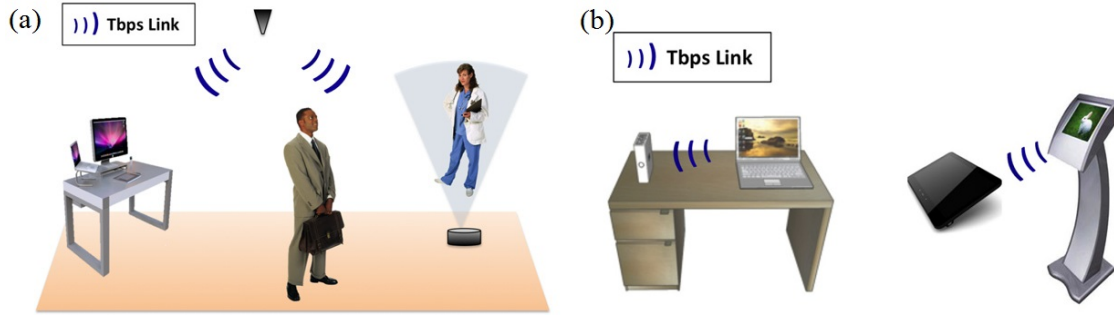


Figure 1.5: THz Band communication providing terabits per second wireless links (a) wireless local area network and (b) wireless personal area network [Akyildiz et al., 2014].

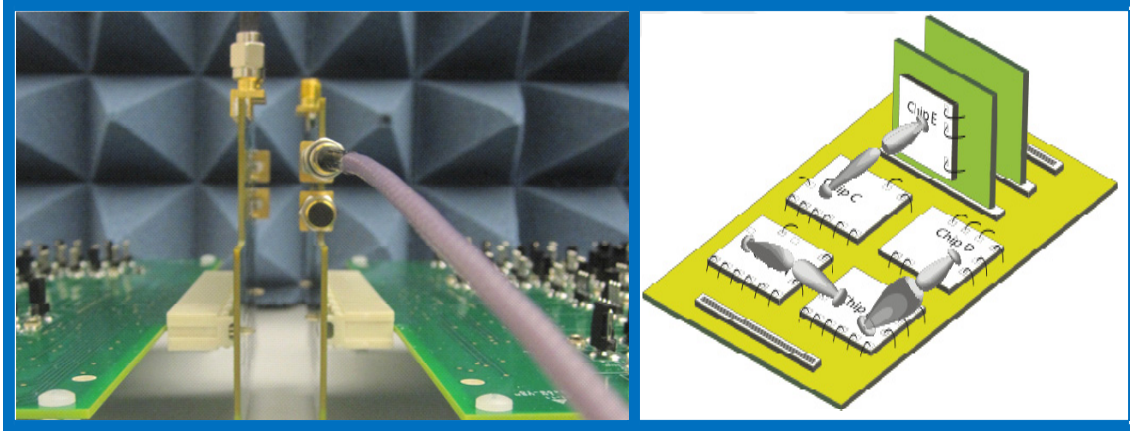


Figure 1.6: THz Band communication providing terabits per second wireless chip to chip link [dong Park, 2013].

normally centered around 94, 140, 250, 350, 410, 500, 650, 850, 1035, 1350, and 1500 GHz. Consider the attenuation peak at 557 GHz having attenuation larger than 104 dB/km [Schneider et al., 2012]. such atmospheric windows must be avoided for communications applications as much of the signal power is wasted in the air. Furthermore, the attenuation at 360 GHz is around 10 dB/km and must also be avoided [Schneider et al., 2012]. In this thesis, we carried experiments in the frequency range of 260 to 400 GHz.

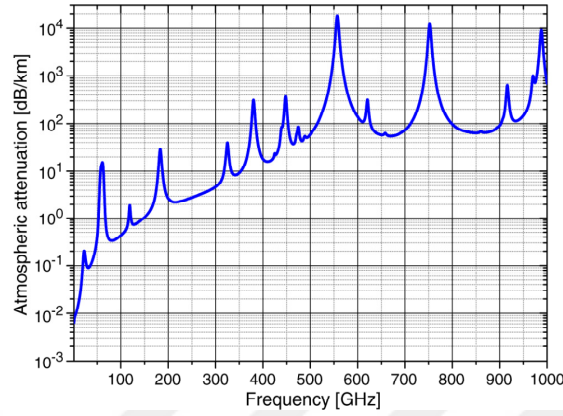


Figure 1.7: Specific attenuation of the earth's atmosphere at sea level in the frequency range between 1 and 1000 GHz for an air pressure of 1013 hPa, a temperature of 15 °C and a water vapor density of 7.5 g/m [Schneider et al., 2012].

1.6 Thesis Outline

In this thesis, we first present a SISO THz Band channel measurement setup in Chapter 2. The material is based on the work presented in [Khalid and Akan, 2016b]. Using this equipment we determine the channel characteristics of the wideband THz Band channel. In the next chapter, we use our described equipment to determine the path loss exponent and channel coherence bandwidth. Based on the work in [Khalid et al., 2017b], Chapter 4 proposes the most energy-efficient modulation scheme for communication in terahertz band. We also propose a SISO THz Band transceiver design in Chapter 5, which is based on the work in [Khalid et al., 2017a]. In the following part of this thesis, we discuss the possible usage of a MIMO channel based principles of diffraction limited optics in Chapter 6. This chapter is based on our work published in [Khalid and Akan, 2016a]. In Chapter 7, we propose a concept of guard width in an FD-MIMO system to provide optimum spacing between the adjacent beam widths in a THz Band network. Finally, the thesis is concluded in Chapter 8.

Chapter 2

CHANNEL MEASUREMENTS SETUP

The emerging technology Terahertz Band (0.3 - 10 THz) communication is envisioned to accommodate high speed wireless communication. Large bandwidth makes it a good candidate for 5G mobile networks. In this chapter, fundamental experiments on channel modeling at THz Band are presented with detailed analysis of the setup. The measurement setup consisted of subharmonic mixer and vector network analyzer. Path loss and phase delay measurements from 260 GHz to 400 GHz for different distances, angles of arrival and objects acting as reflectors were examined along with their capacity limits. We have shown that LOS link can reach speeds of terabits per second. In addition, reflections from materials were also examined and results indicated that, in case of signal obstruction, a reflector can be used for establishing NLOS link.

2.1 Introduction

Demands for ubiquitous access to information and entertainment at higher data rates are increasing. Consumers are expecting wireless communication to reach the throughputs of wired communications. This demand suggests that wireless data rates reaching terabits per second (Tbps) speed must be realized in 5G wireless networks [Huang and Wang, 2011, Daniels and Heath, 2007]. While current development in 60 GHz spectrum is expected to increase data rates up to 6 Gbits/s [IEEE, 2009], the available spectrum around this frequency is limited to 9 GHz and cannot satisfy the increasing demands of higher data rates [Ericsson, 2015]. To support such high throughputs of Tbps, enormous bandwidth is required, and hence, researchers around the world are exploring the Terahertz (THz) Band.

Increased research activities in THz Band have introduced equipment with improved performance. This equipment is categorized mainly into two groups depending on its approach for generating THz Band signal. In a top-down approach, photonics technology is used by translating back the optical frequencies, while in a bottom-up approach, the operating frequency of electronics components is increased [Piesiewicz et al., 2007]. The latter approach is preferred in communication systems due to compact sized components and reduced cost.

Channel characterization is the preliminary step in design of a communication system, case studies and experiments are being carried out to understand channel characteristics and material behavior in THz Band [Priebe et al., 2011, Kim and Zajić, 2015, Jornet and Akyildiz, 2011]. The setups commonly used are based on subharmonic Schottky diode mixers and are thus categorized into bottom-up approach [Priebe et al., 2010, Crowe et al., 2011, Jastrow et al., 2008a]. However, to develop accurate indoor channel models, there is a need to further case study the response of THz frequencies, in variant environments and to the materials that are still unexamined.

In order to get insight of channel behavior at THz frequencies, in this work a channel sounding setup was installed and detailed experiments were performed. The setup used was based on Vector network analyzer and subharmonic mixers. Currently, network analyzers available in the market support frequencies up to 70 GHz. Therefore, to characterize frequencies in THz Band, external subharmonic mixers were used.

For thorough channel modeling, experiments for both line-of-sight (LOS) and non-line-of-sight (NLOS) were conducted. LOS experiments were carried by varying distance between transmitter and receiver and recording path loss at frequencies ranging from 260 GHz to 400 GHz. Whereas NLOS experiments included response measurement for different angles of arrival, shadowing effect and reflection from various materials. An indoor wireless channel frequently encounters obstructions, therefore, detailed experiments for NLOS link are required. To collect further data, experiments were carried, in which objects like human hand, fiber mat and glass were modeled as

obstructions. In addition, effects of common objects like acoustic ceiling panel and wooden tables were also measured. These experiments help in determining absorption coefficient of the materials and their potential to be used as reflector in NLOS link.

Results suggest that huge bandwidth available in THz Band will introduce variety of applications, which require very high data rates. Some of these applications include wireless interconnection of desktop gadgets, THz femtocell mobile communication, secure Terabit wireless communication and high speed wireless back-haul connection. It can also be inferred that the higher path loss and antenna directivity can enhance frequency reuse and multiply capacity of the system 1000 times when deployed in femtocell regime. Therefore, channel characteristics are investigated for different distances, antenna orientations and materials acting as obstruction or reflector. Although huge bandwidth and directional nature of THz Band opens door for many novel applications, some apparent shortcomings suggest that it cannot replace gigahertz communication. The major drawbacks are smaller coverage area, power inefficient circuits and expensive deployment.

The remainder of this chapter is organized as follows. In Section 2.2, we describe our measurement system used for experimentation along with its characterization. In Section 2.3, we present channel measurements at THz Band for LOS and NLOS configurations. Various channel responses are shown for different distances, antenna misalignment and shadowing to account for maximum possible cases. Finally, we conclude the chapter in Section 2.4.

2.2 THz Band measurement system

2.2.1 Channel Sounding Testbed Specifications

Channel modeling is a preliminary step in design of a communication system and requires careful experimentation. Special equipment is required to carry out these experiments. For millimeter wave, fully developed and integrated equipment is easily available, and hence, accurate models are already constructed by following predefined step by step approach. However, for THz Band channel modeling, the experimen-

tation is rather difficult and much more specialized equipment is required. This is because single solution supporting up to THz Band is yet not available and separate modules must be added to the contemporary equipment. These modules must be dealt with extreme care to obtain accurate results. In this section, we discuss the equipment used for the experimentation and describe their suitability for THz Band channel modeling.

The measurement setup consists of two major parts Anritsu Vector Network Analyzer (VNA) MS4647B and VDi WR2.8MixAMC module. VNA MS4647B is a wide-band equipment used to measure scattering parameters. However, the upper limit of this device is 70 GHz. To measure channel characteristics at THz frequencies, extension modules WR2.8MixAMC are attached to the VNA. These modules contain subharmonic mixers that upconverts the test signal from VNA before transmitting. This signal after passing through the channel is downconverted at the receiver module and fed back to the VNA. Difference in the transmitted and received signal is analyzed to find channel characteristics.

VDi WR2.8MixAMC is an ultra broadband module that operates from 260 GHz to 400 GHz. Typical Single-Sideband (SSB) mixing loss of this device is 15 dB. Test port waveguide has a cutoff frequency of 211 GHz. These modules are attached externally to the VNA to perform the up and down conversion function [Crowe et al., 2011]. Maximum bandwidth supported is 20 GHz, however, single sweep bandwidth is limited to 19 GHz by the upconversion feature, as it operates from 1 GHz to 20 GHz. Hence, all single sweep measurements recorded are 19 GHz wide.

Due to stringent stability requirements, Yttrium iron garnet (YIG) based tunable synthesizer is used, which generates local oscillator signal f_{LO} . This signal is doubled, and then, multiplied by a factor of 12 to generate THz Band signal, which is fed to f_{LOTHz} port of subharmonic mixer. Intermediate frequency f_{IF} port of this mixer is attached to the VNA that sweeps the output frequency up to 20 GHz ($f_{LOTHz} + f_{IF}$). To generate frequencies higher than the single sweep limit, f_{LO} is varied from 10.79 GHz to 15.875 GHz, which allows covering the whole band of 260 GHz to 400 GHz.

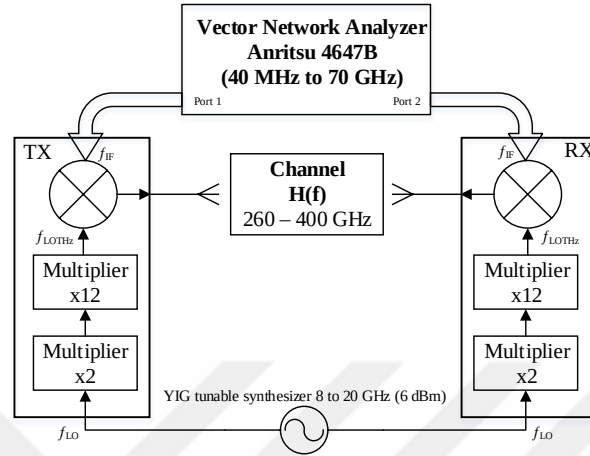


Figure 2.1: Block diagram of THz Band measurement setup.

Phase coherence in our setup is achieved by using a common local oscillator for both the transmitter and the receiver module. The corresponding block diagram of our setup is shown in Fig. 2.1. The subharmonic mixer used produces Double Sideband (DSB) and thus using common f_{LO} causes two frequencies to accumulate after homodyne down conversion, this behavior can be avoided if a small offset is maintained between upconverted and downconverted signal so that the unwanted frequency may be easily filtered out [Hesler et al., 2010].

2.2.2 Channel Sounding Testbed Configuration

Channel transfer function is acquired by recording scattering parameters (s-parameters) using the THz Band measurement setup. One of the most important part in making good s-parameters measurement is VNA calibration. While VNA is designed with quite precaution to attain high-linearity and spectral purity in its sources, some imperfections like frequency mismatch, imperfect directivity of the internal coupler and added test cables and modules makes it necessary to perform calibration. Calibrating the VNA corrects these imperfections along with other defects in measurements. Several types of calibrations can be performed depending on the application of interest [Dunsmore, 2012, Priebe et al., 2011, Alshabo et al., 2014]. For channel modeling,

Table 2.1: Setup parameters

Parameter	Symbol	value
<i>Measurement points</i>	N	1000
<i>Test signal power</i>	P_{in}	0 dBm
<i>Start frequency</i>	f_{start}	1 GHz
<i>Stop frequency</i>	f_{stop}	20 GHz
<i>Bandwidth</i>	B.W.	19 GHz
<i>IFBW</i>	Δf_{IF}	1 KHz

through/reciprocal calibration is performed with direct interconnection of the module's waveguide. All the later measurements are taken with diagonal horn antenna attached at both the transmitter and the receiver. Gain of the antennas at center frequency is 25 dBi each. Whereas at the lowest frequency (260 GHz) it is 2 dB lesser and at the highest frequency (400 GHz) it is 1.5 dB higher.

Full 19 GHz band measurements are recorded with 10 points averaging and 1 KHz IFBW. These parameters significantly reduces noise floor and improves dynamic range [Dunsmore, 2012, Agilent, 2000]. Test signal with power of 0 dBm is used that provides dynamic range of 60 dB. Details of the measurement parameters are given in the Table 2.1. Trace smoothing of 5% is used for more robust interpretation of the results.

2.3 THz Band channel measurements

2.3.1 Line-of-sight Channel Measurements

Simplest channel model is the LOS channel, in which the transmitted signal is directly received without any obstruction. The measurement setup is shown in Fig. 2.2. In LOS channel measurements, we measured transmission coefficient at different distances from 10 cm to 95 cm, that are often expected in a typical office desktop table. All the measurements were performed on a wooden base and antenna height in



Figure 2.2: THz Band channel measurement setup.

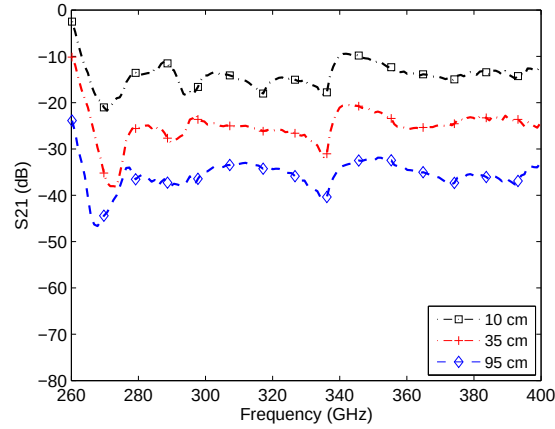


Figure 2.3: Full band channel response.

the system was kept as 3 cm. Path loss measurements were recorded in segments of 19 GHz and compiled to produce the frequency response from 260 GHz to 400 GHz in Fig. 2.3. Graph depicts how the path loss increases as the distance is increased and gives a first insight into the basic capabilities and practical success of THz Band communication system. These measurements include antennas with gain of 25 dBi, and hence, considered as a part of system. Therefore, while considering channel transfer function the gain of antennas may be subtracted.

For the known frequency, distance and antenna gain, one can easily find the path loss between transmitter and receiver using friss transmission equation [Shaw, 2013]. For distances, $d = 10\text{cm}$, $d = 35\text{cm}$ and $d = 95\text{cm}$ the theoretical calculated free

space loss at 290 GHz is 15.7 dB, 26.57 dB and 35.24 dB, respectively. Whereas the experimental results recorded are 17.2 dB, 27.4 dB and 36.1 dB, respectively. This variation can be reduced by thorough calibration, using anechoic environment and perfect module alignment.

To examine results more carefully, single sweep channel response from 285 GHz to 304 GHz is shown in Fig. 2.4 and Fig. 2.5 for 35 cm distance. It can be seen from the results that phase shift is mostly linear throughout the band but shows non-linear characteristics at some frequencies with unexpected magnitude response. This is due to unwanted resonances that can be avoided in an anechoic environment.

To further elaborate our results, Shannon's capacity theorem [Shannon, 1948] was applied and capacity (C) of our measured channel was calculated using

$$C = W \cdot \log_2(1 + P/N) \quad (2.1)$$

where W is the bandwidth, P is signal power and N is noise power. For the distances, $d = 10\text{cm}$, $d = 35\text{cm}$ and $d = 95\text{cm}$ average path loss from 260 GHz to 400 GHz was measured as 14.23 dB, 24.14 dB and 35.75 dB, respectively and the corresponding channel capacity was calculated as 9.26 b/s/Hz, 7.62 b/s/Hz and 5.713 b/s/Hz. The noise floor for the proposed setup was approximately -70 dB. It can be seen that, if distance is increased 3.5 times, the capacity of the channel falls by 18%. However, for short distances, Tbps throughput is achievable by the large bandwidth. Consider for the case of 10 cm, using bandwidth of 140 GHz, one can reach speeds of 1.29 Tbps, which is more than 200 times faster than the speeds of currently available technologies.

2.3.2 Non-line-of-sight Channel Measurements

Future wireless communication is expected to support high speed data communication reaching throughput of Tbps. Applications will include mobile communications, wireless local area networks or interconnection of gadgets on desktop like HDMI and flash memory. These applications require channel measurements with regard not only

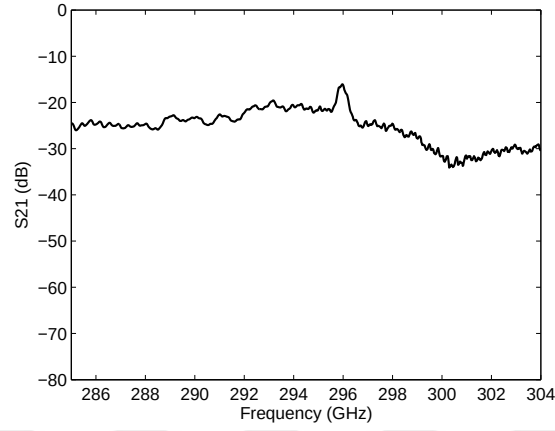


Figure 2.4: Single sweep magnitude response of THz channel.

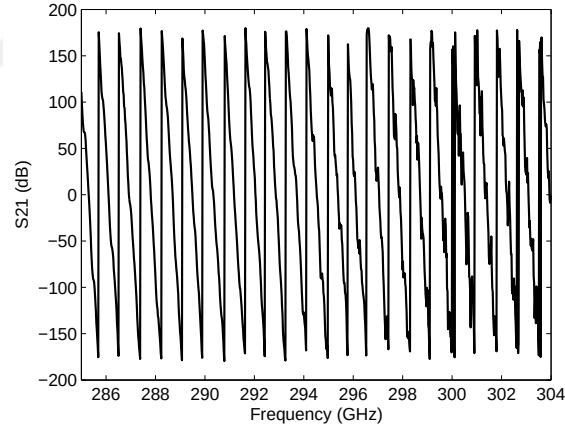


Figure 2.5: Single sweep phase response of THz channel.

LOS communication but also NLOS communication as users cannot be expected to place their devices perfectly aligned. Hence, for more comprehensive results, antenna misalignment effects were measured by varying the angle of arrival from 0 to 30 degrees in azimuth direction. All measurements were recorded with 35 cm distance and shown in Fig. 2.6. Due to highly directive antennas used, signal power drops rapidly. This behavior suggests that robust beamforming algorithms will be required to make Tbps communication viable. These algorithms must easily adapt alterations

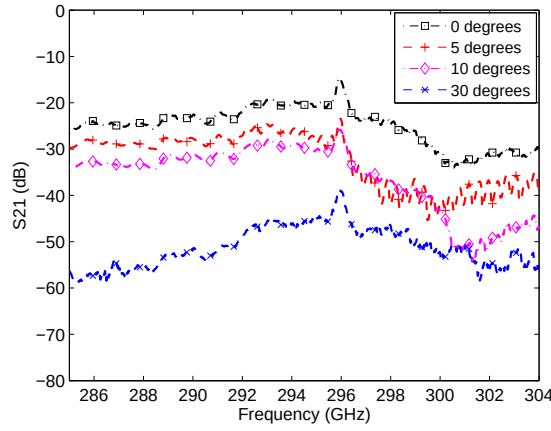


Figure 2.6: Transfer function for antenna misalignment.

and mitigate the affect of misalignment.

Indoor wireless radio channel is rather more challenging as it is frequently susceptible to shadowing. Mainly shadowing is caused by any obstacle appearing between transmitter and receiver. This obstacle attenuates the signal power depending on the material type. A strong attenuation can even block the signal and stop communication [Goldsmith, 2005]. To model channel characteristics in the presence of obstruction, experiments were performed for fixed distance of 35 cm between transmitter and receiver. The attenuation due to shadowing by human hand, empty glass of water and a 80 grams fiber mat with 3 mm thickness is shown in Fig. 2.7. No obstruction shows transmission loss without any obstruction in transmission path.

Absorption coefficient of different materials was determined using the transmission loss. The attenuated signal power was subtracted from the reference signal to get added transmission loss due to the object. Finally, this extra loss was used to calculate the absorption coefficient of the material. For instance, a clear glass with thickness of $t_{glass} = 2 \text{ mm}$ or a 80 gram fiber mat with thickness of $t_{fiber} = 3 \text{ mm}$, attenuation of approximately 6.14 dB and 27.55 dB, respectively, were measured at 290 GHz. In the case of a human hand placed as obstruction, 45.6 dB of attenuation was observed. The attenuation coefficient of different materials is summarized in the Table 2.2.

Table 2.2: Absorption coefficient of different objects

Object	α_{290GHz}	α_{350GHz}
<i>Fiber mat</i>	0.168 mm^{-1}	0.171 mm^{-1}
<i>Glass</i>	0.049 mm^{-1}	0.048 mm^{-1}
<i>Empty water glass</i>	0.958	0.962
<i>Hand</i>	0.995	0.982

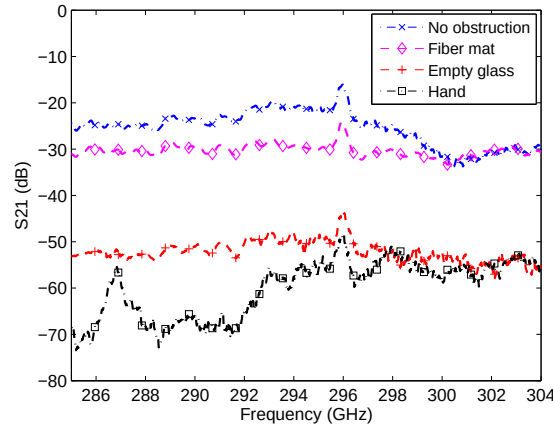


Figure 2.7: Transfer function for shadowing effect.

The results show that, if a thin glass or a fiber sheet appears as obstruction in the communication link, the signal to noise ratio (SNR) of channel decreases, and hence, the data rate also decreases. However, if obstruction is a human hand, the intense signal attenuation might lead to disconnection of communication link. To reestablish communication, the system should turn to a NLOS path that utilizes reflection properties of the materials in the surroundings.

To measure reflections from different materials available in a typical indoor channel, system configuration was kept so that the transmitted signal was incident at 45 degree on the object under observation. Distance between antenna and the reflecting object was kept 5 cm, and thus, the total distance our signal travels is 10 cm. Setup for measuring material reflection is shown in Fig. 2.8. Path loss for signal reflected

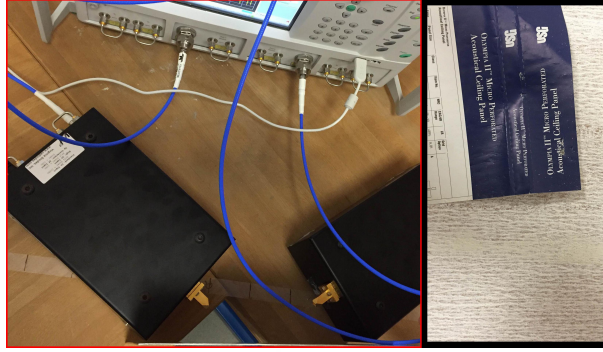


Figure 2.8: Reflections measurement setup.

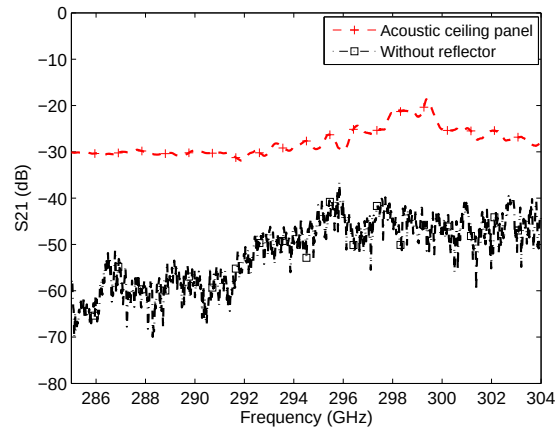


Figure 2.9: NLOS magnitude response with acoustic ceiling panel as reflector.

from an acoustic ceiling panel of 4 mm thickness and wooden ply of 3 mm thickness are shown in Fig. 2.9 and 2.11, respectively. Ceiling panels are often used in indoors and its transfer function depicts that the panel can be used to establish NLOS link in case when LOS link is blocked. However, signal reflected from wood was very weak and thus wood can't be considered a good candidate for NLOS links. Phase measurement for both configurations is shown in Fig. 2.10 and 2.12 to confirm that it delays linearly in case of reflected signal as well.

Channel capacity for LOS and NLOS channels are compared in Table 2.3. Huge decrease in capacity can be seen in NLOS links. For instance, if angle of arrival is 10

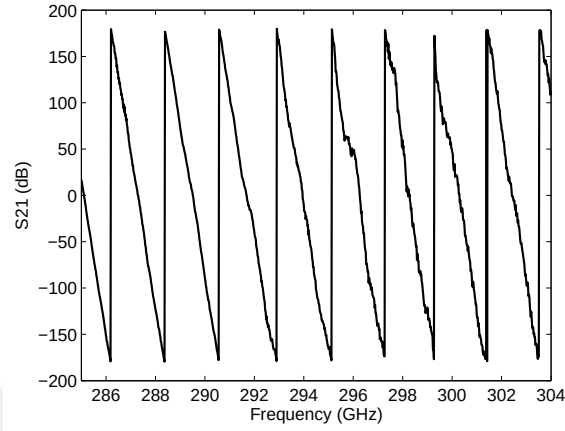


Figure 2.10: NLOS phase response with acoustic ceiling panel as reflector.

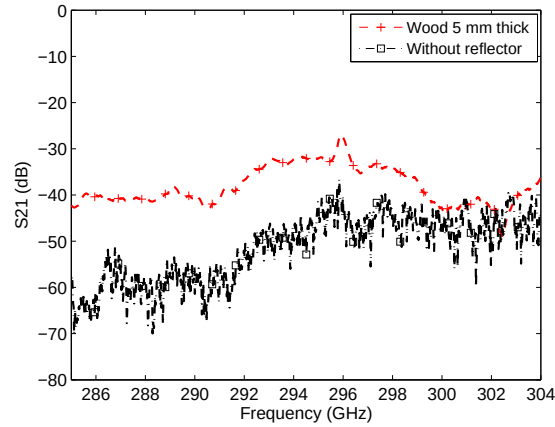


Figure 2.11: NLOS magnitude response with wood as reflector.

degrees, the capacity of the system falls by 30%. However, in case of reflection analysis, acoustic ceiling panel shows very promising results. It can be used in establishing NLOS link with decent throughput. These panels are widely installed in offices and hence, if used, can be a very cost effective solution.

Further case study on this topic can include multiple-input multiple-output (MIMO) antenna technique. MIMO is a very popular technique that can enhance the capacity

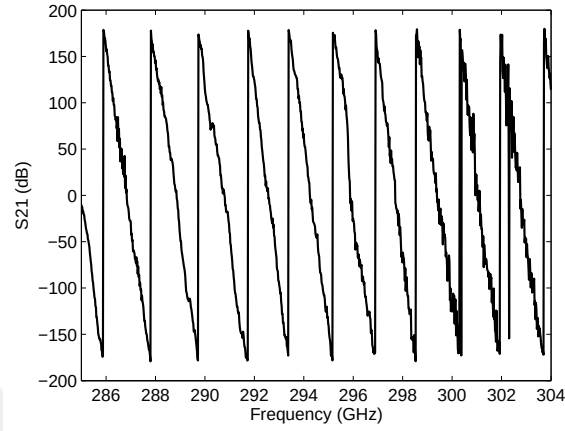


Figure 2.12: NLOS phase response with wood as reflector.

Table 2.3: Capacity comparison for measured scenarios

Link	Distance	Obstruction	Angle of arrival (degrees)	Capacity (b/s/Hz)
<i>LOS</i>	10 cm	-	-	9.26
	35 cm	-	-	7.26
	95 cm	-	-	5.713
<i>NLOS</i>	35 cm	-	5	7.03
	35 cm	-	10	6.44
	35 cm	-	30	3.23
	35 cm	Fiber mat	-	6.732
	10 cm	Ceiling panel	90	6.719

of a wireless link by introducing multiplexing gain. The basic idea is to use multiple transmit and receive antennas to exploit the multipath behavior of the channel. Although the results show that the THz band channel has very weak multipath components, but as long as reflectors like metal or even ceiling panel exists in the path, MIMO systems in THz band can be implemented. Our setup can be easily modified

to implement virtual antenna arrays to observe MIMO characteristics of the channel as shown in [Ranvier et al., 2009].

2.4 Conclusion

We have presented Terahertz Band channel characteristics using Vector Network Analyzer and subharmonic mixers based measurement system. Channel impulse response from 260 GHz to 400 GHz are recorded and presented. System capabilities and parameters are investigated in detail. We have measured simple transmission coefficient for LOS link to analyze basic capabilities and practical success of THz Band communication system. Measured results show that Tbps throughput is achievable in THz Band. It has become evident that THz Band will enable a plethora of applications such as high speed mobile communications, wireless local area networks or interconnection of gadgets on desktop like HDMI and flash memory. Results for LOS measurements confirm that high gain antennas are necessary to mitigate the path losses at THz frequencies. In addition, the experimental setup demonstrated in this work is a significant example that can boost studies on 5G indoor wireless communication.

Chapter 3

STATISTICAL CHARACTERIZATION AND ANALYSIS

This chapter presents measurements and statistical characterization to compare three potential bands of low-THz channel for Internet of things (IoT) applications, namely, the 300 to 319 GHz, 340 to 359 GHz and 380 to 399 GHz bands. From the large set of measurements performed in line-of-sight (LOS) and non-LOS (NLOS) environments, parameters for path loss model with shadowing are evaluated. Our results show that the path loss exponent for 310 GHz band, 350 GHz band and 390 GHz band is 2.03, 1.90 and 1.95, respectively. The impacts of different materials acting as surfaces in LOS desktop channels and reflectors in NLOS environments are also examined. Finally, the statistical analysis due to temporal, spatial and multipath propagation is performed to determine the best fit distributions.

3.1 Introduction

In this work, we first installed a Vector Network Analyzer (VNA) and subharmonic mixer based THz Band equipment and performed detailed experiments for channel sounding. The current generation of network analyzers available in the market support frequencies up to 70 GHz. Therefore, to characterize frequencies in THz Band, external subharmonic mixers are used. Limited literature is available that provides in-depth analysis of THz Band indoor channels. Most of the previous work uses narrow band equipment and explore only 300 to 320 GHz Band [Priebe et al., 2011, Kim and Zajić, 2015, Jastrow et al., 2008b, Salhi et al., 2013, Priebe et al., 2013, Priebe and Kurner, 2013, Priebe et al., 2014, Kürner et al., 2015]. Secondly, the materials used were very limited in them.

For thorough statistical channel characterization, three bands of 19 GHz were

selected from low-THz Band and a large set of experiments for line-of-sight (LOS) and non-line-of-sight (NLOS) cases were conducted. The tested bands included the ranges of 300-319 GHz, 340-359 GHz and 380-399 GHz. LOS experiments were carried by varying distances between the transmitter and receiver to record path loss in all bands. Several obstacles that are commonly encountered in an indoor channels such as transparent plastics, keys and mobile phones were used in the LOS experiments to measure the effects of shadowing. From the large set of data, path-loss exponents and variations due to shadowing were measured for all the tested bands. NLOS experiments included response measurements of reflections from various surfaces also present in indoor environments, such as ceiling panels and various metals. Additionally, experiments were performed to test various desktop table materials such as wood, metal and rubber to find the effect of the surfaces on which the communication takes place. Based on the analysis of multipath propagation, we show unprecedented power delay profile, RMS delay spread, mean excess delay and maximum excess delay with picoseconds accuracy and coherence bandwidth for LOS and NLOS environments for the all the bands under consideration.

The goal of this research is threefold;

- extensive measurement campaign to understand channel characteristics with different materials and environments expected in a THz Band wireless channel used for IoT;
- full analysis of three potential bands of 300 to 400 GHz channel with in-depth comparison integrated in single article;
- high resolution time domain measurements with picosecond accuracy.

The remainder of this chapter is organized as follows. In Section 3.2, we describe our measurement system used for experimentation along with its characterization. Statistical characterization of the low-THz Band is provided in Section 3.3. Section 3.4 provides the small-scale multipath propagation analysis. Finally, we conclude the chapter in Section 3.5.

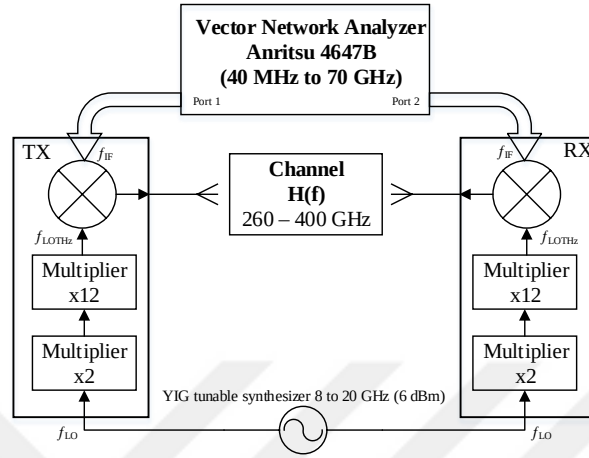


Figure 3.1: Block diagram of THz Band measurement setup.

3.2 The Measurement Equipment and Site

The THz Band indoor channel measurements were conducted with a custom configured wideband frequency domain channel sounder. The equipment was installed in a typical office environment to measure indoor channel propagation characteristics. Experimentation in THz Band is rather difficult due to extremely small wavelength (1 - 0.1 mm) and the fact that single solution testbeds are not yet available, therefore, modifications are required to the contemporary equipment. In this section, we describe our testbed used for the experimentation and provide details of the experiment site.

3.2.1 Testbed Specifications

The measurement setup can be categorized into two main parts, Anritsu VNA MS4647B and VDi WR2.8MixAMC extension modules. The VNA MS4647B measures channel impulse response in frequency domain. However, the upper limit of this device is 70 GHz and to measure the channel impulse response at 300 GHz and above, WR2.8MixAMC extension modules are attached to the VNA. These modules contain subharmonic mixers that upconvert the VNA test signal and transmit it to the

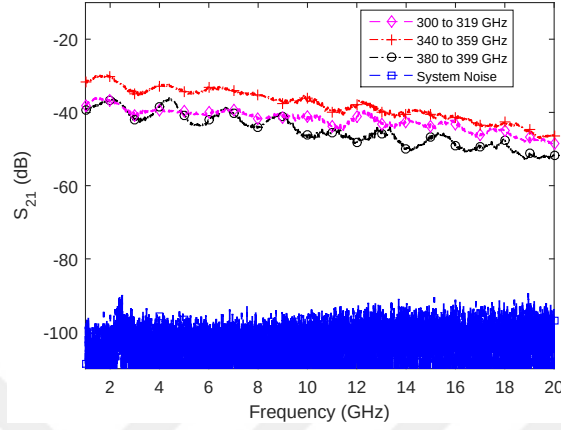


Figure 3.2: Measurement equipment losses.

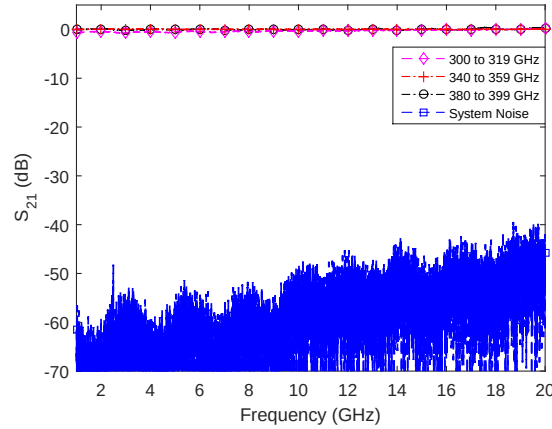


Figure 3.3: Measurement equipment losses after calibration.

channel. This signal is received at the receiver and fed back to the VNA after down-conversion. The VNA monitors changes in the amplitude and phase of the transmitted and received signals to show the scattering parameters (s-parameters).

The extension module, VDi WR2.8MixAMC operates from 260 GHz to 400 GHz with a typical Single-Sideband (SSB) mixing loss of 15 dB. The test port waveguide has a cutoff frequency of 211 GHz. A pair of these modules is attached externally to the VNA to measure THz Band channel [Crowe et al., 2011]. The maximum band-

width supported by the module is 20 GHz, however, the coupler used for upconversion limits it to 19 GHz as its operating range is from 1 GHz to 20 GHz. Hence, all our measurements are for 19 GHz wide bands. The block diagram of our setup is shown in Fig. 3.1. The local oscillator signal f_{LO} , required by extension module, is provided using a Yttrium Iron Garnet (YIG) based tunable synthesizer. The frequency of f_{LO} is selected based on the frequency band required at the antenna. For instance, to generate {300, 340, 380} GHz signal, we set $f_{LO} = \{12.45833, 14.12500, 15.79166\}$ GHz, respectively which is then multiplied by a factor of 24 to be fed to the f_{LOTHz} port of subharmonic mixer. Intermediate frequency f_{IF} port of the mixer is swept by the VNA from 1 to 20 GHz to generate $(f_{LOTHz} + f_{IF})$ at the output. Phase coherence in our setup is achieved by using a common local oscillator that provides synchronized frequency signals to all the components. However, the subharmonic mixer used produces Double Sideband (DSB) and thus using a common f_{LO} at both the transmitter and the receiver causes two frequencies to accumulate after homodyne down conversion. To avoid this behavior, a small offset should be added between the f_{LO} of the transmitter and the receiver so that the unwanted frequency may be easily filtered out [Hesler et al., 2010].

3.2.2 Testbed Configuration

The full 19 GHz channel impulse response at {300, 340, 380} GHz was measured by recording s-parameters S_{21} for the $f_{IF} = 1$ to 20 GHz. The transmitter f_{IF} input power was set as -10 dBm (0.1 mW). The relatively low input power was necessary in order to avoid damaging the transmitting subharmonic mixer, as well as the receiving subharmonic mixer at short link distances. Additionally, the low output power of -10 dBm EIRP (considering 10 dB coupler loss, 15 dB mixing loss and 25 dB antenna gain) emulated office desktop channel where it is required to keep a transmitted signal within the boundaries of a table. The transfer function of the transceiver is shown in Fig. 3.2, which is obtained by directly interconnecting the transmitter and the receiver. It is evident from the transfer function that these losses

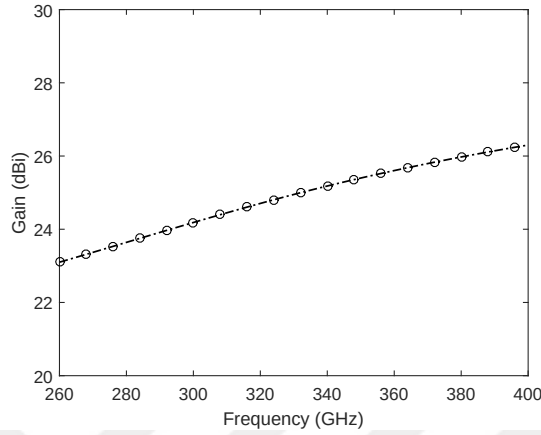


Figure 3.4: Antenna gain 260 GHz to 400 GHz range.

should be excluded while measuring the channel impulse response. Calibrating the VNA can exclude these losses along with other defects in measurements. Several types of calibrations can be performed depending on the application of interest [Dunsmore, 2012, Priebe et al., 2011, Alshabo et al., 2014, Lewandowski and Wiatr, 2010]. In our case, through/reciprocal calibration is performed to correct the amplitude of the measured channel impulse response. The transceiver response after calibration is shown in Fig. 3.3.

To compensate the significant path loss at THz frequency, high gain diagonal horn antennas are used at the transmitter and the receiver. Gain of the antennas at center frequency is 25 dBi each. Whereas at the lowest frequency (260 GHz) it is 2 dB lesser and at the highest frequency (400 GHz) it is 1.5 dB higher. The half power beamwidth is 12° . The gain of our antenna, from 260 to 400 GHz, is shown in Fig. 3.4. For all the three bands, transmitter and receiver antennas are vertically polarized. In the later sections of this chapter, since all measurements are performed while an antenna is attached at both ends, the antennas will be considered a part of the transfer function.

S-parameters for the 19 GHz bands were recorded with 2 MHz of frequency step, 10 points averaging and 1 KHz IFBW. These parameters significantly reduce noise floor and improve dynamic range [Dunsmore, 2012, Agilent, 2000] to provide spatial

Table 3.1: Setup parameters.

Parameter	Symbol	Value
<i>Measurement points</i>	N	9501
<i>Test signal power</i>	P_{in}	-10 dBm
<i>Start frequency</i>	f_{start}	1 GHz
<i>Stop frequency</i>	f_{stop}	20 GHz
<i>Bandwidth</i>	B.W.	19 GHz
<i>IFBW</i>	Δf_{IF}	1 KHz
<i>Average noise floor</i>	P_N	-103.4 dB
<i>Dynamic range</i>	R_{DYN}	60 dB
<i>Time domain resolution</i>	Δt	0.5 ps
<i>maximum excess delay</i>	τ_m	4 ns

and temporal resolutions of 0.1578 mm and 0.526 ps, respectively. The average noise floor of the VNA is -103.4374 dB (Fig. 3.2), however, for our calibrated system, the average noise floor is -59.4123 dB (Fig. 3.3), which is also the dynamic range of our testbed. The number of sweep points is set to 9501 and the maximum excess delay is 4 ns. A summary of measurement parameters is given in Table 3.1.

3.2.3 Site Description

Our experiments were performed in NWCL lab, which is a typical indoor office environment. On a typical office desktop, the LOS communication is operated at distances between 5 and 70 cm. Our experimental setup as shown in Fig. 3.5 was used for LOS measurements and the scheme for measurements at different transmitter-receiver separations is shown in Fig. 3.6. Several measurements were taken at each transmitter-receiver separation on different time instances to capture temporal variations for each of the frequency bands and then the transmitter-receiver pair was moved further apart to a new position. An average measurement, involving the mean of 10

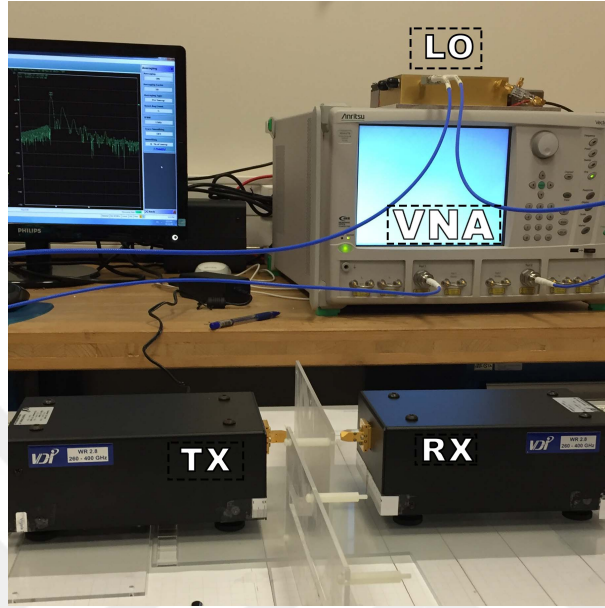


Figure 3.5: Measurement setup for LOS experiments with linear displacement.

previous temporal reading was also generated. A similar mechanism was adopted to characterize various desktop table materials to check the effects of table materials on THz Band communication.

To create shadowing in the channels, different objects were placed between the transmitter and receiver, and again several temporal instantaneous and average readings were recorded. The setup of shadowing by means of plexiglass is shown in Fig. 3.5. The LOS measurement setup and scheme for spatial displacement is shown in Fig. 3.7 and Fig. 3.8, respectively. Keeping the distance between the transmitter and receiver fixed at 10 cm, the setup was spatially displaced, each time with a difference of 30° from the previous position. Thus, the spatial data from 12 different locations was collected for the spatial LOS measurements.

Finally, for the NLOS reflections, the transmitter and receiver were placed perpendicular to each other and a reflection surface at 45° was placed between the two as shown in Fig. 3.9 and Fig. 3.10. Different reflective surfaces that are common in an indoor office environment such as ceiling panels, paint-coated aluminum surfaces

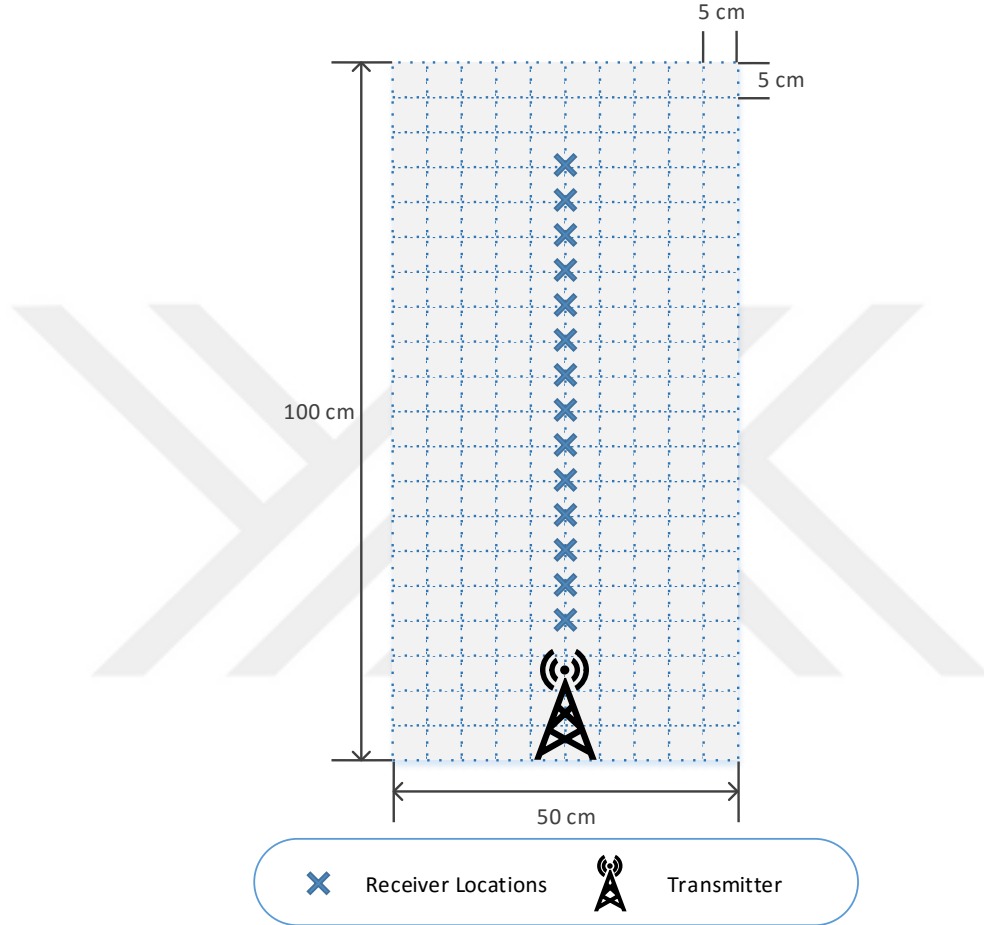


Figure 3.6: Illustration for LOS experiments with linear displacement.

of typical office cabinets and copper were used. Fig. 3.9 shows a measurement in progress using a ceiling panel as a reflection surface.

3.3 Statistical Characteristics of Low-THz Band Channel

Channel modeling is one of the most difficult parts of wireless communications as its prediction requires detailed knowledge of the environment. Usually, site description is not easily available and thus statistical models can be very helpful in predicting the channel behavior. It should be noted that the statistical characteristics of the

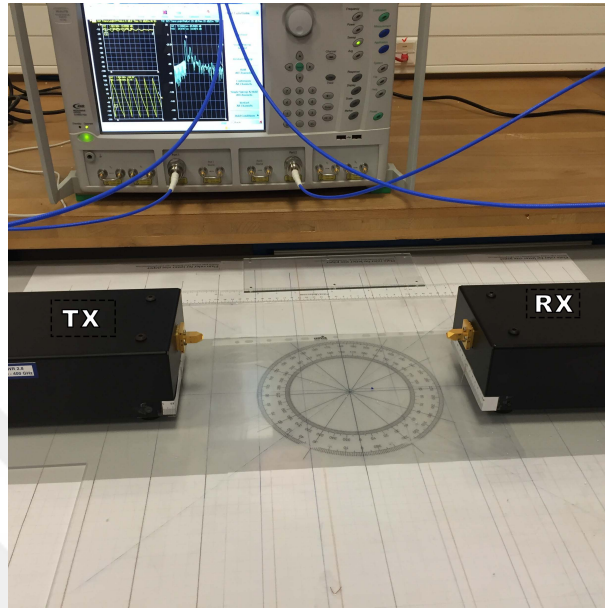


Figure 3.7: Measurement setup for spatial displacement.

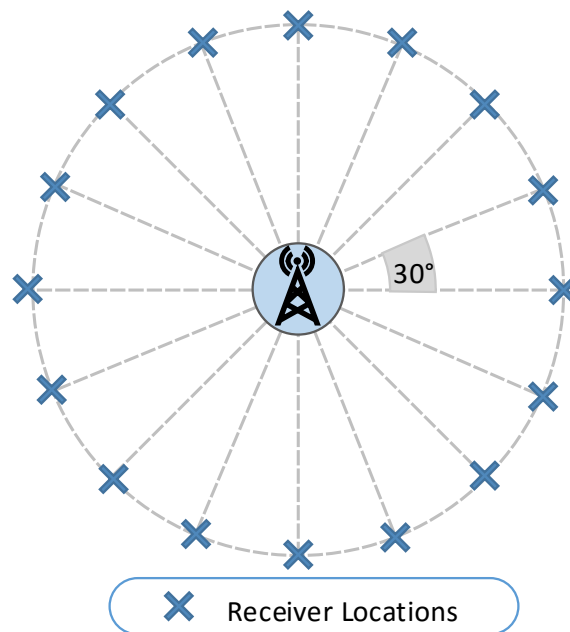


Figure 3.8: Illustration for spatial displacement.

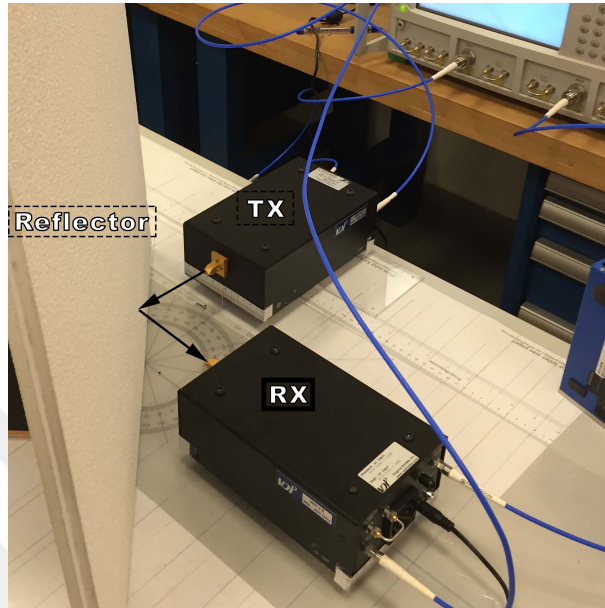


Figure 3.9: Measurement setup for NLOS experiments with a reflective surface.

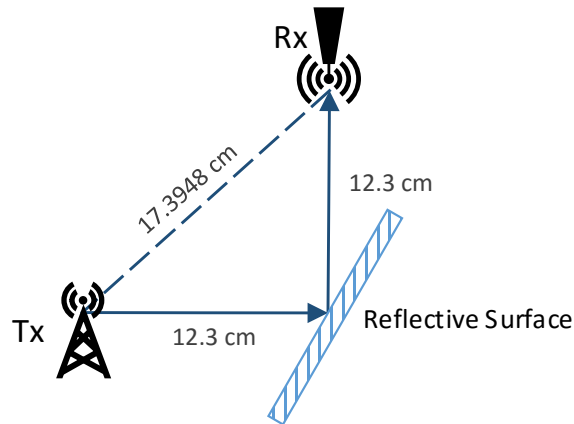


Figure 3.10: Illustration for NLOS experiments with a reflective surface.

measured channel depend on the antenna radiation pattern and the location. In this section, we statistically determine a large-scale propagation model for an indoor desktop table.

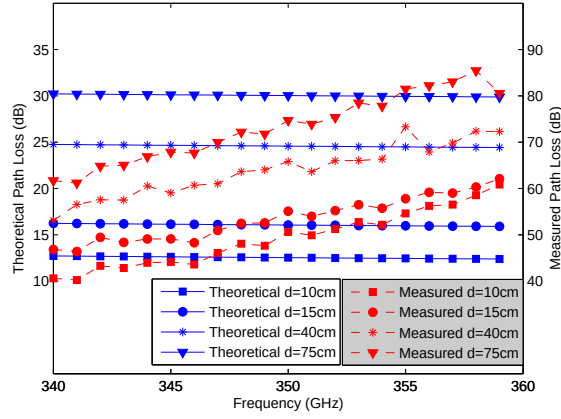


Figure 3.11: Uncalibrated measured and theoretical path losses with transceiver response included.

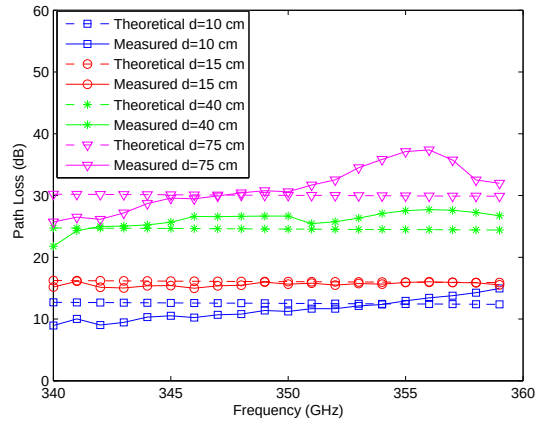


Figure 3.12: Measured and theoretical path losses after calibration of VNA.

3.3.1 Path loss model and shadowing

To determine the path loss model for a desktop channel, we have measured s-parameters of the channel with antennas attached at the transmitter and receiver for different

distances. The path loss at a particular distance can be referred as,

$$PL = \frac{P_t}{P_r} = -S_{21}. \quad (3.1)$$

where P_t and P_r are transmit and received power, respectively and S_{21} are the s-parameters. The local mean path loss is obtained by averaging measured impulse response over time and frequencies, i.e.,

$$\overline{PL}(d) = \frac{1}{MN} \sum_{i=1}^N \sum_{j=1}^M |H(f_i, t_j, d)|^2, \quad (3.2)$$

where, $H(f_i, t_j, d)$ is the measured impulse response of the channel, N is the number of observed frequencies, M is the number of times impulse response is measured and d is the distance in meters.

LOS channels make the case for simplest channel model and a sample measurement setup is shown in Fig. 3.5. In these experiment, we measured path loss at different distances ranging from 10 cm to 75 cm, that are often expected in a typical office desktop table or a TV stand. All the measurements were performed on a wooden base and the antenna height in the system was kept at 7 cm. The number of frequency-response snapshots over time was $M = 10$ which was observed to be adequate for temporal analysis.

To appreciate the importance of calibration in our setup, we have shown s-parameters of different distances at frequencies ranging from 340 to 359 GHz for the case when the transceiver was not calibrated, in Fig. 3.11. We can observe that the path loss increases with frequency and does not follow the theoretical path loss. The reason for this apparent aberration is the frequency-dependent response of the transceiver shown previously in Fig. 3.2. Note that the response includes antennas with a gain of 25 dBi and is hence considered as a part of system. Therefore, while considering channel transfer function, the gain of antennas may be subtracted using Fig. 3.4. After the calibration was performed, the results were fairly improved and a better

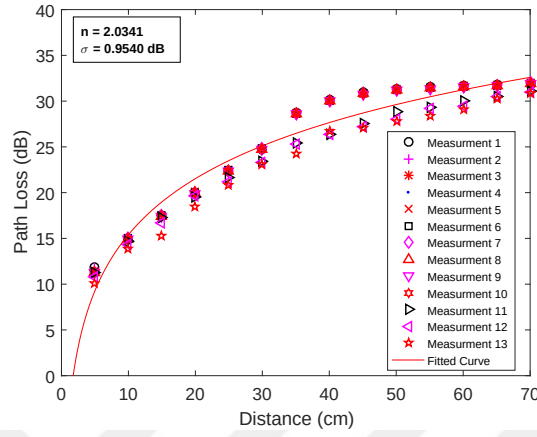


Figure 3.13: Path loss versus transmitter receiver separation in LOS environment for 300-319 GHz.

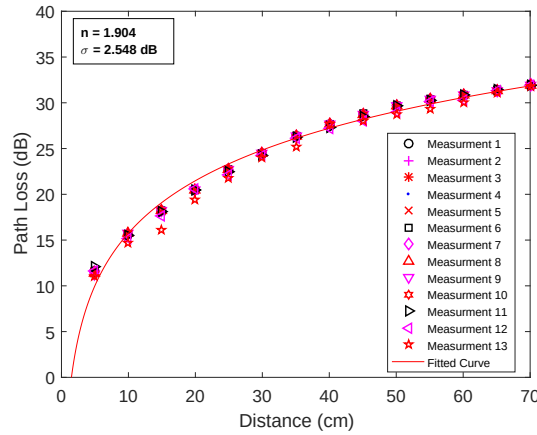


Figure 3.14: Path loss versus transmitter receiver separation in LOS environment for 340-359 GHz.

match between the measured and theoretical path loss was observed, as shown in Fig. 3.12. However, the path loss variations across the frequency indicate that extensive equalization would be required for such a wide band system. Moreover, increased loss at higher distance was observed and can be attributed to the higher system noise floor.

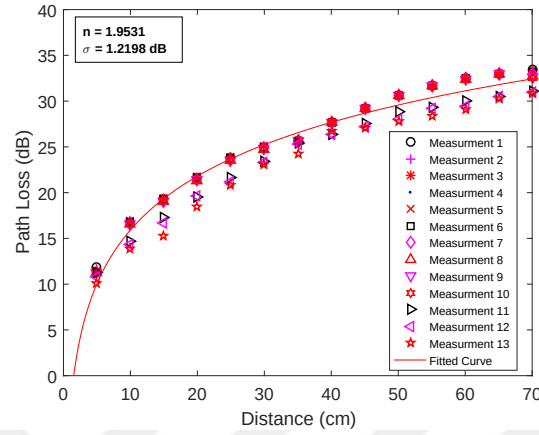


Figure 3.15: Path loss versus transmitter receiver separation in LOS environment for 380-399 GHz.

The path loss for different distances on desktop tables with different surface materials was measured at different times and mean path loss was calculated and shown in Fig. 3.13 for 300-319 GHz band measurements. It can be observed in Fig. 3.13 that the temporal variations over time are minimal. Similar results were observed for 340-359 GHz band and 380-399 GHz band. The first 10 measurements are temporal measurements in which the antenna was kept 7 cm above the table. While measurement 11, 12 and 13 correspond to measurements taken with antenna height of 1 cm above wood, aluminum and rubber surface respectively. Since the channel is static, there is no additional clutter added over time that would cause any significant variation. This concludes to the fact that 10 measurements are fairly sufficient to capture all temporal variations in a THz Band desktop channel.

For a more extensive path loss model, we observed the effects of shadowing in our channel by rotating our transmitter and receiver on several spatial positions as well as placement on the desktop tables with surfaces of wood, aluminum and rubber by placing them 3 cm above the antennas, to observe spatial variation. It was observed that, path loss on a wooden table was slightly less than the rubber followed by an aluminum table. The average values are summarized in Table 3.2. Moreover, we

Table 3.2: Path loss for different desktop surfaces.

Freq (GHz)	Free Space(dB)	Wooden (dB)	Aluminium(dB)	Rubber(dB)
300-319	15.3754	13.2688	14.28	13.3464
340-359	12.4336	10.8362	11.8505	11.4571
380-399	11.4274	9.7417	10.7517	10.4883

Table 3.3: Values of measured path loss exponent n and σ for low-THz Bands.

Freq (GHz)	n	Standard Deviation	σ (dB)	Standard Deviation(dB)
300-319	2.0341	4.417	0.9540	0.9767
340-359	1.9040	0.1229	2.548	1.59
380-399	1.9531	4.0732	1.2198	1.1044

placed different objects such as plexi glass, metal keys, mobile phone and human hand to observe the affect of all these obstructions. The cumulative distribution functions (CDF) of all the shadowing objects in 300-319 GHz, 340-359 GHz and 380-399 GHz band are shown in Fig. 3.16, Fig. 3.17 and Fig. 3.18, respectively. It can be observed that the shadowing follows normal distribution. For the CDFs, it was concluded that only optically transparent objects would allow the THz Band signal to pass through them and thus the opaque objects were not included in the shadowing variable. For an observed channel, path loss over a particular distance can be modeled by the path loss exponent model with log-normal shadowing as [Rappaport et al., 1996, Samimi et al., 2015, Sun et al., 2016, Xu et al., 2002, Kivinen, 2007]

$$\overline{PL}(d) = PL(d_0) + 10n \log\left(\frac{d}{d_0}\right) + X_\sigma, \quad (3.3)$$

where, $\overline{PL}(d)$ is the average path loss in decibels at the distance d , n is the path loss exponent which indicates the rate at which the path loss increases with d , $PL(d_0)$ is the path loss at reference distance d_0 and X_σ is a zero-mean Gaussian distributed random variable (in dB) with standard deviation σ (also in dB).

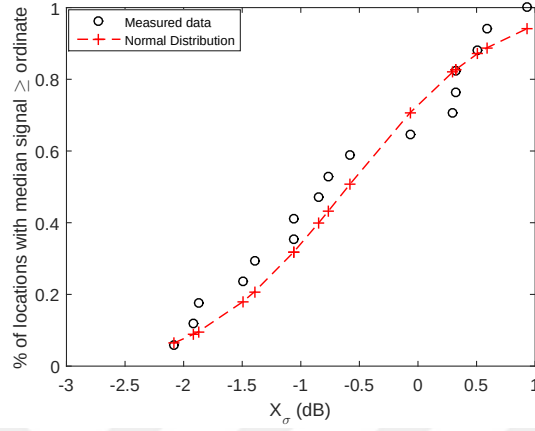


Figure 3.16: Scatter plot of shadowing CDF in 300-319 GHz.

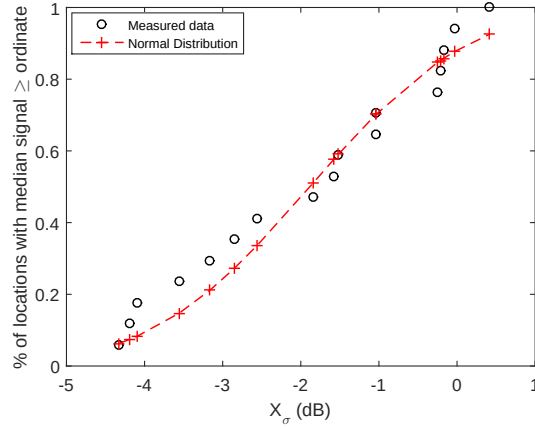


Figure 3.17: Scatter plot of shadowing CDF in 340-359 GHz

The values of n and σ in (3.3) are computed from the measured data, using least square linear regression fitting through the scatter of data such that the error between measured and estimated path loss is minimized in a mean square error sense over the numerous measurements for different transmitter-receiver distances. The model parameters were calculated for all three bands with the reference distance $d_0 = 1\text{ m}$. The free space path loss including the antenna gains at this distance is 33.37 dB, 32.52 dB and 31.97 dB, respectively for all these bands. The results are summarized

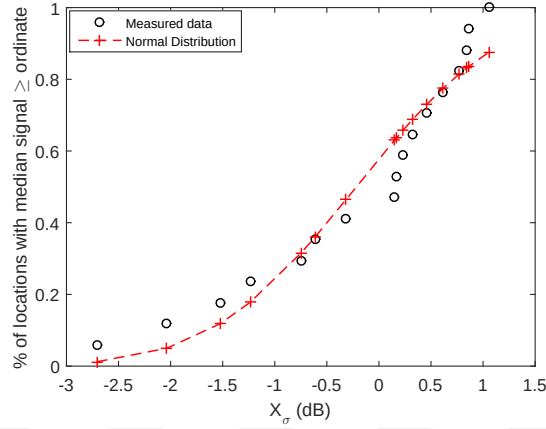


Figure 3.18: Scatter plot of shadowing CDF in 380-399 GHz.

in Table 3.3. The path loss exponent, n , is calculated as 2.0341, 1.9040 and 1.9531 for 300-319 GHz, 340-359 GHz and 380-399 GHz bands, respectively. The values of σ along with Fig. 3.16, 3.17 and 3.18 show that the shadowing X_σ is indeed Gaussian.

3.3.2 Non-line-of-sight Channel Measurements

The indoor desktop communication channel that we have considered in this work is susceptible to significant losses due to the presence of shadowing [Goldsmith, 2005]. Mainly shadowing is caused by any obstacle appearing between transmitter and receiver. This obstacle attenuates the signal power depending on the material type. A strong attenuation can even block the signal and stop communication. Absorption coefficient of different materials was determined using the transmission loss. The attenuated signal power was subtracted from the reference signal to get added transmission loss due to the object. Finally, this extra loss was used to calculate the absorption coefficient of the material [Khalid and Akan, 2016b]. For instance, a clear glass with thickness of $t_{glass} = 2 \text{ mm}$ or a 80 gram fiber mat with thickness of $t_{fiber} = 3 \text{ mm}$, attenuation of approximately 6.14 dB and 27.55 dB, respectively, were measured at 310 GHz. In the case of a human hand placed as obstruction, 45.6 dB of attenuation was observed. The attenuation coefficient of different materials is summarized in the

Table 3.4: Absorption coefficient of different objects

Object	α_{310GHz}	α_{350GHz}	α_{390GHz}
<i>Fiber mat</i>	0.169 mm^{-1}	0.171 mm^{-1}	0.173 mm^{-1}
<i>Glass</i>	0.049 mm^{-1}	0.048 mm^{-1}	0.049 mm^{-1}
<i>Empty water glass</i>	0.959	0.962	0.965
<i>Hand</i>	0.990	0.982	0.994

Table 3.4.

The results show that, if a thin glass or a fiber sheet appears as obstruction in the communication link, the SNR of channel decreases, and hence, the data rate also decreases. However, if obstruction is a human hand, the intense signal attenuation might lead to disconnection of communication link. To reestablish a link in such scenarios, the system should turn to NLOS paths using reflection properties of the materials present in the environment. In a typical indoor environment, objects like metallic surfaces of aluminum and copper coated with paint and ceiling panels are ubiquitous. Using these three materials as reflectors, we measured NLOS path loss with the experimental setup discussed in Fig. 3.9 and 3.10. The reflecting surface was equidistant from both the transmitter and the receiver at a distance of 12.3 cm making the effective distance our signal travels following a reflective path is 24.6 cm. Fig. 3.19 compares the measured NLOS path loss for paint-coated aluminum, copper and acoustic ceiling panel with the theoretical path loss. It can be seen that the signals reflected from these objects are strong enough to provide a NLOS in a desktop channel with the metal plates working almost equivalent to the theoretical path loss expectation. Ceiling panels do not perform as well as metals due to their rough surfaces but still the signal reflected by them is above noise level in any case.

It is important to note that if the reflective surface is moved to a different position, a corresponding change in the path loss may occur based on the new effective distance between the transmitter and the receiver as well as antenna properties such as the

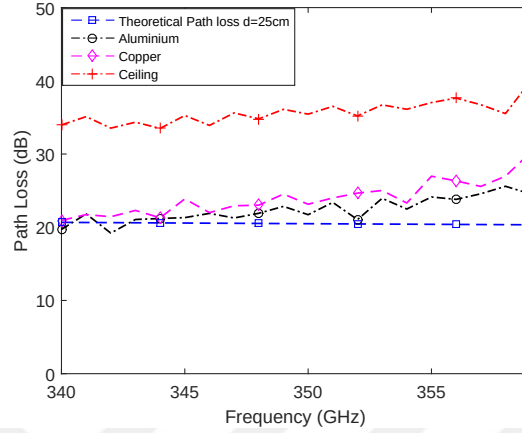


Figure 3.19: NLOS communication with different reflective surfaces.

direction of the antenna beam width and the shape of the antenna beam.

3.4 Small-Scale Multipath Propagation

Multipath in a wireless channel creates small-scale fading effects that manifest as rapid changes in the signal strength. It happens when a transmitted signal reaches the receiver from different directions with different propagation delays. These waves add up vectorially at the receiver antenna and can cause severe distortion or fading. Such a channel impulse response can be characterized as [Rappaport et al., 1996]

$$h(t, \tau, d) = \sum_{k=1}^L a_k(t, d) \exp(j\theta_k(t, d)) \delta(t - \tau_k), \quad (3.4)$$

where τ_k is the excess delay of the k th multipath component relative to the first arrival, L is the number of multipath components, a_k represents the amplitude of the k th component, θ_k is its phase and $\delta(\cdot)$ is the Dirac delta function.

Using the dual relationship of time domain and frequency domain technique, the channel impulse response was calculated by taking inverse discrete Fourier transform (IDFT) of the measured frequency response. This helps in determining the time dispersive properties of a wide band multipath channel. The common parameters used to quantify the dispersive property are the mean excess delay (τ_m) and RMS

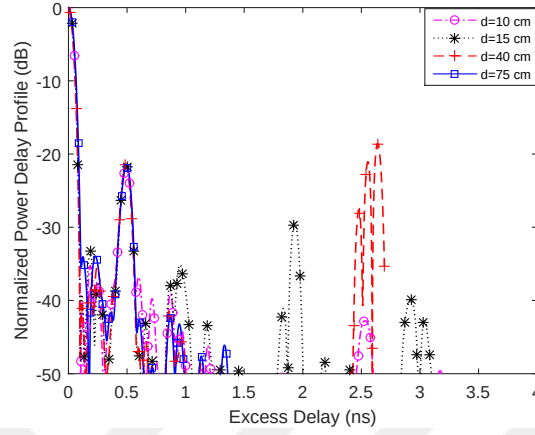


Figure 3.20: Normalized power delay profile in LOS channel.

delay spread (σ_τ). The mean excess delay is the first moment of the power delay profile and is defined as

$$\tau_m = \frac{\sum_{k=1}^L a_k^2 \tau_k}{\sum_{k=1}^L a_k^2}, \quad (3.5)$$

whereas the RMS delay spread is the second central moment of the power delay profile and is defined as

$$\sigma_\tau = \sqrt{\tau_c^2 - \tau_m^2}, \quad (3.6)$$

where,

$$\tau_c^2 = \frac{\sum_{k=1}^L a_k^2 \tau_k^2}{\sum_{k=1}^L a_k^2}. \quad (3.7)$$

These equations only rely on the amplitudes that are relative to the first detectable signal arriving at the receiver and do not rely on the absolute power level of the received signal. Thus, the impulse response is normalized to clearly observe the power delay profile and excess delay. The noise threshold level is chosen 10 dB above the system noise level which comes out to be approximately -50 dB. The mean values of τ_m and τ_{RMS} are 44.2926 ps and 231.435 ps, respectively for the 340-359 GHz band.

Table 3.5: Time dispersion parameters in LOS channel.

Freq (GHz)	Distance (cm)	τ_m (ps)	σ_τ (ps)	τ_{max} (ns)	B_c (GHz)
300-319	10	6.88	71.48	2.50	2.79
	15	14.41	114.87	1.92	1.74
	40	58.76	375.00	2.62	0.53
	75	104.85	410.43	2.74	0.50
340-359	10	3.99	44.58	0.87	4.48
	15	7.02	84.62	2.92	2.36
	40	58.21	388.56	2.63	0.51
	75	107.94	407.97	2.70	0.49
380-399	10	18.92	92.31	0.91	2.16
	15	23.01	117.86	2.93	1.69
	40	54.18	312.03	2.63	0.64
	75	103.90	409.65	2.74	0.53

The power delay profile for 10 cm, 15 cm, 40 cm and 70 cm transmitter-receiver separation is shown in Fig. 3.20. The results show that there are multipath signals, above the threshold value, arriving after the main signal. These mulipaths reduce the coherence bandwidth (B_c) of the channel which is the bandwidth over which the channel can be considered flat. For a bandwidth above which correlation function is above 0.5, $B_c \approx 1/5\sigma_t$ [Rappaport et al., 1996]. The τ_m , τ_{RMS} , the maximum excess delay (τ_{max}) for -40 dB threshold and B_c for the various frequency bands are summarized in Table 3.5. The results show that the B_c is much less than the actual bandwidth and utilizing the full bandwidth efficiently would not be straight forward without stringent channel equalization. Furthermore, B_c decreases with the increase in the distance between the transmitter and receiver due to normal path losses.

Finally, we also analyze the temporal autocorrelation function in Fig. 3.21 for the frequency band of 340-359 GHz. The normalized temporal correlation function is

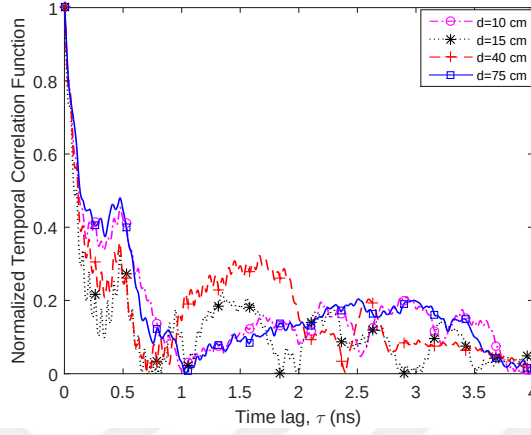


Figure 3.21: Normalized temporal correlation function in LOS channel.

calculated using

$$R(\tau) = \frac{E[h(t)h(t+\tau)^*]}{\sqrt{E[|h(t)|^2]E[|h(t+\tau)|^2]}}, \quad (3.8)$$

where $(\cdot)^*$ denotes complex conjugate operation, and $E[\cdot]$ is expectation operator. The results in Fig. 3.21 show that the correlation function drops below 0.2 after 0.71 ns for all the tested distances. However, at some distances in between, the function value increase above 0.2 and the signal may get correlated again. Similar results were generated for the other two bands as well.

3.5 Conclusion

In this chapter, we have provided and compared the statistical characteristics of several low-Terahertz Band channels, namely, the 300 to 319 GHz band, 340 to 259 GHz band and 380 to 399 GHz band. The experiments were carried using a Vector Network Analyzer and subharmonic mixers based channel sounder. Testbed specifications, configurations and site descriptions are discussed in detail. Around 600 different measurements were recorded with spatial and temporal variations, that included LOS and NLOS environments, for all the three potential bands. The large-scale path loss models for all the three bands were evaluated using the single slope path loss model

with shadowing. Our work shows that the models are quite comparable as the path loss exponent for 310 GHz band, 350 GHz band and 390 GHz band is 2.07, 1.90 and 1.96, respectively, while the variations due to shadowing were confirmed to be normally distributed, for all the three bands. Moreover, NLOS measurements were performed by using the optically opaque objects as reflectors to measure the path loss. It was observed that metals, wood and even acoustic ceiling panel can be used as a reflector in THz Band channel.

Furthermore, small-scale multipath propagation was also statistically analyzed in an indoor environment. The channel impulse response was calculated by taking IDFT of the measured frequency response. It was observed that strong multipath components were present in LOS and NLOS environments. To describe the multi path properties of the channel, we have calculated mean excess delay, RMS delay spread and the coherence bandwidth. The results show that the high RMS delay reduces the coherence bandwidth a lot, making it necessary to use equalizers in THz Band communications. This finding suggests that efficient equalizer should be designed in order to fully benefit from the wide available bandwidth of THz Band channel. Finally, we have also calculated the autocorrelation function of the channel impulses response and it was observed that at times the channel might show correlation higher than 0.2.

By applying virtual antenna array technique on the similar equipment, first results of a Terahertz (THz) Band line-of-sight (LOS) 2×2 multiple-input multiple-output (MIMO) channel were presented in [Khalid and Akan, 2016a]. This progress shows the potential of THz Band and further encourage its usage in 5G Indoor Internet of Things.

Chapter 4

ENERGY-EFFICIENT MODULATION SCHEME

Terahertz Band (0.3-10 THz) communication is envisioned to accommodate high throughput wireless communication that will open plethora of new applications such as Internet of things (IoT), wireless personal area networks (WPAN) and ultra-high-speed wireless backhaul. Nodes in these applications will communicate massive amount of data that will consume a lot of energy. Hence, hardware of these applications must be designed so that minimum energy is used in transmitting data while satisfying the given throughput and delay limits. In this context, we analyze the best modulation scheme for these applications to minimize the total energy consumed while sending given number of bits. The total energy consumption for MQAM and MFSK system is modeled in detail so that it includes both the transmission energy required on a THz Band channel and circuit energy consumption. By numerical calculations, we have shown that for low distance communication networks, femtocells, MQAM is the most energy efficient scheme that saves up to 90 % energy.

4.1 Introduction

Consumers of wireless communications are growing exponentially and demands for ubiquitous access to information and entertainment content at higher data rates are being intensified. According to a recent survey conducted by Ericsson, the mobile subscription has reached 7.5 billion and it is anticipated that by 2020, it will increase by 23%. As smartphones constitute the majority of subscribers utilizing wireless broadband services, further development of smartphone applications will increase data usage such that the worldwide data usage per smartphone is anticipated to reach 25000 petabytes until 2020 [Ericsson, 2016]. This huge increase will require

very high data transfer rates. The current wireless systems support data throughput up to few hundred megabits per second, which is quite low in comparison to the wired communication standards such as Gigabit Ethernet and high-definition multimedia interface (HDMI) and anticipated future requirements. Furthermore, wireless data throughput is approaching the capacity of the current allocated spectrum and hence, to support the increasing demand in wireless communications, higher frequency bands must be exploited. Among the high frequency candidates, current research is more focused on 60 GHz ISM band which can provide data throughput up to few gigabits per second due to its limited bandwidth. To fulfill the anticipated requirements, wireless technology must shift to an even higher spectrum where the bandwidth can reach from few to several GHz. In this context, THz Band is the most suitable candidate as it can provide contiguous bandwidth of several GHz or even up to THz and thus it must be included in communication spectrum of 5G wireless networks [Akyildiz et al., 2014, Khalid and Akan, 2016b, Yilmaz and Akan, 2015a, Huang and Wang, 2011, Yilmaz and Akan, 2015b].

Research activities in THz Band are increasing, already channel characterization and semiconductor device behaviors are being examined [Khalid and Akan, 2016a]. As THz Band communication is still in its infancy, the standardization process has not yet begun. Unlike the 60 GHz band, modulation schemes for THz Band communications are yet to be defined. In [Han and Akyildiz, 2014], a distance aware modulation scheme is proposed that uses distance and frequency dependent behavior of THz Band channel to adaptively utilize the bandwidth. An important factor in selection of modulation scheme for THz Band is high amount of energy required for transmitting the huge amount of data. Hence, minimization based on energy consumption factor is necessary to make transmission viable and cost effective. Most of the previous research on energy consumption in communication has focused transmitting small amount of data at low frequency. In [Cui et al., 2005], energy consumption analysis of communication system operating at 2.5 GHz is analyzed by including transmission and circuit energy consumption. It is shown that for short-range appli-

cations, MQAM is more energy efficient as compared to MFSK. In this chapter, we study cases where huge amount of data is being transmitted in THz Band and determine the best possible modulation scheme that consumes minimum energy while transmitting the data. Components use in our model to perform the analysis are based on semiconductor technology, as for inexpensive products and bulk fabrication, Si CMOS and SiGe are the best solutions.

The remainder of this chapter is organized as follows. In Section 4.2 we describe our THz Band transceiver model used for energy consumption analysis. In Section 4.3 we perform the energy consumption analysis of MQAM and MFSK modulation scheme. In Section 4.4 we optimize modulation parameters and compare the results of MQAM and MFSK modulation at different distances. Finally, we conclude the chapter in Section 4.5.

4.2 THz Band System Model

In order to analyze energy consumption of the system, we have modeled a wireless communication link between two nodes operating in THz Band. Generally, a wireless communications system can be decomposed into five major components: the transmitter, receiver, synthesizer, power amplifier and antenna. The specification of each block determines the energy consumed by the system. In addition, all signal processing blocks need to be considered in energy consumption analysis. In THz Band, signal generation is a challenge and requires a special synthesizer and thus it is examined in detail. On the other hand, bandwidth in THz Band is huge and small electrical length enables highly directive communication. Therefore, complicated signal processing techniques such as multiuser detection or Time Division Multiplexing (TDM) are not very advantageous and hence neglected in our model. The resulting system model of the THz Band transmitter and receiver is shown in Figure 4.1.

In the THz Band communication system, the signal synthesizer mainly consists of a Local Oscillator (LO), balun, buffer amplifier and a multiplier [Chakraborty et al., 2014]. The multiplying factor depends on the frequency required to be generated.

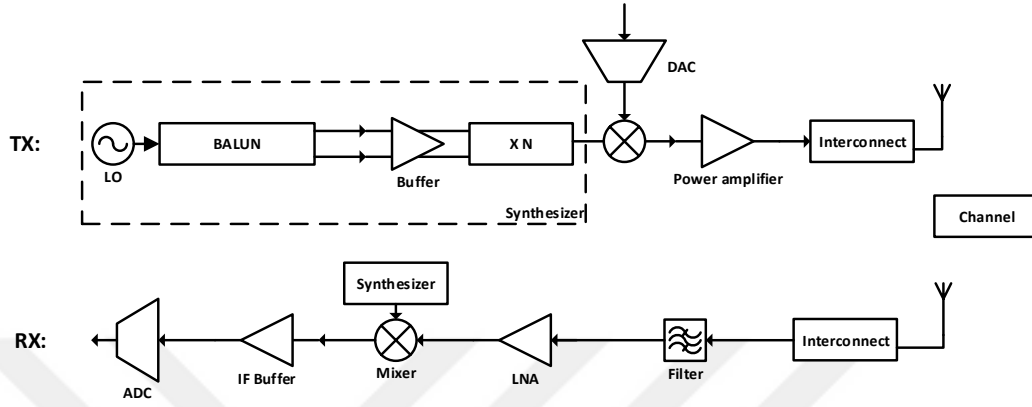


Figure 4.1: THz Band transceiver architecture.

First, the baseband digital signal is converted to analog signal using a digital-to-analog (DAC) converter. This signal then modulates the frequency synthesizer signal being fed into the mixer. The modulated signal is then amplified using a power amplifier (PA) and transmitted into the channel through the antenna. Usually, antennas in THz Band are connected to the circuit using special interconnects for which power dissipation is considerable and cannot be neglected. While on the receiver side, the THz signal is first filtered inherently by the interconnect and then amplified by the low noise amplifier (LNA). The amplified signal is then downconverted by the mixer and amplified again by the intermediated frequency (IF) buffer. Finally, the IF signal is converted back to digital signal using analog-to-digital converter (ADC).

In our analysis, we considered that the transceiver operates on hybrid mode basis. This implies, when there is a signal to transmit, all the transceiver components operate in active mode and consume full power. On the other hand, when there is no signal to transmit, the transceiver switches to inactive mode, which saves significant energy. When switching from inactive mode to the active mode, there is a transient mode, which is the time required to turn all the components into active state, usually dominated by the phase lock loop (PLL).

To calculate power consumption, a simplified case of wireless communications

is considered in which a node has I bits to transfer in time T . The transmitter turns on for time T_{act} to transmit these bits and then goes back to inactive mode. The parameter T_{act} is active mode time and needs to be optimized based on the modulation scheme. The transmission period consists of $T = T_{act} + T_{settl} + T_{off}$, where T_{settl} is transient time whose duration depends mainly on the PLL settling time in the frequency synthesizer and T_{off} is the inactive mode duration. Thus, the total energy consumption E_{total} required to send I bits is written as

$$E_{total} = P_{act}T_{act} + P_{off}T_{off} + P_{settl}T_{settl}, \quad (4.1)$$

where, active mode power P_{act} is composed of transmitted power and circuit power consumption. In the circuit power consumption, we include the power consumed by LO, balun, buffer amplifier, mixer, low noise amplifier, filter, IF buffer, ADC, DAC and PA. The power consumption of PA is $P_{amp} = \alpha P_t$ where $\alpha = (\xi/\eta) - 1$, η is the drain efficiency and ξ is the peak-to-average power ratio (PAPR) depends on the modulation scheme. Thus, we can write the corresponding active power as

$$P_{act} = P_t + P_{amp} + P_{ct} + P_{cr}, \quad (4.2)$$

where, P_{ct} is the circuit power consumption at the transmitter and P_{cr} is the circuit power consumption at the receiver. Moreover, $P_{ct} = P_{LO} + P_{balun} + P_{buffer} + P_{\times N} + P_{mix} + P_{DAC} + P_{interconnect}$ and $P_{cr} = P_{LO} + P_{balun} + P_{buffer} + P_{\times N} + P_{mix} + P_{LNA} + P_{interconnect} + P_{IFBuffer} + P_{ADC} + P_{fil}$.

The inactive mode power consumption is orders of magnitude smaller than other states and can be neglected in the current comparison [Itoh et al., 2004]. On the other hand, when turning on the transmitter, the LO is first turned on until it settles and then other circuits are turned on. Thus, the transient state power depends on power consumed by LO at the transmitter and the receiver and we can approximate it to $P_{settl} \approx 2P_{LO}$.

Using all the above defined parameter, energy consumption per information bit,

$E_b = E_{total}/I$, is given by

$$E_b \approx \frac{(1 + \alpha)E_t + P_c T_{on} + 2P_{LO}T_{tr}}{I}, \quad (4.3)$$

where, $E_t = P_t T_{act}$ is the energy transmitted and $P_c = P_{ct} + P_{cr}$ is the circuit power consumption while P_{LO} and T_{settl} are the constants. Here optimal T_{act} needs to be determined in order to obtain the optimal constellation size for any modulation scheme. It can be seen from the equation that energy consumption per bit increases as the PAPR factor of the modulation scheme increases. Hence, the modulation scheme must be selected very carefully in order to optimize the energy consumption of the system. In addition, [Cui et al., 2003] has shown that energy per bit increases with the increase of bandwidth efficiency, defined by $B_e = \frac{I}{BT_{act}}$.

Owing to the small distances in a femtocell and spectral window centered around 300 GHz, atmospheric absorption can be neglected [Akyildiz et al., 2014]. Hence for simplicity, we have used γ th power path loss model, in which transmitted power, P_t is expressed as

$$P_t = P_r d^\gamma M_l G_1, \quad (4.4)$$

where G_1 is the gain factor for unit distance. This includes the gain of transmitting antenna, receiving antenna and the path loss at one meter. For an indoor office environment, γ is considered to be 4 [Huang et al., 2016]. Also, path loss for one meter distance at THz frequency is approximately 82 dB [Khalid and Akan, 2016b], whereas the antenna gain of 22 dBi is easily realizable [Li et al., 2015]. Correspondingly, $G_1 = 38$ dB is calculated and used in our analysis.

In THz Band, bandwidth is not scarce and hence to transmit huge amount of data, most energy efficient modulation scheme should be used. To find the optimum modulation scheme, the above equations are evaluated for MQAM and MFSK in the following section.

4.3 Energy Consumption Analysis of MQAM and MFSK Modulation

We first analyze the THz Band communication system model with MQAM modulation scheme and then with MFSK modulation scheme over an additive white Gaussian noise (AWGN) channel.

4.3.1 MQAM modulation scheme

In an MQAM system, symbols required to send I bits are found using $I_s = I/b$, where $b = \log_2 M$ are the number of bits assigned per symbol. Moreover, I_s can also be represented in terms of symbol period as $I_s = T_{act}/T_s$. If the system utilizes square pulses, then we have $T_s \approx 1/B$ and hence the relationship can be defined as

$$b \approx \frac{I}{BT_{act}} = B_e. \quad (4.5)$$

Whereas the required received power for specific constellation size, noise figure, bandwidth and probability of error for MQAM operating in AWGN channel is determined using the upper bound on error probability [Proakis, 2001] given by

$$P_r = \frac{4}{3} N_f B \sigma^2 (2^b - 1) \ln \frac{4(1 - \frac{1}{\sqrt{2^b}})}{b P_b}, \quad (4.6)$$

where, N_f is the receiver noise figure, σ^2 is the power spectral density of AWGN channel and P_b is the probability of bit error.

For MQAM modulation scheme, PAPR is calculated using $\xi = 3(\sqrt{M} - 1/\sqrt{M} + 1)$ and $M = 2^{I/BT_{act}}$. Finally, by substituting all the necessary parameters for MQAM in (4.3), energy per bit is expressed as

$$E_b = \frac{4}{3} (1 + a) N_f \sigma^2 (2^b - 1) \frac{G_d}{b} \ln \frac{4(1 - \frac{1}{\sqrt{2^b}})}{b P_b} + \frac{P_c T_{act} + 2 P_{LO} T_{settl}}{I}. \quad (4.7)$$

Circuit parameters for THz Band transceiver are given in Table 4.1 [Instruments, 2009, Huang et al., 2016, Instruments, 2013, Thomas et al., 2005, Samoska, 2011, Diebold et al., 2015, Khalid et al., 2017a]. Due to high PAPR, MQAM requires

Table 4.1: System Parameters used for Analysis.

Parameters	Value	Parameters	Value
η_{MQAM}	0.15	γ	4
G_1	38 dB	M_l	40 dB
B	1 GHz	P_{mix}	7.9 mW
P_{LNA}	6.75 mW	P_{maxt}	53 mW
T_{tr}	10 μ s	T_{MQAM}	100 ms
N_f	10 dB	f_c	300 GHz
σ^2	-174 dBm/Hz	I	200 Mb
P_{LO}	100 mW	P_{fil}	1 mW
P_b	10^{-3}	T_{MFSK}	1.07 s
P_{ADC}	1.4 W	η_{MFSK}	0.35
P_{DAC}	2.5 mW	P_{balun}	1.2 mW
$P_{IFBuffer}$	5 mW	P_{buffer}	8 mW
$P_{\times N}$	13 mW	$P_{interconnect}$	1.8 mW

linear power amplifier. Amplification at THz Band is hard to achieve using solid state devices due to their low transistor cut off frequency. Therefore, class-A power amplifier is selected whose efficiency is $\eta_{MQAM} = 0.15$ [Diebold et al., 2015]. While, I is found using $b_{min} = I/(BT)$ and the fact that minimum constellation size, b_{min} , is 2. Here, bandwidth B and time period T are kept for specific case of IoT.

To determine the optimum active time for transmitting the data at different distances, Energy per bit is plotted against the normalized T_{act} in Fig. 4.2. It can be seen that for small distance, the energy consumption is less for small T_{act} and hence significant amount of energy can be saved using optimum T_{act} .

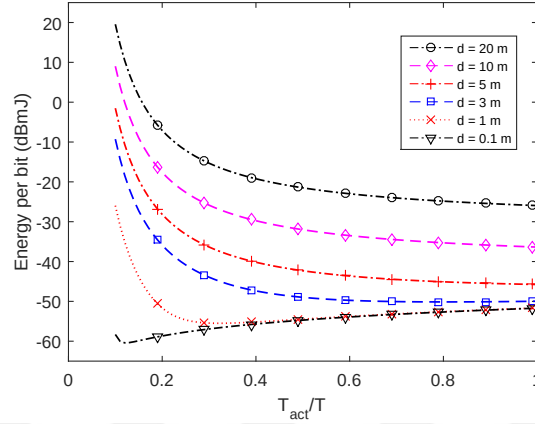


Figure 4.2: Energy consumption per bit at different distances versus normalized T_{act} , for MQAM operating in THz Band channel.

4.3.2 MFSK Modulation scheme

In MFSK, the number of orthogonal carriers equals the constellation size given by $M = 2^b$, with minimum inter-carrier separation of $1/(2T_s)$. Hence, the total bandwidth is $B \approx 2^b/(2T_s)$ and the bandwidth efficiency is $B_e = 2b/2^b$. By comparing bandwidth efficiency of MFSK with general equation of bandwidth efficiency $B_e = \frac{I}{BT_{act}}$, we obtain the following relation for T_{act} in MFSK

$$T_{act} = \frac{2^b I}{B 2b}. \quad (4.8)$$

The relation between, required received power for a given probability of error, bandwidth, noise figure and constellation size for MFSK can be determined using the upper bound on error probability [Proakis, 2001]. The corresponding equation is given as

$$P_r = 4N_f \sigma^2 \frac{2B}{2^b} \ln \frac{2^{b-2}}{P_b}. \quad (4.9)$$

Peculiar nature of MFSK modulation requires some modification in circuit parameters and the model. In MFSK, $\xi = 1$ due to constant signal amplitude. This allows the amplifier to be used in saturation and achieve maximum efficiency. Furthermore,

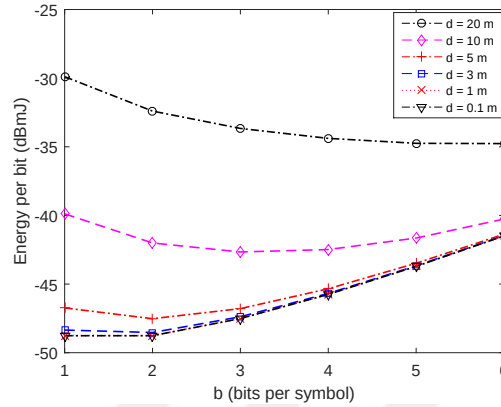


Figure 4.3: Energy consumption per bit at different distances versus bits per symbol (b), for MFSK operating in THz Band channel.

it allows the use of non-linear amplifiers, such as class-B amplifier, that provides better drain efficiency [Khalid et al., 2015]. Hence, $\eta = 0.35$ is kept for THz Band power amplifier. On the other hand, in the transmitter, we need to remove mixer and DAC since frequency modulation is carried directly in the synthesizer. Finally by plugging all the required parameters for MFSK in (4.3), energy per bit is expressed as

$$E_b = 4(1 + \alpha)N_f\sigma^2 \ln \frac{2^{b-2} G_d}{P_b b} + \frac{P_c T_{act} + 2P_{LO} T_{settl}}{I}, \quad (4.10)$$

where, bandwidth and the packet size for MFSK were kept similar to MQAM configuration. However, the maximum time delay, T_{MFSK} , was changed to 1.07 s, which restricts maximum value of b to 6. Unlike MQAM, MFSK modulation is less bandwidth efficient and thus it requires more transmission time for same number of bits and for a given bandwidth.

Energy consumption per bit for MFSK modulation scheme is shown in Figure 4.3. It can be seen that for small distances, smaller value of b , bits per symbol, is more energy efficient and increasing b , increases energy consumption. This is due to the effect of circuit power consumption, which increases for higher b . On the other hand, for distances higher than 10 meters, the transmission power dominates the circuit power consumption and thus higher b becomes more energy efficient. The reduction

Table 4.2: Optimized parameters for MQAM /MFSK in THz Band channel for various distances.

$d(m)$	0.1	1.1	5.1	10.1	20.1	25.1
b_{opt}	16.66 /2	5.88 /2	2 /2	2 /3	2 /6	2 /6
T_{act} (s)	0.012 /0.2	0.034 /0.2	0.1 /0.2	0.1 /0.26	0.1 /1.067	0.1 /1.067
E_b (nJ)	0.909 /13.29	3.04 /13.31	27.94 /18	238.3 /55.47	2581.2 /338.1	5609.1 /652.47
Rate (Gbps)	16.66 /1	5.88 /1	2 /1	2 /0.75	2 /0.1875	2 /0.1875

in transmission energy at higher distance is not surprising because in an AWGN channel, the larger the M is, the more energy efficient the MFSK [Proakis, 2001]. Consider $d = 10$ m case, where by using $b = 4$, about 50 % energy saving is achieved.

4.4 Modulation Parameters Optimization

MQAM is a bandwidth efficient modulation scheme and thus it requires less time to transmit a given amount of data. The energy consumption analysis of MQAM in THz Band shows that for distances less than 1 meter, active state, T_{act} , should be small enough to reduce energy consumption, Fig. 4.2. However, for distance greater than 3 meters, a higher T_{act} will reduce energy consumption per bit. This is due to the fact that at higher distance, transmission power overrides the effect of circuit power. Hence, above 3 meters of distance, optimizing T_{act} will no longer help in energy saving and therefore maximum value of T_{act} should be used.

In terms of transmission time, MFSK is clumsy. Thus the time required to send a certain amount of bits is higher. On the other hand, the constant envelope ($\xi = 1$) significantly relaxes the power amplifier linearity requirement and thus power efficient transmission can be achieved. The energy consumption analysis of MFSK in THz Band channel shows that for distance smaller than 7 meters, $b = 2$ is the optimum

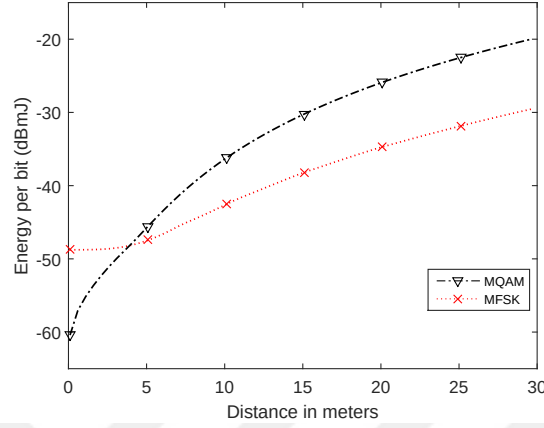


Figure 4.4: Energy consumption per bit at different distances for optimized MQAM and MFSK.

choice, Fig. 4.3. Whereas for distance greater than 7 meters, higher value of b yields better energy efficiency.

A comparison of transceiver energy consumption using MFSK and MQAM modulation scheme is shown in Figure 4.4. For each distance in MQAM transceiver, optimum T_{act} value is used to minimize energy consumption. These parameters are obtained from the energy per bit analysis of MQAM. As an example, for 0.1 meter distance case $T_{act} = 12$ ms was used. Whereas for 4 meters of distance, $T_{act} = 100$ ms was used. Similarly, optimum b was selected for the case of MFSK modulation scheme energy consumption analysis. The comparison of both modulation schemes shows that for distance less than 4 meters, MQAM modulation requires less energy per information bit given optimum T_{act} is used. Whereas for distance greater than 4 meters, MFSK modulation is more energy-efficient and hence consumes less energy than MQAM modulation scheme operating in THz Band. The optimized parameters for MQAM/MFSK operating at different distances for THz Band network are given in Table 4.2.

The potential applications for THz Band can be divided into three main categories. The first category includes very short distance communication devices which includes

applications such as gadgets connected to a computer, ultra-high-definition (UHD) displays, ultra fast portable data drives or mobile data synchronization. In these applications, the distance between transmitter and receiver is very short, usually less than 2 meters. Hence, to minimize energy consumption of the information transmission, we propose to use MQAM modulation scheme since it saves more than 90 % of energy. While, the second category is THz Band 5G indoor wireless communication devices which will provide femtocells with ultra high speed communications within the coverage of 10 meters. Specific applications include UHD multimedia streaming, UHD video conferencing, ultra-high-speed wireless data transfer etc. Due to higher mobility and distance requirement in these applications, we propose to use MFSK modulation scheme to reduce energy consumption of the information transmission. Finally, the third category is the wireless back haul operating over the ranges of distance greater than 10 meters. Similar to the second category, the characteristics of this application also suggests the use of MFSK modulation scheme.

4.5 Conclusion

In this chapter, we have analyzed the total energy consumption of a wireless communication system operating in THz Band channel. System models are based on MQAM modulation and MFSK modulation scheme. THz Band applications are expected to provide high data throughput that will consume a lot of energy, therefore, we focus to devise modulation schemes that will minimize the energy consumption per information bit. Based on our analysis, we have shown that short-range THz Band applications such as wireless computer gadgets, should use MQAM modulation scheme. Whereas, medium and long range such as 5G cellular networks and ultra-high-speed wireless backhaul should use MFSK modulation scheme to achieve energy-efficient communications.

Chapter 5

300 GHZ BROADBAND TRANSCEIVER DESIGN

This chapter presents the architectural design of a 300 GHz transceiver system that can be used to explore the high speed communication opportunities offered by the Terahertz (THz) band for advanced applications of Internet-of-Things (IoT). We use low cost industry ready components to prepare a fully customizable THz band communication system that provides a bandwidth of 20 GHz that is easily extendable up to 40 GHz. Component parameters are carefully observed and used in simulations to predict the system performance while the compatibility of different components is ensured to produce a reliable design. Our results show that the receiver provides a conversion gain of 51 dB with a noise figure (NF) of 9.56 dB to achieve a data rate of 90.31 Gbps at an operation range of 2 meters, which is suitable for high speed indoor IoT nodes. The flexible design of the transceiver provides groundwork for further research efforts in 5G IoT applications and pushing boundaries of throughputs to the order of terabits per second (Tbps).

5.1 Introduction

The Internet is now ubiquitous and widely used for applications like information transfer, social networking, banking and online shopping. Emerging applications of IoT require information sharing technologies that connect smart machines to the Internet along with people. Such applications of IoT are expected to transform our world from household life to modern businesses [Mavromoustakis et al., 2016]. Many IoT applications such as video surveillance, ultra-high-definition UHD displays and wireless back haul demand high data rates that are expected to increase further. In order to develop the technologies to satisfy consumer needs, 5G wireless networks

need to realize data rates reaching terabits per second (Tbps) [Huang and Wang, 2011]. The 60 GHz spectrum is expected to increase data rates up to 6 Gbps [IEEE, 2009], however, the available spectrum around this frequency is limited to only 9 GHz, which may not be able to completely satisfy the high data rate demands. Therefore, there is a lot of interest in exploring THz band communications (0.3 - 10 THz). This motivates the implementation of transmission and reception frontends for THz band that operate above 300 GHz.

Higher frequencies can support higher data rates, however, being highly directional, they are more easily blocked by obstacles on their path. This provides a unique set of challenges for THz band communications. To overcome these problems, THz band equipment is constantly improving performance with current research activities. The experimental THz band equipment can be categorized mainly into two groups depending on the approach for generating THz band signals. In a top-down approach, photonics technology is used to achieve communication that is translated from electronic to optical frequencies, while in a bottom-up approach, the operating frequency of electronics components is increased to the order of THz [Piesiewicz et al., 2007]. The bottom-up approach is a preferred course for consumer designs because optical components are more expensive than silicon components used in the bottom-up approach. Apart from cost, the size of silicon-based components is also more compact. Commonly used THz band experimental setups are based on bottom-up approach involving subharmonic Schottky diode mixers [Khalid and Akan, 2016b, Khalid and Akan, 2016a].

The bottleneck of THz band communication is the technology supporting it, and thus, there is an enormous need to explore different techniques as potential candidates. As discussed earlier, encouraging results for THz band are demonstrated using frontends based on photonic and electronic technologies. Using a 250 nm Indium Phosphide (InP) double heterojunction bipolar transistor (DHBT), [Carpenter et al., 2016] shows the transmission of 48 Gbps using 64-QAM/QPSK Direct-Conversion Transceiver Chipset. In [Katayama et al., 2016], a 40 nm CMOS transmitter is

presented that uses 32-QAM to provide 17.5 Gbps per channel capability over six channels. A Gallium Arsenide GaAs-based 35nm metamorphic high electron mobility transistor (mHEMT) integrated transmitter and receiver designed using Monolithic Microwave Integrated Circuits (MMIC) is shown to provide a 64 gigabits per second (Gbps) point-to-point link at 300 GHz for a distance of 2 meters in [Kallfass et al., 2015b]. Although MMIC based designs are more robust from a manufacturing perspective, their in-flexibility and lack of customization makes them unsuitable for research efforts where a number of system parameters need to be optimized. Whereas for experimentation purposes, it is very important that the system customizable such that a single component block can be easily replaced or modified, unlike the case with MMICs.

Therefore, in this work, we provide a fully customizable transceiver design that is capable of operating with a 20 GHz bandwidth. The design uses state-of-the-art discrete components such as WR2.8 subharmonic mixer (SHM) to operate in THz band. Full transceiver design is discussed and simulations based on the components characteristics are performed to achieve first pass success. The results for single-tone analysis, two-tone analysis, phase noise, receiver noise and receiver gain are discussed. Simulation results show that the transceiver operates reliably at 2 meters and provides data rates of up to 90.31 Gbps. The transceiver design can be further modified with ease to operate with a 40 GHz bandwidth.

The remainder of this chapter is organized as follows. In Section 5.2, we discuss the design and selection of components for the transceiver. Section 5.3 describes the simulation results for the first pass success of our design. Finally, we conclude the chapter in Section 5.4.

5.2 Transceiver Design

To achieve high throughput from the bandwidth available in the THz band, an ultra-wideband frontend is required. The huge path loss in the band limits communication distances to only a few meters and also requires highly directive antennas to transmit

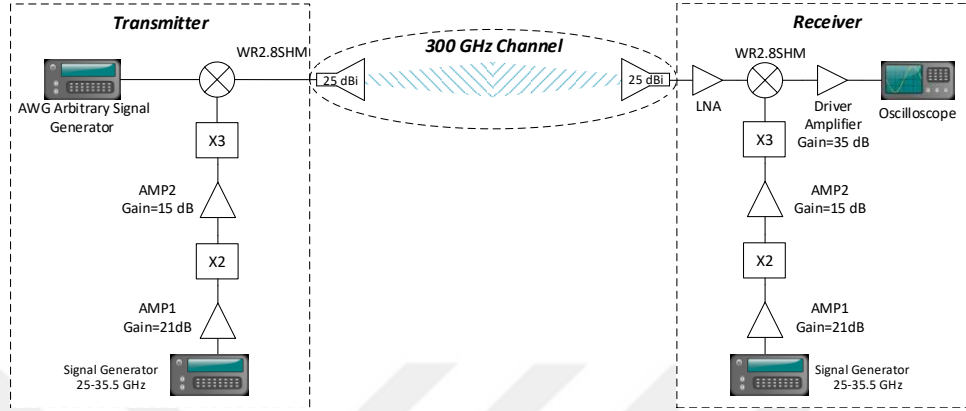


Figure 5.1: Simplified block diagram of the THz band transmitter and receiver.

the limited power generated from solid-state electronics. Therefore, during the design of the transceiver, our goal is to select ultra-wideband components that would provide at least 20 GHz of bandwidth and operate for a minimum distance of 2 meters.

To get an estimate of the characteristics of the required components, we first calculate noise floor values and the expected channel path loss for 20 GHz Bandwidth and a 2 meters channel length. The Jhonson noise power for a bandwidth (B) of 20 GHz in a wideband channel is calculated as -79.96 dBm using,

$$N_R = 10 \cdot \log\left(\frac{kTB}{10^{-3}}\right), \quad (5.1)$$

where, $k = 1.38 \times 10^{-23}$ J/K is Boltzmann's constant and T is the noise temperature in Kelvin. To predict the path loss, we use the free space path loss (FSPL) expression given as,

$$FSPL = 20 \log(d) + 20 \log(f) + 20 \log\left(\frac{4\pi}{c}\right), \quad (5.2)$$

where, d is distance in meters and f is frequency in Hz. For a distance of 2 meters and a frequency of 300 GHz, the $FSPL$ is 88 dB. This dictates that either we provide high output power at the transmitter or use components that can overcome the path loss. For the first option, if we consider a margin of 15 dB (10 dB receiver NF and 5 dB for antenna misalignment), the minimum detectable signal at the receiver would

be -64.96 dBm and this suggests that the output power of the transmitter must be higher than $+23.04$ dBm (provided 88 dB FSPL), which is not a good choice because of its high value. Alternatively, a benefit of high frequency is a small effective aperture that can provide high gain. This is because antenna gain is inversely proportional to the wavelength [Balanis, 2016]. Hence, in our design we prefer using high gain antennas at the transmitter and the receiver along with state-of-the-art components that minimize receiver noise floor. A simplified block diagram of our design is shown in Fig. 5.1. In a typical transceiver, a transmitter and receiver are housed together.

5.2.1 Receiver

We now discuss the components selected in our receiver design based on the requirements discussed in the preceding sections.

Antenna

The significantly high loss at THz Frequency is fortunately overcome by the small sized high gain antennas [Johansson and Whyborn, 1992]. The antennas used in our design is a pyramidal horn antenna with gain of 25 dBi and horn length of 21.4 mm. Whereas at the lowest frequency (260 GHz) it is 2 dB lesser and at the highest frequency (400 GHz) it is 1.5 dB higher. The half power beamwidth is 12° . The gain of our antenna, from 260 to 400 GHz, is shown in Fig. 5.2.

Low Noise Amplifier

The receiver requires high sensitivity, which is typically influenced by its initial components and hence there should be minimal thermal losses at the input. Low noise amplifiers are designed to meet this criterion by matching the transistor's input for optimum NF while the output is conjugately matched to the system impedance (usually 50 ohms) for maximum gain and return loss. For a robust production, a low noise amplifier is usually designed using a transistor model and it is sufficient to use a linear model [Huang et al., 2006]. The noise input power is modeled by a resistor where

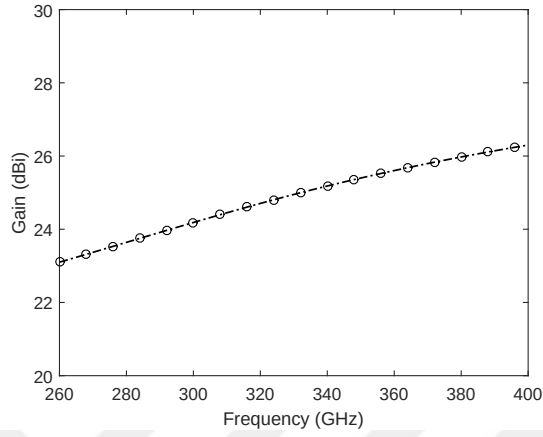


Figure 5.2: Antenna gain 260 GHz to 400 GHz range.

Table 5.1: Local oscillator single-sideband phase noise.

Frequency Range	10 Hz	100 Hz	1 kHz	10 kHz	100 kHz	1 MHz
20 to 40 GHz	-48 dBc	-72 dBc	-94 dBc	-100 dBc	-100 dBc	-124 dBc
40 to 67 GHz	-42 dBc	-66 dBc	-88 dBc	-94 dBc	-91 dBc	-118 dBc

its thermal agitation acts as noise [Bahl, 2009]. Transistor technology selection is a critical parameter and for low noise devices, pseudomorphic HEMT (pHEMT) is the most popular technology due to its low NF and high gain, however, at the expense of cost. mHEMT, on the other hand, combines performance of the pHEMTs and affordability of GaAs process. Therefore, the transceiver design is based on InAlAs/InGaAs based mHEMT technology fabricated by a 35 nm process. Initial design of the LNA is discussed in [Tessmann et al., 2008] and an improved version that provides small signal gain of 25 dB from 220 to 325 GHz with a NF of 7.2 dB is described in [Kallfass et al., 2015b].

Mixer

To process the information in the modulated signal, it is necessary to down convert the signal to a band that can be digitized and processed easily. Heterodyne receivers that employ a mixer which multiplies the two input signals, namely the RF and the local oscillator (LO), and outputs the sum or difference intermediate frequencies (IF) are commonly used in such cases. For communication in 300 GHz band, the required LO signal needs to be at very high frequency, however, due to the cost and difficulty of providing high frequency LO signal, subharmonic types of mixers are generally used that require N times smaller LO frequency (where $N \geq 2$).

Usually planar Schottky diodes are preferred as they exhibit better reproducibility, operation at room temperature and enable the integration of several diodes in balanced or anti-parallel configurations [Miao et al., 2014, Thomas et al., 2005]. Hence the mixer used in our system is WR2.8SHM by Virginia diodes, that operates in 260 to 400 GHz band with an LO input frequency of 130 to 220 GHz. The mixer exhibits double side band (DSB) equivalent noise temperature between 800 and 1500 Kelvin. We calculate the mixer NF using $NF = 10\log(1 + T_e/T_0)$. Where T_e is the equivalent noise temperature and T_0 is the room temperature (290 K). Considering 1000 K equivalent noise temperature in the required band, this comes out to be 6.48 dB. With a 3 to 6 dBm of LO power, typical DSB conversion loss is 9 dB and optimal IF input power is -10 dBm. A DSB type mixer is used without any input image rejection and thus both bands are equally propagated.

LO Signal

The LO signal of the mixer is the driver signal that has a switching impact. The drive power level has direct impact on the performance of the mixer. Since the selected mixer, WR2.8SHM is a subharmonic mixer, the required LO frequency is from 130 to 200 GHz with a power range of 3~6 dBm. To generate such a high frequency signal with considerable power, we designed a multiplier amplifier chain (Fig. 5.1) where required input signal frequency range is 25 to 35.5 GHz. This signal can be easily

generated using any available frequency synthesizer with low phase noise. For this purpose we used, MG3690C by Anritsu whose output power is up to +6 dBm and phase noise characteristics are given in the Table 5.1 [Anritsu, 2015]. The generated signal is then amplified using Ka band ERZ-HPA-2300-3700-27 power amplifier by ERZIA, which is shown as AMP1 in Fig. 5.1. The saturation power of the amplifier is 27 dBm and the gain is 21 dB. The output of the amplifier is fed to the frequency doubler, SFP-152KF-S1. Since this is a passive X2 frequency multiplier, the required input power is very high but can be provided by the amplifier preceding this multiplier.

The output signal frequency of the multiplier is 50 to 75 GHz provided by WR-15 waveguide with the output power of +5 dBm. Before further multiplication, this signal is amplified using QPW50662315 power amplifier by QuinStar Technology Inc, shown as AMP2 in Fig. 5.1. It operates from 50 to 66 GHz and provides a gain of 15 dB with 1 dB compression point at 23 dBm. The signal after amplification is fed to passive X3 frequency multiplier WR5.1x3HP by Virginia diodes Inc, that has 3% efficiency. The input signal power should be at least 18 dBm to compensate for the multiplier losses of approximately 15 dB and thus provide the LO signal to the mixer with the required minimal power. This output is fed to the LO port of WR2.8SHM mixer.

Driver Amplifier

The IF signal generated by the mixer is a down-converted signal that is amplified before digitization for better receiver gain. The amplifier used is DBLNA500013000A from DBWave Technologies, that operates from 0.01 to 30 GHz range with the NF of 4.5 dB. The gain of the amplifier is 35 dB with a 1 dB compression point at 28 dBm.

The signal generated at the amplifier's output is ultra-wideband and to capture the signal we used state-of-the-art real-time oscilloscope (RTO) DPO72304SX by Tektronics, that features sampling rate of 100 GSps with an analog bandwidth of 23 GHz.

5.2.2 Transmitter

The block diagram of transmitter is shown in Fig. 5.1. The LO signal generation for the transmitter is same as the receiver. RF path power amplifier is not used in the design for simplicity and because of the fact that the link is being designed for short distances. High speed signal generation is an intricate issue and it is resolved by using state-of-the-art M8195A 65 GSaps Arbitrary Waveform Generator from Keysight that provides analog signal bandwidth up to 25 GHz.

The saturation power of the WR2.8SHM is -10 dBm and hence the IF port power is limited to this value. Due to 9 dB mixer conversion loss, the maximum output power of the mixer is calculated as -19 dBm. To overcome high path loss, antennas used in our design are 25 dBi each at the transmitter and the receiver.

5.3 Results

To ensure first pass success of our design, we performed necessary system design simulations to evaluate the transmitter and the receiver. The lower limits for the effectiveness of a system are established by the noise. In principal, three main parameters that establish the performance of a wireless link are: dynamic range, sensitivity and selectivity. The first two are dictated by the receiver NF and linearity characteristics while selectivity is dictated by the phase noise contribution by the oscillator [Laskar et al., 2004]. The linearity performance of the receiver is determined by performing single-tone and two-tone analysis. To analyze selectivity, we perform single sideband phase noise measurements and discuss our results. Finally, we show the receiver gain and the NF of the receiver.

5.3.1 Single-Tone Analysis

In this analysis, the performance of the proposed transceiver is analyzed for a single-tone input signal. We apply a 2 GHz signal with -10 dBm power to the IF port of the WR2.8SHM mixer. This analysis measures the power signal spectrum at the output

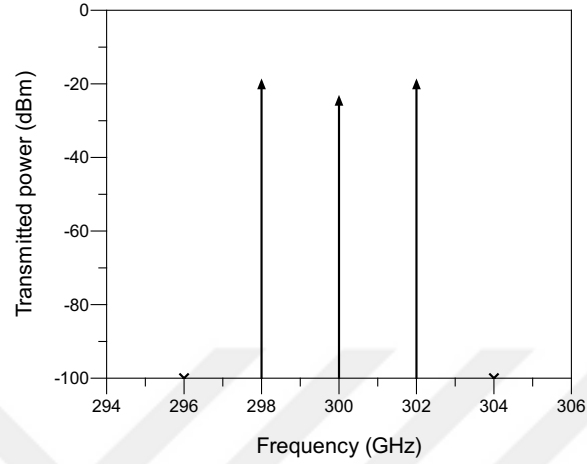


Figure 5.3: Output spectrum of the 300 GHz band transmitter.

of mixer. Due to the DSB nature of the mixer, the output spectrum contains the upper sideband signal of 302 GHz and lower side band signal of 298 GHz with output power of -19.114 dBm each, as shown in Fig. 5.3. The carrier frequency of 300 GHz can also be seen with output power of -23.58 dBm. This signal is passed through a channel model where at a distance of 2 m provides 38 dB of pathloss (considering antenna gain of 25 dBi each at Tx and Rx).

The transmitted signal is then fed to the receiver and is downconverted by the mixer after amplification. This baseband signal is amplified using a driver amplifier and the output spectrum is shown in Fig. 5.4. This signal is fed to the high sampling rate oscilloscope for signal processing. Several mixing products are observed in the baseband spectrum, however, the highest spurious signal is at least 60 dB below the required signal, which is much higher than the rule of thumb of 30 dBc and hence can be ignored.

5.3.2 Two-Tone Analysis

Non-linearity in the transfer function of receiver can be problematic for the communication system as the modulated signal is distorted causing difficulty in information

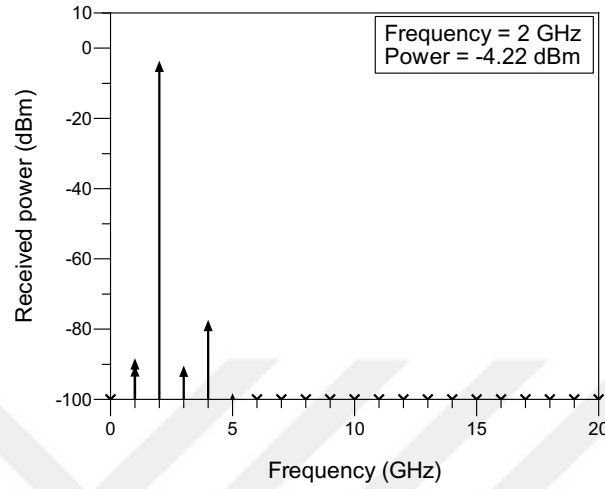


Figure 5.4: Baseband output spectrum of the receiver when excited by the DSB signal.

retrieval. One of the main contributors that cause non-linearity are the amplifiers because when the device enters its compression region, the non-linear effects start to become apparent [Khalid et al., 2015]. Two-tone analysis is a popular technique in which non-linearity of the circuit is determined by finding the carrier to intermodulation ratio (C/I) that is defined as the ratio of useful component output power to the intermodulation distortion (IMD) output power [Bahl, 2009]. In this technique, more than one carrier frequency with slight frequency spacing are applied to the circuit resulting in multiple sidebands due to the mixing that is introduced by the non-linear behavior of the device. Usually third order intermodulation products ($IM3$) appearing at $2f_1 - f_2$ and $2f_2 - f_1$ are the major source of distortion and are sufficient to predict the device performance.

$IM3$ of the proposed receiver is analyzed by applying input signals centered at 302 GHz with tone spacing (Δf) 2 MHz and calculating the ratio of intermodulation products and the carrier given by the following equation,

$$IM3(dBc) = 10 \log \left(\frac{P_{2f_2-f_1}}{P_{f_2}} \right) = 10 \log \left(\frac{P_{2f_1-f_2}}{P_{f_1}} \right), \quad (5.3)$$

The output spectrum of the receiver, when excited by two input signals of -32.5

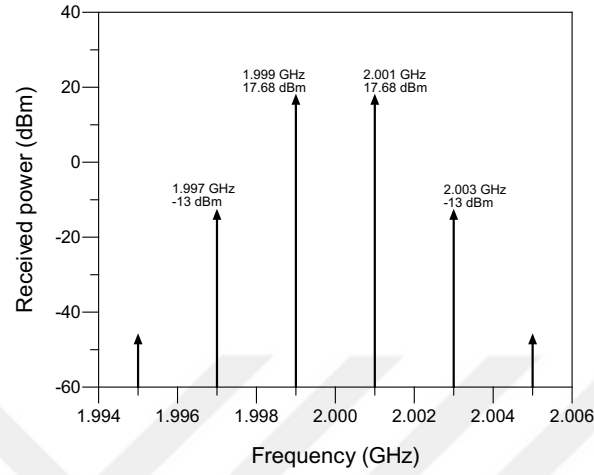


Figure 5.5: Baseband output spectrum of the receiver when excited by a two-tone input signal of -32.5 dBm.

dBm with tone spacing (Δf) 2 MHz, is shown in Fig. 5.5. The calculated C/I with these conditions is -30.69 dBc. This shows that our proposed transceiver design enters into non-linear region for an input power greater than -29.5 dBm that appears for inter-transmitter-receiver distances smaller than 8 cm.

5.3.3 Phase Noise Analysis

Phase noise is a key parameter in determining the minimum frequency spacing required in between adjacent channels. It is primarily generated by the signal generator, which is usually low for highly stable signal synthesizers, however, the amplifiers in the system might increase the phase noise leading to performance degradation. To get an insight of the selectivity of our receiver, we performed phase noise analysis. The results in Fig. 5.6 shows that the phase noise initially decreases with the slope of 30 dB/decade and after 100 KHz, it gets below -80 dBc. To illustrate this further, consider an example in which received power is -57 dBm at distance of 2 meters, while the noise floor of the receiver is -79.96 dBm. The signal to noise ratio for this particular example is calculated as 22.96 dB. On the other hand, it can be seen from

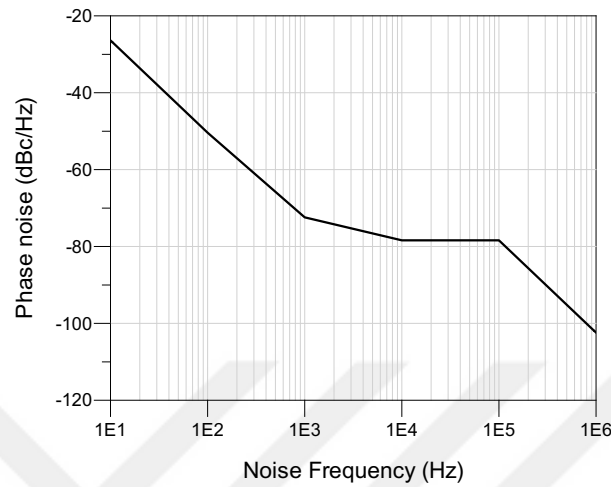


Figure 5.6: Phase noise of the receiver determined at the output of the driver amplifier.

the Fig. 5.6 that, at 10 Hz offset from the carrier, the phase noise is 27 dB below the carrier. Thus, for this particular case, the adjacent channel separation is suggested to be at least 10 Hz. However, for better compatibility, the recommended separation is 100 KHz. This separation will ensure that the interference in adjacent channels remain minimum.

5.3.4 Receiver Gain and Noise Figure

Another important parameter to determine is the overall receiver gain and NF. This helps in determining the minimum detectable signal by a receiver. Based on the gain, loss and compression of components in the receiver chain, ratio of output power to the input power is calculated to determine the receiver gain. The NF of the system is calculated based on the NF and gain of each component. It should be noted here that the NF is dominated by the first few components in the chain and once the desired signal is amplified well above the noise of the following components, their impact in increasing the signal noise is minimized. Results in Fig. 5.7 and Fig. 5.8 show that the receiver gain is higher than 51 dB whereas the NF is 9.56 dB, respectively. The

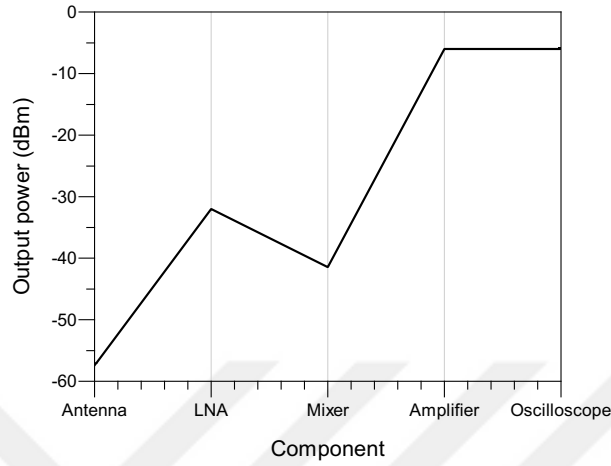


Figure 5.7: Output power of each component when the receiver is excited by -57 dBm signal power.

receiver gain results are shown for an inter-transmitter-receiver distance of 2 m. This suggests that the minimum detectable signal for our receiver is -70.4 dBm. Hence for the distance of 2 m, the receiver signal to noise ratio is 13.4 dB. Using Shannon's capacity formula, we determine the capacity of the proposed system as 90.31 Gbps. The link budget is summarized in Table 5.2. While the performance comparison of our system is shown in Table 5.3.

5.4 Conclusion

In this chapter, we propose a broadband transceiver design for 300 GHz band wireless communication based on discrete components that may be used in future IoT applications. We use low cost industry ready components that can be easily modified based on application specific requirements. The proposed transceiver provides a bandwidth of 20 GHz making it a good candidate for exploring wide band channel capabilities around the 300 GHz frequency band for 5G applications of IoT. Component parameters are used to simulate the system design to ensure first pass success. Linearity, noise and selectivity characteristics of the system are analyzed based on single-tone,

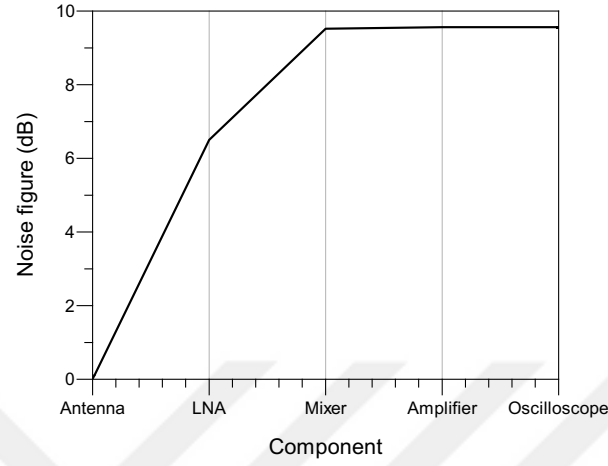


Figure 5.8: Noise figure of the receiver determined based on the noise and gain contribution of each component.

two-tone, receiver gain, noise figure, and phase noise analysis, which provides detailed insight of our proposed system. Finally, it is shown that our system is capable of providing 90.31 Gbps data throughput at 2 m. Future work includes analysis of digital data transmission on the proposed transceiver to determine practically achievable data rates.

Table 5.2: Link budget.

Parameter	Unit	Value
Start frequency	GHz	300
Stop frequency	GHz	320
Bandwidth	GHz	20
RF Tx Power	dBm	-19
Tx and Rx antenna gain	dBi	25
Link Distance	m	2
FSPL	dB	88
Rx LNA input power	dBm	-57
System temperature	K	290
Mixer noise temperature	K	1000
Rx noise figure	dB	9.56
Receiver gain	dB	51
Minimum detectable signal	dBm	-70.4
IM3 of the receiver	dBm	-29.5
SNR @ 2 m	dB	13.4
Max Range @ SNR = 10 dB	m	2.9

Table 5.3: Performance comparison of the proposed system.

Parameter	This work	[Carpenter et al., 2016]	[Katayama et al., 2016]	[Kallfass et al., 2015b]
Bandwidth SSB (GHz)	20	17	4.7	32
Output power (dBm)	-19	9	-14.5	-4
Antenna gain (dBi)	25	40	-	24.2
Receiver noise figure (dB)	9.56	9.5	-	6.7
Link distance (m)	2.9	1.8	-	2.4

Chapter 6

EXPERIMENTAL THROUGHPUT ANALYSIS OF A MIMO COMMUNICATION CHANNEL

This chapter presents first results of a Terahertz (THz) Band line-of-sight (LOS) 2×2 multiple-input multiple-output (MIMO) channel. The system architecture is based on a subharmonic mixer that translates the measuring frequency of vector network analyzer (VNA) in the range of 298 - 313 GHz. The system uses virtual antenna array technique to characterize a MIMO channel. The antenna element spacing is derived from the principles of diffraction limited optics to establish parallel channels for higher system throughput and reliability. The MIMO channel measurements are then used in simulations to evaluate performance of the communication system operating in THz Band MIMO communication channel. Finally, we have shown the MIMO link in THz Band operating at 7 Gbps, with higher reliability, in comparison with the single channel operating at 5.55 Gbps.

6.1 Introduction

Terahertz Band (0.3 - 10 THz) communication is attracting more interest due to its large available bandwidth, small antenna size and high directivity. The large bandwidth is fundamental in enabling very high speed communication that is envisioned in future generation of wireless communication. Since the demand for higher data rates is ever increasing, MIMO techniques should be incorporated in the THz Band communication. This technique has the potential to not only increase the data throughput but also improve reliability of the systems. MIMO communication links provide higher data throughput without increasing the system bandwidth. Multiple closely spaced transmitters (TX) and receivers (RX) are used to establish parallel commu-

nication channels that can simultaneously enhance the throughput and reliability of the system.

MIMO in conventional low frequency communication systems (2 - 5 GHz) exploits multipath signals in a non-line-of-sight (NLOS) channel to achieve spatial diversity. However, in THz Band communication, LOS MIMO is more suitable due to small antenna size and high directivity. Although high directivity is more prone to link breakage caused by path obstructions, a large array of MIMO links can easily mitigate this shortcoming.

To characterize THz Band radio channel, several channel sounders and measurements have been reported. A 300 GHz measurement system is described using single transmitter receiver with 9 GHz bandwidth in [Jastrow et al., 2008b]. In [Priebe et al., 2010] and [Priebe et al., 2011], channel response of single-input single-output (SISO) at 300 GHz is shown using subharmonic schottky diode mixer and a VNA. LOS and NLOS measurements at 300 GHz are performed with a system bandwidth of 20 GHz in [Khalid and Akan, 2016b]. However, to date, no MIMO link operating in THz Band has been investigated.

To evaluate a MIMO channel operating in THz Band, we have emulated a 2×2 MIMO system by linearly displacing a single transmit and receive antenna to form a virtual antenna array as in [Gustafson et al., 2014]. Separation between the transmitter and receiver are calculated using the principles of diffraction limited optics, [Torkildson et al., 2006]. The measured MIMO channel matrix is then used in a communication system simulation to evaluate the channel performance. Results confirm that THz Band LOS MIMO link can be used to transmit 7 Gbps of data. The higher path loss and antenna directivity can enhance frequency reuse and multiply capacity of the THz Band system when deployed in a femtocell regime.

The remainder of this chapter is organized as follows. In Section 6.2, we describe our THz Band MIMO measurement system and channel response. In Section 6.3, we show the performance of our emulated MIMO link. Finally, we conclude the chapter in Section 6.4.

6.2 THz Band MIMO measurement system

6.2.1 Channel Measurements Testbed Specifications

The setup used for MIMO channel measurements consists of two major parts, Anritsu Vector Network Analyzer (VNA) MS4647B and VDi WR2.8MixAMC modules. VNA MS4647B is a wideband equipment with the upper frequency limit of 70 GHz. To extend the frequency of the system to 300 GHz, WR2.8MixAMC are attached with the VNA. These extension modules are based on subharmonic mixers that upconvert and downconvert the RF signal of the VNA. Detailed description of the setup is given in [Khalid and Akan, 2016b]. The block diagram of our setup is shown in Fig. 6.1.

Yttrium iron garnet (YIG) based synthesizer was used to generate a common local oscillator signal f_{LO} of 12.357 GHz for both the upconverter and downconverter modules to achieve phase coherence. WR2.8MixAMC module multiplies this f_{LO} signal 24 times to generate THz Band signal, which is fed to f_{LOTHz} port of subharmonic mixer. Intermediate frequency f_{IF} port of this mixer is attached to the VNA that generates a sweep signal from 1 - 15 GHz. After mixing, the signal generated at the output of the subharmonic mixer contains frequencies from 298 - 313 GHz ($f_{LOTHz} + f_{IF}$). This signal undergoes attenuation and phase alteration as it passes through the wireless channel. The signal is received at the receiver and fed back to the VNA after down conversion. The VNA determines changes in the received signal based on the information of transmitted signal and calibration data to show scattering parameters (s-parameters) of the channel. For measuring channel response, through/reciprocal calibration is performed with direct interconnection of the module's waveguide [Khalid and Akan, 2016b, Priebe et al., 2011]. All the later measurements are recorded with horn antenna attached at the output and input port of the transmitter and receiver, respectively. Antenna gain is 25 dBi each, with 3 dB beamwidth of 12° .

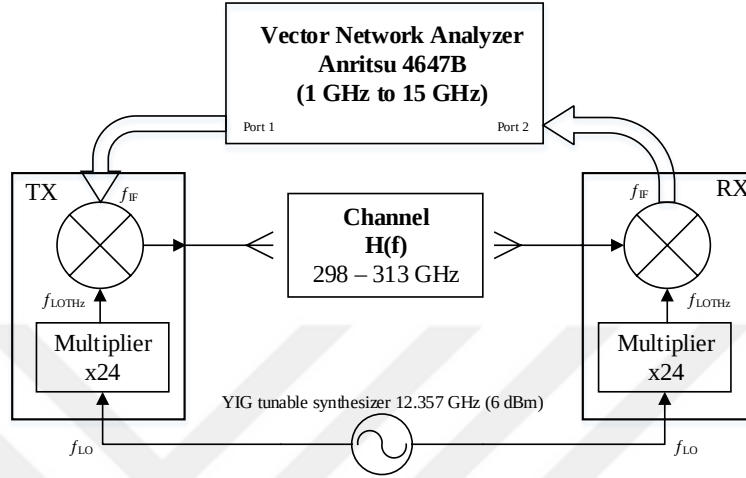


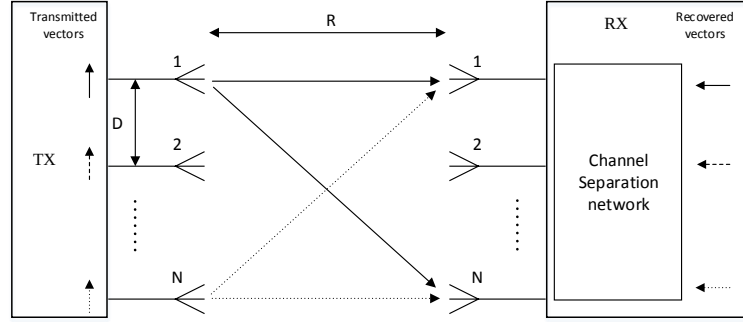
Figure 6.1: Block diagram of THz Band measurement setup.

6.2.2 MIMO Measurements Based on Diffraction Limited Optics

Unlike conventional low frequency MIMO that depends on NLOS multipath signals, THz Band MIMO is more appropriate with LOS spatial diversity scheme. This is because in THz Band, the high pathloss causes the multipath signals to get very attenuated. On the other hand, the peculiar behavior of small antenna size makes LOS MIMO more promising technique as large number of antenna elements can be placed alongside a small area to make multiple parallel channels. The high directivity will make sure that the inter channel interference, generated due to closely placed transmitters, is minimal. The concept of THz Band $N \times N$ LOS MIMO is shown in Fig. 6.2.

In THz Band LOS MIMO, antenna separation D is a very important parameter ensuring multiple transmitters can be distinguished at the receiver. This problem is addressed by using diffraction limited optics theory [Torkildson et al., 2006, Sheldon et al., 2008], which states that for LOS MIMO, the antenna element spacing should be kept as

$$D = \sqrt{\frac{R \cdot \lambda}{N}}, \quad (6.1)$$

Figure 6.2: THz Band $N \times N$ LOS MIMO system.

where R is the distance between transmitter and receiver, λ is the free space wavelength of the carrier frequency and N is the number of elements at the receiver. This condition is very similar to Rayleigh criterion. It is evident that there will be cross channel interference among different channels, this interference can degrade the performance and must be reduced using a channel separation network, as shown in Fig. 6.2.

Measurements were conducted in NWCL lab, which is a typical single floor office environment. The setup for 2×2 LOS MIMO channel experimentation was placed on a solid aluminum optical breadboard with matte anodized finish that reduces unwanted reflection. Distance between the transmitter and the receiver was kept as $R = 25$ cm. For such small distance and narrow antenna beamwidth along with high pathloss, reflected NLOS signals can be neglected. At a center frequency of 305.5 GHz, the corresponding inter element spacing calculated using (6.1) is $D = 1.2$ cm. On the other hand, VNA was configured to record full 14 GHz band measurements with 1 KHz intermediate frequency bandwidth (IFBW). Details of the measurement parameters are given in Table 6.1 and our measurement setup is shown in Fig. 6.3.

To acquire a full 2×2 LOS MIMO channel matrix, we first recorded the response of MIMO TX 1 in both the channels. This was carried out by placing transmitter module at antenna location 1 of the TX and the receiver module at antenna location 1 (RA1) of RX. The receiver module was then displaced to antenna location 2 (RA2)

Table 6.1: Measurement parameters of VNA

Parameter	Symbol	value
<i>Measurement points</i>	N	7368
<i>Test signal power</i>	P_{in}	-10 dBm
<i>Start frequency</i>	f_{start}	1 GHz
<i>Stop frequency</i>	f_{stop}	15 GHz
<i>Bandwidth</i>	B.W.	14 GHz
<i>IFBW</i>	Δf_{IF}	1 KHz

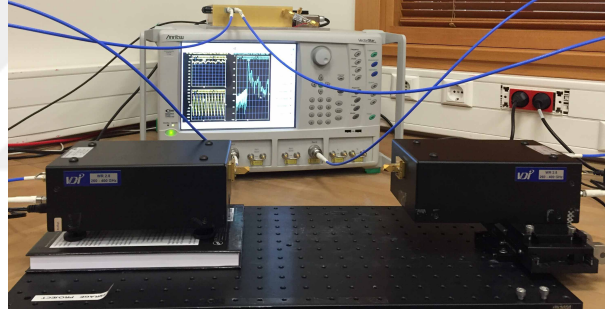


Figure 6.3: THz Band MIMO channel measurement setup.

of RX. In our setup, the desired phase response was obtained at 1.257 cm away from the previous position, which agrees with our theoretical estimate based on (6.1). The first configuration measures the response of signal transmission in channel 1 (CH1) and the second configuration measures cross channel interference in channel 2 (CH2) due to MIMO TX 1. The downconverted magnitude response is shown in Fig. 6.4 and phase response in Fig. 6.5. It can be seen in Fig. 6.5 that the phase difference of the received signal at adjacent antennas is 180° . Similarly, frequency response of MIMO TX 2 was recorded at both RX antenna locations to create a virtual array. The recorded measurements were combined to generate a full 2×2 channel matrix. This matrix was used to perform post processing on a computer to evaluate the performance of THz Band 2×2 MIMO channel.

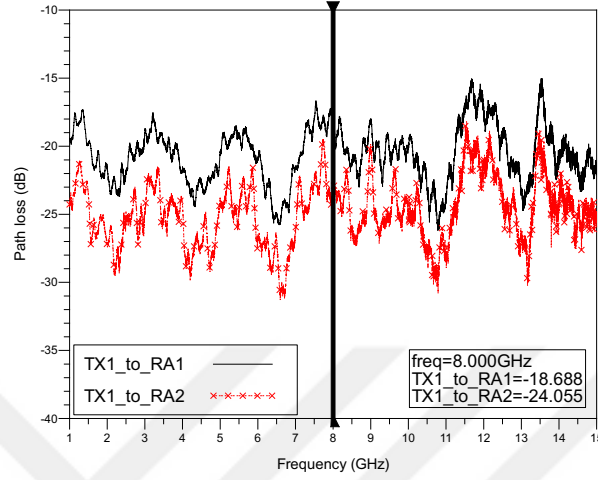


Figure 6.4: Signal magnitude of TX 1 at RA 1 and cross channel interference magnitude at RA 2.

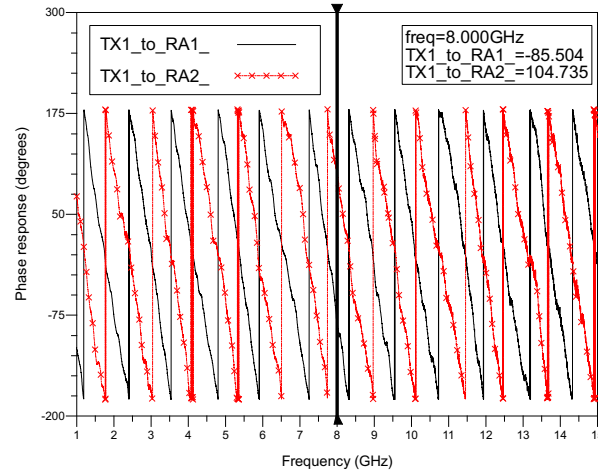
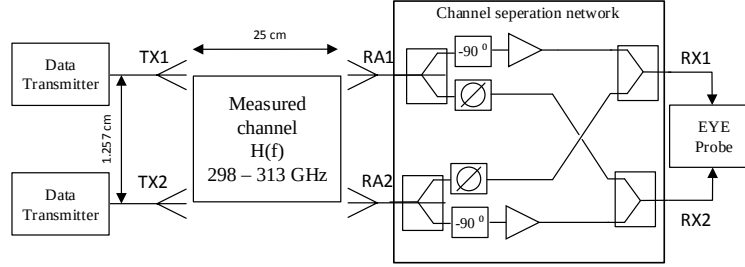


Figure 6.5: Signal phase of TX 1 at RA 1 and cross channel interference phase at RA 2.

The cross channel signal at any receiver is out of phase and can be suppressed using a channel separation network. This network mainly consists of delay elements, amplifiers, and splitter/combiner. The architecture of channel separation network is shown in Fig. 6.6. The optimal values of the amplifier and phase delay element

Figure 6.6: Schematic of 2×2 THz Band MIMO system.

were iteratively determined while minimizing cross channel interference. In CH1, the corresponding optimal value for amplifier gain was found as 7.24 dB and the phase delay was 85.19° whereas, in CH2, the amplifier gain was 7.69 and the phase delay was 84.58° . The slight difference in these component parameters is due to the fact that the response for both channels was not exactly identical in an office environment. Signal magnitude after channel separation network is shown in Fig. 6.7. It can be seen that the cross channel interference is 13.9 dB below the main signal magnitude and is much higher than the noise level, which is -40 dB. Hence the throughput will be determined by the signal to interference ratio. However, for distances greater than 50 cm, the noise will dominate and hence the throughput will be chiefly determined by the signal to noise ratio (SNR).

6.3 Results of 2×2 LOS THz Band MIMO

The THz Band MIMO channel was evaluated offline for communications by transmitting and receiving bits over it, in the baseband. The setup consists of a transmitter that generates pseudorandom binary sequences, the measured MIMO channel response, a channel separation network and an equalizer. The system was analyzed for different configurations such as single transmit mode, MIMO CH1 and MIMO CH2. While operating in single transmit mode, other channel was kept silent, whereas, in MIMO channel configuration, other channel was excited by uncorrelated bits to imi-

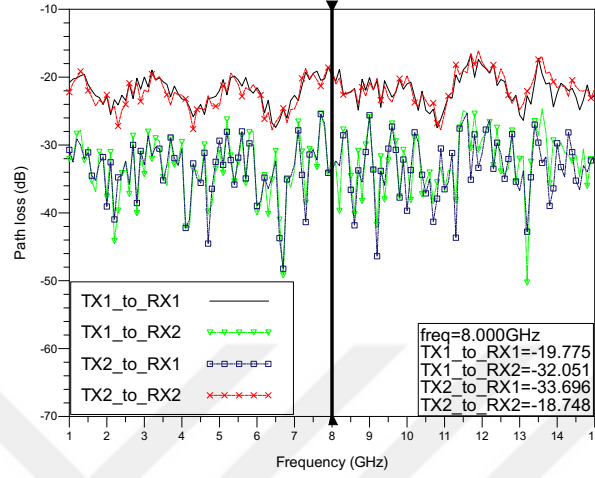


Figure 6.7: Path loss of CH1 (TX1.to_RX1) and CH2 (TX2.to_RX2) along with cross channel interference magnitude at receiver 1 (TX2.to_RX1) and receiver 2 (TX1.to_RX2), after channel separation network.

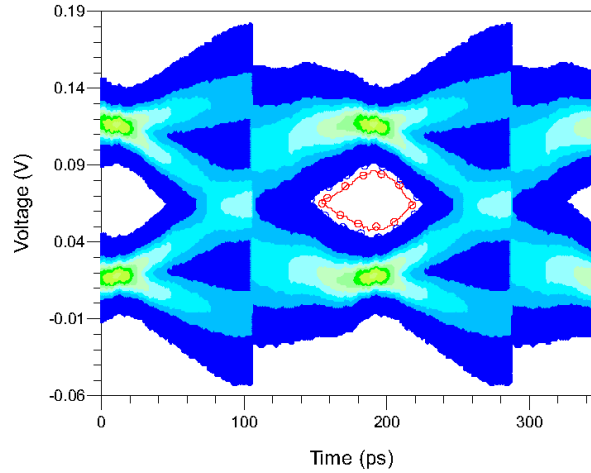


Figure 6.8: Eye patterns at 5.55 Gbps without cross talk.

tate cross talk scenario. Output of the equalizer was used to generate eye diagram, bit error rate (BER) contours and Bathtub using statistical eye methodology, [Sanders et al., 2004]. All configurations are compared based on same eye height.

In single transmit mode, the channel was excited with data rate of 5.55 Gbps.

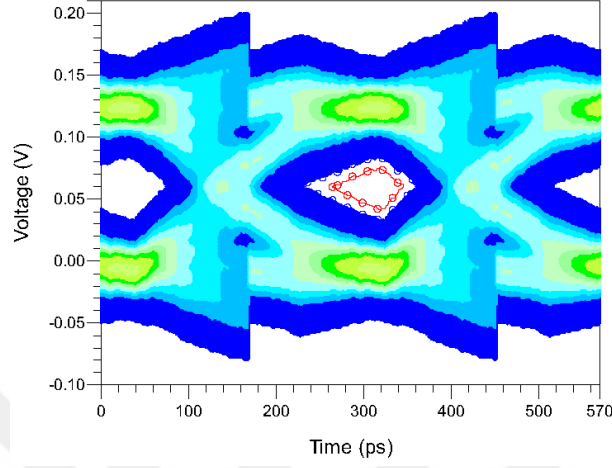


Figure 6.9: Eye patterns at 3.5 Gbps on CH1 with cross talk.

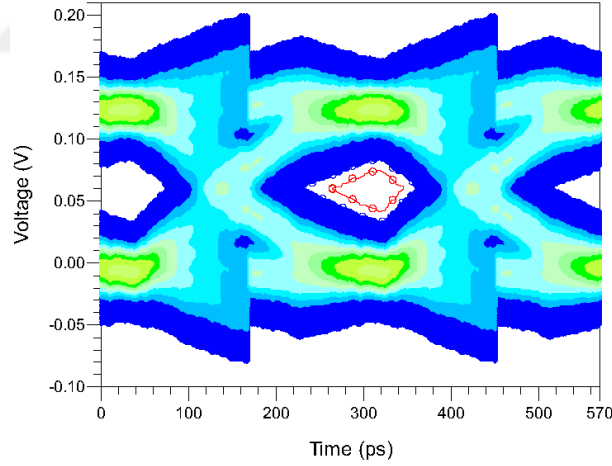


Figure 6.10: Eye patterns at 3.5 Gbps on CH2 with cross talk.

The corresponding eye diagram with BER contour, for the value of 10^{-12} , is shown in Fig. 6.8. Height of the eye is 0.052 volts and height of BER contour is 0.038 volts. One can calculate the capacity for a flat channel using Shannons capacity formula as 148.8 Gbps, with an SNR of 32 dB and bandwidth of 14 GHz. However, in our case, neither the channel is flat nor the phase response is perfectly linear thus causing the significant decrease in the throughput.

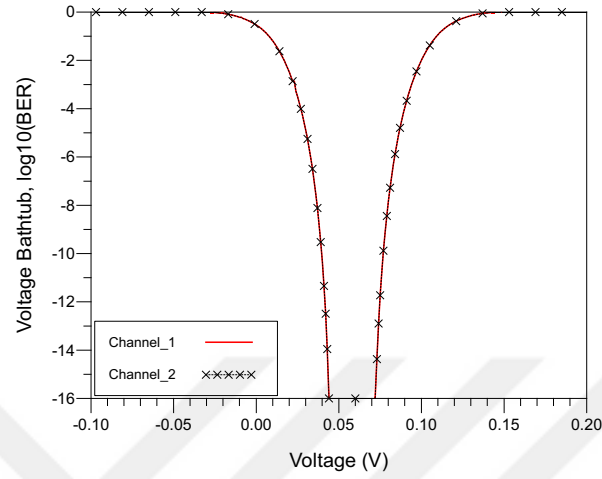


Figure 6.11: Voltage Bathtub for both channels operating at 3.5 Gbps.

In the MIMO channel mode, eye height of 0.052 volts for individual channel was achieved at 3.5 Gbps with 10^{-12} BER contour height of 0.033 volts. Ideally, the data rate must remain the same and overall system throughput should have doubled, however, due to addition of cross channel interference, the data rate was decreased. However, the combined system throughput of 7 Gbps is still higher than the single transmit mode. The corresponding eye diagram and BER contour is shown in Fig. 6.9 and Fig. 6.10, for CH1 and CH2, respectively. The BER performance of the MIMO link from 10^0 to 10^{-16} , for both channels, is depicted using voltage bathtub, in Fig. 6.11. The curve tells maximum tolerable noise for a given BER and shows how an increase in noise can decrease the BER performance of our system, for instance, if signal is sampled exactly at 320 ps, the maximum tolerable noise, depicted by the tub width at BER 10^{-16} , would be 0.028 Volts. Whereas, a noise of 0.033 volts will reduce the system performance to a BER of 10^{-12} . The Performance parameters are summarized in Table 6.2.

Table 6.2: Performance of THz Band 2x2 LOS MIMO system

		Single Channel	MIMO Channel	
Parameters	BER		CH1	CH2
Height (V)	10^{-12}	0.038	0.033	0.033
	10^{-6}	0.050	0.051	0.051
	-	0.052	0.052	0.052
Width (ps)	10^{-12}	64.81	82.8	82.86
	10^{-6}	83.71	132.9	132.9
	-	100.8	155.7	155.7

6.4 Conclusion

We have presented THz Band 2×2 LOS MIMO system operating at 7 Gbps. Spatial diversity scheme is utilized with antenna element spacing derived from the principles of diffraction limited optics. The channel transmission and inter channel interference is measured and recorded using virtual antenna array technique and channel sounder based on VNA and frequency extenders. We have shown that cross channel link vectors are out of phase and can be suppressed using channel separation network. Optimal channel separation parameters were determined for minimal cross channel interference. Finally, THz Band 2×2 MIMO link was evaluated for digital communication system and confirmed to achieve 7 Gbps. However, a significant difference was seen between the theoretical and experimental throughputs that opens up the need for further research. Moreover, our results suggest that single carrier might not provide the ultimate benefit of such a wireless communication channel, instead multi-carrier modulation will perform better in terms of spectral efficiency, given its relative immunity to fading and interference. Future works in this domain are to analyze systems with larger one-dimensional and two-dimensional arrays transmitting real time data in different indoor environments. Performance of multi-carrier modulation system may also be analyzed.

Chapter 7

CAPACITY AND COVERAGE ANALYSIS OF FD-MIMO BASED CHANNEL

Current and proposed Internet of things (IoT) applications are expected to bring about a major technological revolutions. Next-generation wireless communications in such devices are expected to support high speed data transfers. Among different candidate technologies, terahertz (THz) band communication seems to be a promising direction due to availability of high bandwidth in the electromagnetic spectrum around this frequency range and its directional nature governed by the directive antennas. In this chapter, we look into some networking scenarios in full-dimension multiple-input multiple-output (FD-MIMO) based THz Band indoor wireless networks to determine the number of nodes that can be connected to a base station as a function of the antenna characteristics. Furthermore, we analyze the performance of the users and network based on their ergodic capacity. Our results suggest fundamental parameters that can be used in future THz Band analysis and implementations.

7.1 *Introduction*

The inter-networking of devices, buildings, vehicles and other items such as wearable electronics that comprise technologies labeled under the banner of Internet of things (IoT), are quickly molding the future of technology in the current era. Cisco envisions this global phenomenon to reach a market value of more than 14 Trillion US Dollar by the year 2022 [Bradley et al., 2013]. On the other hand, increasing data rates for communication are in high demand to cater for soaring data consumption in new applications such as high-resolution videos and 3D experience environments.

In order for the technology to compete with consumer needs, 5G wireless networks

need to realize data rates reaching terabits per second (Tbps) [Huang and Wang, 2011, Daniels and Heath, 2007]. The 60 GHz spectrum is expected to increase data rates up to 6 Gbits/s [IEEE, 2009] but the available spectrum around this frequency is limited to 2 GHz that may not completely satisfy the increasing demands of higher data rates. To support such high throughputs of Tbps, enormous bandwidth is required, and hence, researchers around the world are exploring the Terahertz (THz) Band (0.3 - 10 THz). Higher frequency networks can support higher data rates, however, being highly directional, they are more easily blocked by any obstacles on their path providing a set of unique challenges for THz Band communications. Additionally, IoT framework necessitate extensive and seamless communication between a large number of devices in indoor environment. Therefore, the main motivation for the current study is to provide analytical analysis for networking in THz Band communication systems such that they may be realized as an enabling technology for 5G indoor IoT applications.

An efficient approach to increase system capacity using THz Band is to pack these directional antennas intelligently to form FD-MIMO such that the resulting array provides dedicated coverage to all the devices in its vicinity [Kim et al., 2014]. Several articles are available that analyze the performance of mmwave networks using directional antennas [Thornburg et al., 2015]. However, very limited literature is available that analyze the performance of FD-MIMO based THz Band network.

In this chapter, our aim is to improve spatial multiplexing performance of THz Band networks by introducing a novel concept of spatial separation of antenna beams that is in addition to its half-power beam width (HPBW). We denote this separation parameter as guard width. To evaluate this concept in an FD-MIMO based THz Band network, we try to determine its effect on the user and network performance. We provide optimum guard width value for the scenario where user capacity is to be maximized and in the scenario where the network capacity is to be maximized. Expressions are derived for robust selection of guard width values and interpretation of the system performance. We also analyze the minimum cell separation required based

on lowest possible inter-cell interference. An insight to the area spectral efficiency (ASE) of THz Band network is also provided to appreciate its use in the future indoor IoT applications.

The remainder of this chapter is organized as follows. In Section 7.2, we describe the system model used for simulation and performance analysis of THz Band network. Coverage and capacity modeling is provided in Section 7.3. Section 7.4 discusses the analytical results. Finally, we conclude the chapter in Section 7.5.

7.2 System Model

Consider an indoor IoT network where devices achieve high speed bi-directional communication links by utilizing the capabilities of a THz Band channel and the base stations are equipped with FD-MIMO technology. Due to high path loss, the coverage area of THz Band networks is small, however, this peculiar behavior comes with an inherent advantage of smaller cell sizes that permit frequent channel reuse and more secure connections. To analyze such a network, we consider a scenario where the IoT devices are placed on the Euclidean plane, $\mathbb{R}^2$ with an average distance of r meters from the base station, as

$$r = \sum_{i=1}^N \frac{r_i}{N}, \quad (7.1)$$

where, r_i is the distance of i^{th} device and N is total number of devices connected to the base station.

7.2.1 FD-MIMO

A typical FD-MIMO system consists of an antenna array in 2-dimensional space that provides flexibility of beam-forming in almost any arbitrary direction. At lower frequency bands, the sizes of such antenna arrays become huge making their use in indoor IoT devices impractical. As an example, consider an antenna of 64 elements operating at 2.5 GHz in an 8×8 array with 0.5 wavelength (λ) separation. This would require a nominal area of 50 cm $\times$ 50 cm for the antenna array [Kim et al., 2014],

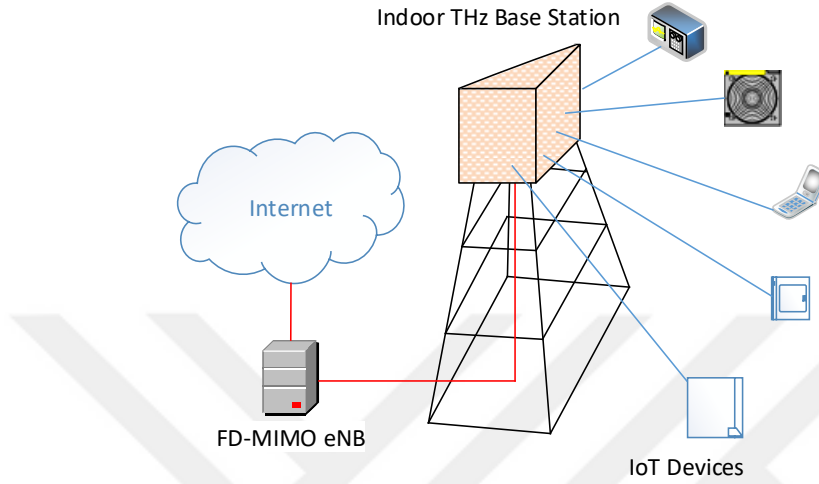


Figure 7.1: THz Band FD-MIMO system.

which clearly is an infeasible solution for indoor IoT networks. On the other hand, at higher frequency, such as a THz Band system operating at 300 GHz, a similar array would require only $0.4 \text{ cm} \times 0.4 \text{ cm}$ area. This is 20 times smaller than its lower frequency counterpart. Thus, THz Band is especially suited to FD-MIMO in indoor IoT networks.

By moving its beam, a typical planar antenna array can provide a coverage of 120° and multiple arrays can be placed in different orientations to increase the coverage [Nam et al., 2013]. Usually three arrays are placed in a triangular geometry to provide full 360° of coverage in a cell, as shown in Fig. 7.1. We denote the angular coverage of an antenna as antenna movement freedom (φ) and use it further in our analysis.

7.2.2 Antenna radiation pattern

Antenna radiation patterns play a very important role in determining the coverage of a system. In our analysis, antenna radiation pattern is approximated with a modified *sinc* function. Such an antenna is characterized by the following three parameters; Half power beam width β_{3dB} , main lobe gain G_t and damping factor ζ . The damping factor controls the sharpness of nulls and improve the realization of a *sinc* function

with an actual high directional antenna radiation pattern. Its value determines the number of neighboring points up to ζ degrees on either side of the specified $f(\theta)$.

The radiation pattern is modeled by the following expression,

$$f(\theta) = \frac{1}{2\zeta + 1} \sum_{m=-\zeta}^{\zeta} \text{sinc}^2(A\theta + m), \quad (7.2)$$

where, the value of A determines the lobe width dependent on β_{3dB} and can be determined considering θ_{3dB} , an angle where the function goes to the half value of its maximum,

$$\text{sinc}^2(A\theta_{3dB}) = 0.5, \quad (7.3)$$

$$\beta_{3dB} = 2\theta_{3dB} = \frac{0.8854}{A}, \quad (7.4)$$

hence,

$$A = \frac{0.8854}{\beta_{3dB}}. \quad (7.5)$$

The gain of the antenna at an arbitrary angle θ can be determined using,

$$G(\theta) = \frac{1}{2\zeta + 1} \sum_{m=-\zeta}^{\zeta} G_t \text{sinc}^2\left(\frac{0.8854}{\beta_{3dB}} \theta + m\right). \quad (7.6)$$

The antenna radiation pattern using ideal *sinc* function is shown in Fig. 7.2. It can be seen from the figure that the pattern returns to null which does not represent a practical radiation pattern. A typical antenna radiation pattern is similar to the one shown in Fig. 7.3 [Balanis, 2016], which is used in our further analytical analysis.

As discussed earlier, thanks to the small antenna size at high frequency, many antennas can be packed to realize FD-MIMO and thus, multiple lobes can be spread around spatially to create several simultaneous communication links. However, very low inter-lobe spacing of antennas can cause severe interference among neighboring devices and downgrade the performance of spatial multiplexing. Using techniques derived from the principles of diffraction limited optics, similar to [Khalid and Akan,

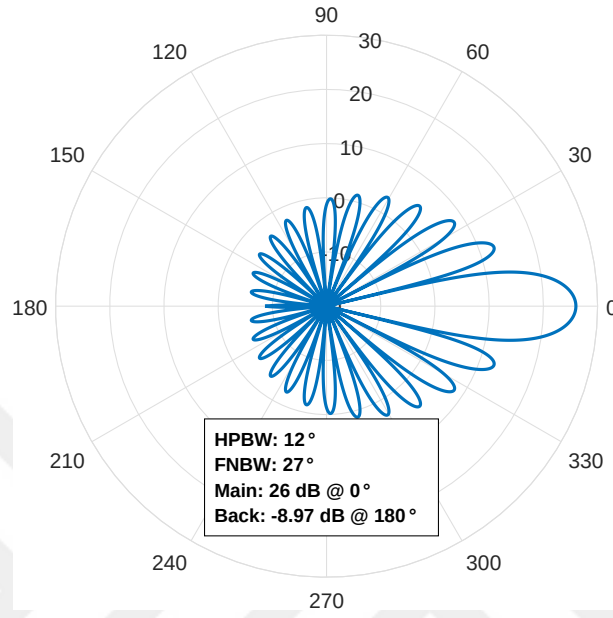


Figure 7.2: Radiation pattern of the antenna based on ideal *sinc* function.

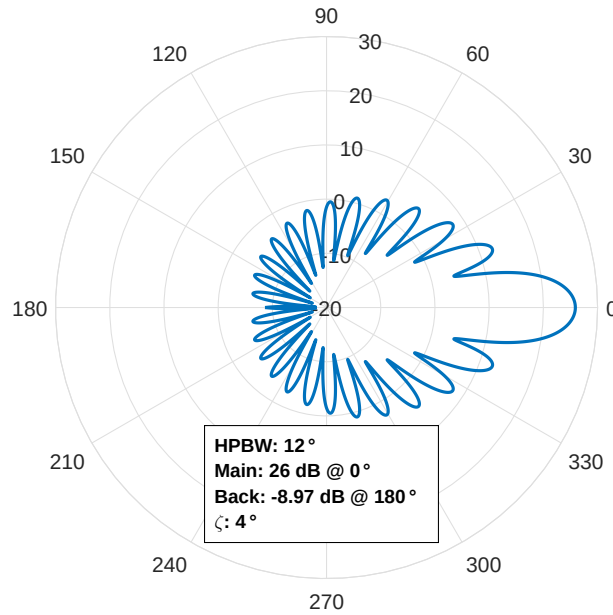


Figure 7.3: typical radiation pattern of the antenna with modified *sinc* function.

2016a], one can significantly reduce the inter-lobe spacing, however, at the increased cost of system complexity. To avoid complexity, in this chapter we propose a simpler

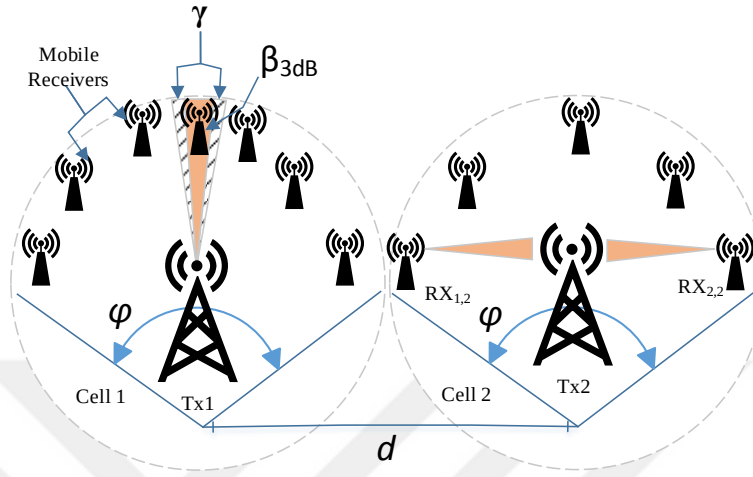


Figure 7.4: THz Band FD-MIMO architecture.

system based on using the minimum gap between each lobe denoted as guard width (γ). We define γ as the angular distance between the β_{3dB} of adjacent lobes as shown in Fig. 7.4 and its optimal value is based on user requirements.

For a given φ , β_{3dB} and γ , number of possible connections (N) can be calculated using,

$$N = \left\lfloor \frac{\varphi}{\beta_{3dB} + \gamma} \right\rfloor, \quad (7.7)$$

where, $\lfloor \cdot \rfloor$ represents floor value of the expression.

7.2.3 Interference and Obstacles

Due to high directivity, THz Band networks are only LOS interference limited, which is why we include a guard width concept to enhance the node and network capacities.

Consider a scenario where the K^{th} user/receiver (K_r) is connected to the K^{th} transmitting beam (K_t). We assume that the receiver and transmitter beam are perfectly aligned to achieve maximum gain. This can be achieved by applying beam alignment technique on beam steering, which is essential to an FD-MIMO antenna array. Interference faced by the user K_r is due to the power transmitted by other lobes in the direction of this receiver. If neighboring beams are very close to a receiver,

higher interference is faced by that node. Conversely, if the angular distance between each lobe is increased, it can be observed by looking at Fig. 7.3 that the interference level will decrease, owing to the fact that transmission level gets weaker moving away from the main lobe at the expense of fewer devices per coverage area. The interference observed by a single user K_r can be determined as,

$$I_{K_r} = \sum_{i=1, i \neq K}^N G_{K_r} G_i(\theta_i) P_{t_i} \left(\frac{c}{4\pi f r} \right)^n, \quad (7.8)$$

where, $G_i(\theta_i)$ is the gain of transmitter i at an angle θ_i found using (7.6), G_{K_r} is the gain of user K_r , c is the speed of light, f is the frequency of transmission, r is the distance between the transmitter and the receiver and n is the path loss exponent. It should be noted here that the receiving antenna and the transmitting antenna array are assumed to be points on the Euclidian plane, $\mathbb{R}^2$ and the distance r is considered equal for all the transmitters. Also, G_{K_r} is considered to have maximum gain, irrespective of the angle θ_i .

It is evident from (7.7) that by decreasing γ , the number of connected user are increased, however, by (7.8) we can also see that this results in increased interference. The combination of both these behaviors directly influence the single-user and network-level capacities.

Due to high material absorption in the THz Band, obstructions have significant impact on the network performance. It is experimentally shown that the obstructions expected in THz Band can absorb transmitted signals in the range of 10-30 dB [Khalid and Akan, 2016b]. Considering that the obstruction between i^{th} transmitter and K^{th} receiver is denoted as $E(i, K)$, the resulting interference with obstruction can be written as,

$$I_{K_r} = \sum_{i=1, i \neq K}^N G_{K_r} G_i(\theta_i) P_{t_i} E(i, K) \left(\frac{c}{4\pi f r} \right)^n. \quad (7.9)$$

In this work, we consider the placement of obstructions as deterministic and leave the stochastic geometry analysis for future work.

7.3 Coverage and Capacity Model

In our system model, we consider that the base station is placed at the origin and that a node K_r receives signal power which is determined by a simple variant of Friss path loss equation, as,

$$S_{K_r} = G_{K_r} G_{K_t} P_{t_K} E(i, K) \left(\frac{c}{4\pi f r} \right)^n, \quad (7.10)$$

where, G_{K_t} and P_{t_K} is the gain and transmitted power of the K^{th} transmitter communicating with the K^{th} receiver, respectively. The user K has coverage if its Signal-to-interference-plus-noise-ratio (SINR) is greater than a threshold (T) that is required for successful packet transmission. The SINR for the stated system can be calculated as,

$$SINR = \frac{G_{K_r} G_{K_t} P_{t_K} E(i, K) \left(\frac{c}{4\pi f r} \right)^n}{N_0 + \sum_{i=1, i \neq K}^N G_{K_r} G_i(\theta_i) P_{t_i} \left(\frac{c}{4\pi f r} \right)^n}. \quad (7.11)$$

The threshold T is usually dependent on the modulation scheme being used in a system. The ergodic capacity of the system is given as [Shannon, 1948],

$$C_{K_r} = B_c \log_2 (1 + SINR), \quad (7.12)$$

where, B_c is the coherence bandwidth of the system. The parameter B_c is very important for wideband systems as it determines the range of frequencies for which the channel can be considered flat. It is affected by the multipath propagating signals that are usually expected in indoor channels.

Another important parameter, network capacity, in $\mathbb{R}^2$ can be calculated as,

$$NC_{\mathbb{R}^2} = N \cdot C_{K_r}, \quad (7.13)$$

which determines the performance of the whole network. Moreover, the full FD-MIMO in $\mathbb{R}^3$ can be determined for the antennas with identical radiation pattern in azimuth and elevation directions as,

$$NC_{\mathbb{R}^3} = N_\theta N_\phi \cdot C_{K_r}, \quad (7.14)$$

where, N_θ is evaluated by considering the $0 \leq \varphi_\theta \leq 360$ and β_{3dB_θ} for the azimuth plane of antenna array, while N_ϕ is evaluated by considering the $0 \leq \varphi_\phi \leq 180$ and β_{3dB_ϕ} for the elevation plane of the antenna array.

In this work, we have considered that each user is equally demanding the same capacity and thus, the parameter γ is identical for all users. For the case where each user demands different quality of service (QoS), this problem can be easily modified to assign a different γ value to each user and optimize it based on the maximum network capacity. We leave QoS based network optimization problem to the future work.

Next, we analyze the inter-cell distance (d) required in THz Band indoor wireless network. This important parameter is calculated based on the criterion of minimum inter-cell interference for a particular quality of communication. We determine this distance by considering the interference power received by two receiver orientations that are expected to be affected most by inter-cell interference. The first is a receiver placed closest to the cell boundary with its beam pointing in the opposite direction i.e. $G(180^\circ)$, RX_{1,2} depicted in Fig. 7.4. While, the second is when a receiver being at the opposite end of the neighboring cell has its main beam pointing towards the local cell, RX_{2,2} depicted in Fig. 7.4. Finally, minimum inter-cell distance is determined by making sure that the interference power is lower than the system noise floor for both the antenna orientations.

THz Band communications, due to its wide bandwidth, not only provides very high speed wireless communication, it also provides higher ASE due to smaller cell sizes such as femtocells. We determine the ASE by dividing the cell capacity with the network area.

$$ASE = \frac{NC_{\mathbb{R}^2}}{B_c \pi d^2}, \quad (7.15)$$

where, d is the inter-cell distance between the two neighboring cells. This parameter determines how efficiently the frequency spectrum is used in THz Band indoor IoT network.

7.4 Simulation Results

We first determine the interference signal level received by the user under examination as a function of guard width (γ) for multiple distances. This analysis gives us a first insight into the performance of the network. Simulations were performed using the parameters given in Table 7.1. System bandwidth B_c was kept as 4.48 GHz which was determined experimentally in [Kim and Zajić, 2015, Priebe et al., 2013] and the center frequency is set to 300 GHz. Thermal noise power of a 4.48 GHz band system at room temperature is -77.3 dBm and with a noise figure of 7 dB, the total noise power is -70.3 dBm. The value, -70 dBm in Table 7.1 is an approximate value. Gain of antenna used is 26 dBi with a HPBW of 12° , which is easily achievable in THz Band [Roh et al., 2014, Khalid and Akan, 2016b, Lin and Li, 2015]. The path loss of 2 is statistically determined in previous experiments in [Kim and Zajić, 2015]. Due to hardware limitations, generating high signal power is very difficult, however, 0 dBm of transmitter power is achievable as shown in [Kallfass et al., 2015a] and hence used in our simulations. Antenna radiation pattern used in simulations was calculated using $\zeta = 4^\circ$. Finally, to satisfy the condition $K_r \leq N$, the receiver is always assigned a fixed value of $K_r = 1$.

The received interference signal power as a function of γ for different distances is calculated using (7.9) and shown in Fig. 7.5. As expected, it can be seen that the interference signal power decreases with increasing γ . However, an interesting point to be noted here is that there are also some smaller values of γ where the interference drops significantly. This enhancement is due to the suitable placement of the nulls of neighboring beams. It can be inferred that these point can produce interesting results by providing low interference even at smaller γ . Furthermore, at smaller distance the interference signal power is higher than the white noise N_0 for lower γ values, while, for distances greater than 1.5 m, the interference signal power gets lower than the N_0 and hence, will not effect the performance of the receiver. It should be noted here that the negative values of γ depicts the separation between the beams to be even less than the β_{3dB} .

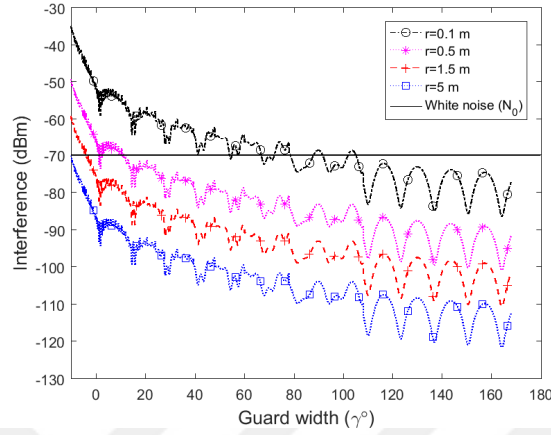


Figure 7.5: Simulation results for total interference power received by the UE for different γ values.

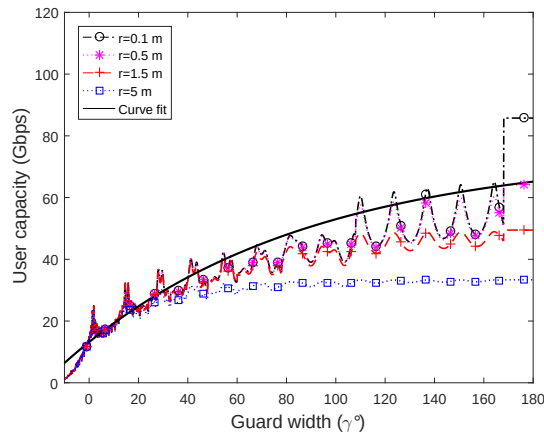


Figure 7.6: Simulation results for capacity of single user for different γ values and distances.

We further examine the effect of γ by determining user capacity in our system by using (7.12) for different distances. It can be seen in Fig. 7.6 that by using the proposed system, we can achieve very high data transmission speed. The user capacity reaches its maximum at 168.1° . Whereas the spikes in the results confirm that the proper placement of the nulls of neighboring beams can be very advan-

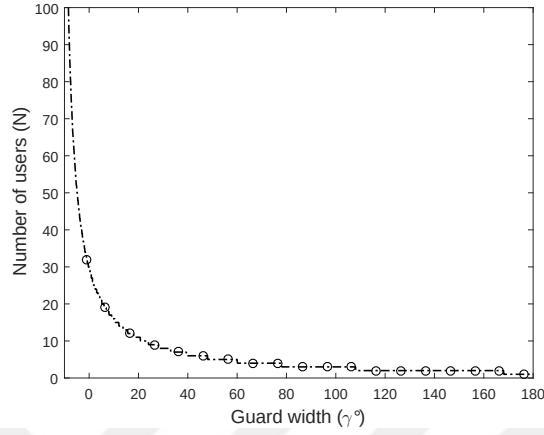


Figure 7.7: Simulation results for maximum number of users at a given γ .

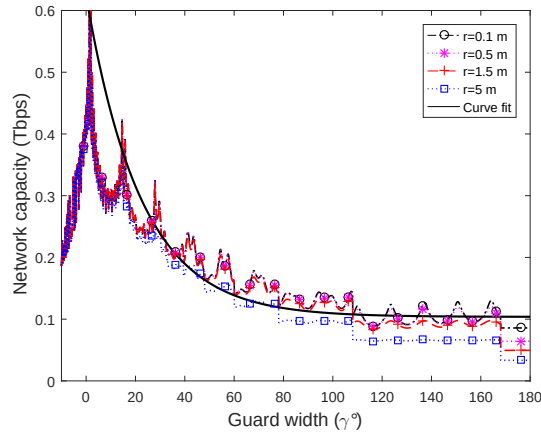


Figure 7.8: Simulation results for capacity of the network for different γ values and distances.

tageous for the system as it provides higher SINR at lower γ . Further, we have evaluated an exponential approximation curve that represents the user capacity by $(-61.24e^{-0.01049\gamma} + 74.43) \times 10^9$. This curve fit is used to analyze the system performance trade-off when combined with other curves that are discussed later.

Although the results show that user capacity is increased by increasing γ , (7.7) suggests that the number of possible user connections will decrease. This attribute is

Table 7.1: Setup parameters.

Parameter	Symbol	Value
<i>Bandwidth</i>	B_c	4.48 GHz
<i>Frequency</i>	f	300 GHz
<i>Speed of light</i>	c	299792458 m/s
<i>Average distance</i>	r	0.1, 0.5, 1.5, 5 m
<i>White noise</i>	N_0	-70 dBm
<i>Gain</i>	G	26 dBi
<i>HPBW</i>	β_{3dB}	12°
<i>Path loss exponent</i>	n	2
<i>Transmitter power</i>	P_t	0 dBm
<i>Pattern smoothing</i>	ζ	4°
<i>Receiver</i>	K_r	1

also shown in Fig. 7.7 and it also depicts that the value of γ directly influence the full network performance and therefore, must be carefully determined.

Next we determine the network performance by using (7.13). It can be seen in Fig. 7.8 that the network performance degrades exponentially as γ increases which is due to the fact that the number of possible user connections are decreased. It should be noted here that, for negative values of γ , the network capacity reduces significantly. This is due to the fact that the interference at negative values of γ is very high and thus, a higher number of users does not increase the over all network performance. The optimum network capacity is achieved at $\gamma = 1.32^\circ$. Here, we again provide a curve fit that approximates the network capacity using an exponential equation $(0.5205e^{-0.045\gamma} + 0.1037) \times 10^{12}$. Finally, as in the earlier cases, we again see spikes in the results occurring due to the suitable placement of the nulls in the network capacity. To further investigate this behavior, we present the radiation pattern of the transmitter at $\gamma = 1.32^\circ$ in Fig. 7.9. Although, in practice, such precise placement

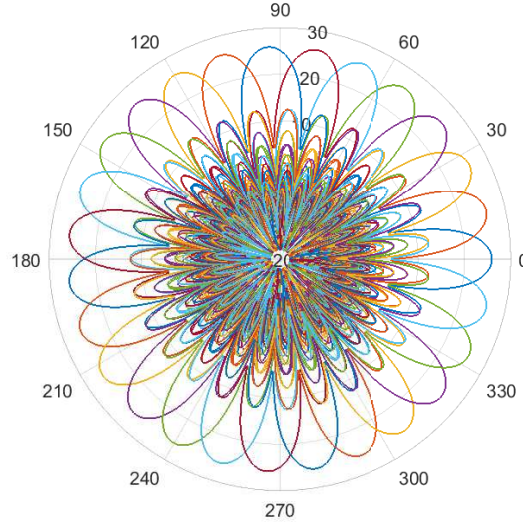


Figure 7.9: Radiation pattern FD-MIMO with $\gamma = 1.32^\circ$.

of antenna might seem difficult, it is possible to place the antenna beams on precise locations by using advanced beam steering techniques as shown in [Sadhu et al., 2017].

It is now evident that a suitable selection of γ is very important for FD-MIMO systems. Hence, to visualize the trade-off process for selecting optimal γ in any scenario, we provide the normalized curve fit graphs of user capacity and network capacity in Fig. 7.10. For instance, in case a guard width of 1.32° is selected, the network performance is optimum, however, the graph depicts that the user performance will be 22% of the maximum performance. An interesting value is where both the curves intersect each other at $\gamma = 27^\circ$. This point is the balanced optimum for both the parameters, as both are operating at 44.26% of their respective maximum values. Similarly, in cases where the objective is to optimize user capacity, $\gamma = 168.1^\circ$ must be used, while the network performance at this γ will be 17.5%. These curves can significantly help in developing optimization algorithms required for FD-MIMO based THz Band indoor IoT networks.

Due to higher path loss, THz Band network provides higher frequency reuse. To determine the minimum inter-cell distance, we observe the interference signal power

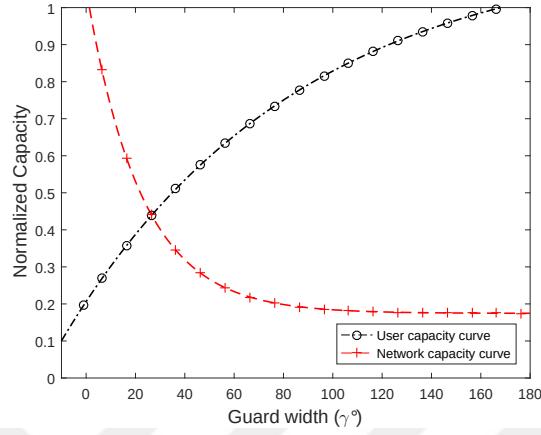


Figure 7.10: Normalized capacity curves to find optimum Guard width (γ) based on the requirements.

received by two receivers of the neighboring cell. The results for both the receivers are shown in Fig. 7.11. It can be seen that the neighboring cell receiver facing in the opposite direction receives an interference signal that stays mostly below the system white noise N_0 . Hence to determine the inter-cell distance, we ensure that the power of interference signal received by the neighboring cell receiver facing directly toward the local receiver remains below -70 dBm. We can observe from Fig. 7.11 that at an inter-cell distance of 1.6 m, the interference power goes below the white noise N_0 .

Finally, we determine the ASE of FD-MIMO based THz Band network. For an inter-cell distance of 1.6 m and $\gamma = 1.32^\circ$, the ASE reaches up to 4.6 b/s/Hz/m² as shown in Fig. 7.12. The ASE of THz Band network is much higher compared to lower frequency networks and can significantly help alleviating high data rates problem in future indoor 5G networks [Khalid and Akan, 2016b].

7.5 Conclusion

In this chapter, we presented an analysis based on a novel node separation scheme to characterize the performance of future THz Band IoT networks. We presented the concept of guard width (γ), that is a beam separation in access to the antenna

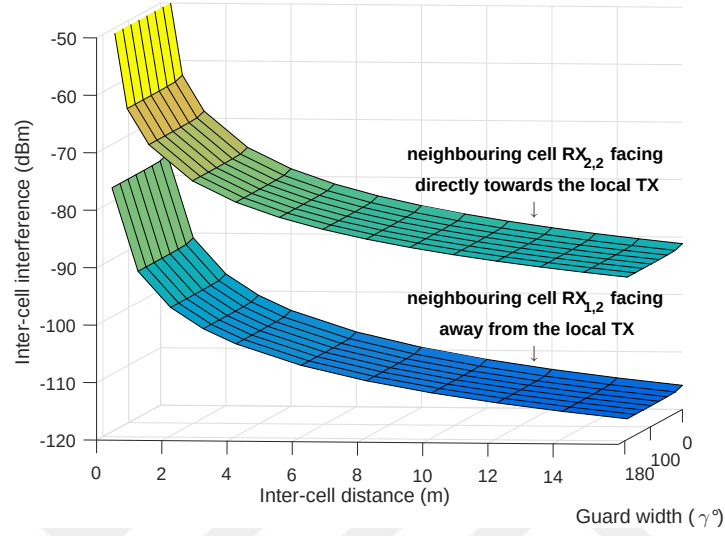


Figure 7.11: Distance required for minimum inter-cell interference.

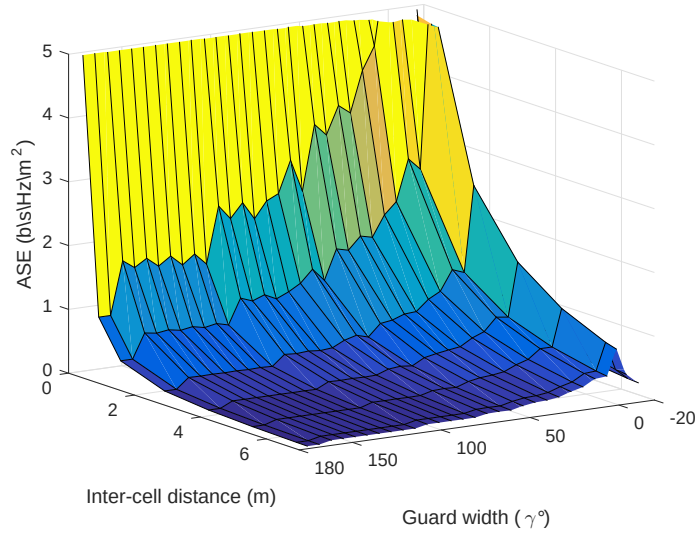


Figure 7.12: Change in area spectral efficiency based on inter-cell distance.

HPBW, to mitigate the effect of interference. By applying this concept on a THz Band network, we analyzed interference as well as user and network performance to determine the optimum value of guard width in a number of scenarios. We also

presented a trade-off that may assist in robust selection of guard width based on the target IoT applications. We have shown that nodes operating in THz Band can reach tens of Gbps speed while the networks can provide Tbps which is an order-of-magnitude more speed as compared to lower frequency networks. In addition to the stated parameters, we have also determined the minimum inter-cell distance required in a THz Band network and shown that it can accommodate a high number of users. Finally, we have shown the area spectral efficiency of the network can reach 1000 times more than the lower frequency networks by using optimum guard width values.

Chapter 8

CONCLUSIONS AND FUTURE WORK

8.1 Contributions

In This thesis, we first have presented Terahertz Band channel characteristics using Vector Network Analyzer and subharmonic mixers based measurement system. Channel impulse response from 260 GHz to 400 GHz are recorded and presented. System capabilities and parameters were investigated in detail. We have measured simple transmission coefficient for LOS link to analyze basic capabilities and practical success of THz Band communication system. Measured results show that Tbps throughput is achievable in THz Band. It has become evident that THz Band will enable a plethora of applications such as high speed mobile communications, wireless local area networks or interconnection of gadgets on desktop like HDMI and flash memory. Results for LOS measurements confirm that high gain antennas are necessary to mitigate the path losses at THz frequencies. In addition, the experimental setup demonstrated in this work is a significant example that can boost studies on 5G indoor wireless communication. Furthermore, we have provided and compared the statistical characteristics of several low-Terahertz Band channels, namely, the 300 to 319 GHz band, 340 to 359 GHz band and 380 to 399 GHz band. Around 600 different measurements were recorded with spatial and temporal variations, that included LOS and NLOS environments, for all the three potential bands. The large-scale path loss models for all the three bands were evaluated using the single slope path loss model with shadowing. Our work shows that the models are quite comparable as the path loss exponent for 310 GHz band, 350 GHz band and 390 GHz band is 2.07, 1.90 and 1.96, respectively, while the variations due to shadowing were confirmed to be normally distributed, for all the three bands. Moreover, NLOS measurements were

performed by using the optically opaque objects as reflectors to measure the path loss. It was observed that metals, wood and even acoustic ceiling panel can be used as a reflector in THz Band channel. In addition, small-scale multipath propagation was also statistically analyzed in an indoor environment. The channel impulse response was calculated by taking IDFT of the measured frequency response. It was observed that strong multipath components were present in LOS and NLOS environments. To describe the multi path properties of the channel, we have calculated mean excess delay, RMS delay spread and the coherence bandwidth. The results show that the high RMS delay reduces the coherence bandwidth a lot, making it necessary to use equalizers in THz Band communications. This finding suggests that efficient equalizer should be designed in order to fully benefit from the wide available bandwidth of THz Band channel. Finally, we have also calculated the autocorrelation function of the channel impulses response and it was observed that at times the channel might show correlation higher than 0.2.

We have also analyzed the total energy consumption of a wireless communication system operating in THz Band channel. System models are based on MQAM modulation and MFSK modulation scheme. THz Band applications are expected to provide high data throughput that will consume a lot of energy, therefore, we focus to devise modulation schemes that will minimize the energy consumption per information bit. Based on our analysis, we have shown that short-range THz Band applications such as wireless computer gadgets, should use MQAM modulation scheme. Whereas, medium and long range such as 5G cellular networks and ultra-high-speed wireless backhaul should use MFSK modulation scheme to achieve energy-efficient communications.

In addition, we have proposed a broadband transceiver design for 300 GHz band wireless communication based on discrete components that may be used in future IoT applications. We use low cost industry ready components that can be easily modified based on application specific requirements. The proposed transceiver provides a bandwidth of 20 GHz making it a good candidate for exploring wide band channel capabilities around the 300 GHz frequency band for 5G applications of IoT.

Component parameters are used to simulate the system design to ensure first pass success. Linearity, noise and selectivity characteristics of the system are analyzed based on single-tone, two-tone, receiver gain, noise figure, and phase noise analysis, which provides detailed insight of our proposed system. Finally, it is shown that our system is capable of providing 90.31 Gbps data throughput at 2 m.

Furthermore, we have presented THz Band 2×2 LOS MIMO system operating at 7 Gbps. Spatial diversity scheme is utilized with antenna element spacing derived from the principles of diffraction limited optics. The channel transmission and inter channel interference is measured and recorded using virtual antenna array technique and channel sounder based on VNA and frequency extenders. We have shown that cross channel link vectors are out of phase and can be suppressed using channel separation network. Optimal channel separation parameters were determined for minimal cross channel interference. Finally, THz Band 2×2 MIMO link was evaluated for digital communication system and confirmed to achieve 7 Gbps. However, a significant difference was seen between the theoretical and experimental throughputs that opens up the need for further research. Moreover, our results suggest that single carrier might not provide the ultimate benefit of such a wireless communication channel, instead multi-carrier modulation will perform better in terms of spectral efficiency, given its relative immunity to fading and interference.

Lastly, we have presented an analysis based on a novel node separation scheme to characterize the performance of future THz Band IoT networks. We presented the concept of guard width (γ), that is a beam separation in access to the antenna HPBW, to mitigate the effect of interference. By applying this concept on a THz Band network, we analyzed interference as well as user and network performance to determine the optimum value of guard width in a number of scenarios. We also presented a trade-off that may assist in robust selection of guard width based on the target IoT applications. We have shown that nodes operating in THz Band can reach tens of Gbps speed while the networks can provide Tbps which is an order-of-magnitude more speed as compared to lower frequency networks. In addition to the

stated parameters, we have also determined the minimum inter-cell distance required in a THz Band network and shown that it can accommodate a high number of users. Finally, we shown the area spectral efficiency of the network can reach 1000 times more than the lower frequency networks by using optimum guard width values.

8.2 Future Research Directions

Further case study on this topic can include larger multiple-input multiple-output (MIMO) antenna technique. MIMO is a very popular technique that can enhance the capacity of a wireless link by introducing multiplexing gain. The basic idea is to use multiple transmit and receive antennas to exploit the multipath behavior of the channel. Although the results show that the THz band channel has very weak multipath components, but as long as reflectors like metal or even ceiling panel exists in the path, MIMO systems in THz band can be implemented. Our setup can be easily modified to implement virtual antenna arrays to observe MIMO characteristics of the channel as shown in [Ranvier et al., 2009]. Moreover, systems with larger one-dimensional and two-dimensional arrays transmitting real time data in different indoor environments. Performance of multi-carrier modulation system may also be analyzed.

For the proposed transceiver, future work can include analysis of digital data transmission on the proposed transceiver to determine practically achievable data rates.



Appendices

Appendix 1

MATLAB CODE FOR ENERGY-EFFICIENT MODULATION SCHEME

A.1 MQAM Matlab code

```
1 % This program simulates energy per bit utilization of a THz
   band
2 % transceiver operating using MQAM modulation. the programm
3 % considers channel to be an AWGN. The architecture used is
   multiplier
4 % based semiconductor system
5 %
6 %
7 %
8 %
9 clear all
10 close all
11 clc
12 fc=300e9;% center frequency of the system
13 % n=0-(1/19);
14 % for index=1:20 %If using this loop, the d should be
   singleton
15 %     n=n+((1/19))
16 k=4;
17 B=1e9;
```

```

18 Pmix=30.3e-3;
19 Plna=20e-3;
20 Pmaxt=250e-3;
21 Ttr=5e-6;
22 T=100e-3
23 Nfdb=7.5;
24 Nf=10^((Nfdb)/10);
25 n=0.2;
26 psddb=-174;
27 psd=10^((psddb)/10)*1e-3
28 L=200e6;
29 Psyn=50e-3;
30 Pifa=3e-3;
31 Pfilt=2.5e-3;
32 Pfilr=2.5e-3;
33 Mldb=17;
34 Ml=10^((Mldb)/10);
35 Pbex=3;
36 Pb=10^(-Pbex)
37 Gldb=30;
38 G1=10^((Gldb)/10);
39 Vdd=3;
40 Lmin=0.5e-6;
41 n1=10;
42 n2=10;
43 fcor=1e6;
44 Io=10e-6;
45 Cp=1e-12;
46 Beta=1;

```

```

47 Pdac=Beta*((0.5*Vdd*Io*((2^n1)-1))+(n1*Cp*(2*B+fcor)*(Vdd^2))
    );
48 Padc=(3*(Vdd^2)*Lmin*(2*B+fcor))/(10^(((−0.1525)*n2)+4.838));
49 Ton=0.01:0.001:0.1
50 % Ton=0.01;
51 b=L./(B.*Ton);
52 M=2.^(L./(B.*Ton));
53 z=3.*((sqrt(M)-1)./sqrt(M)+1);
54 a=(z./n)-1;
55 Pct=Pmix+Psyn+Pfilt+Pdac;
56 Pcr=Pmix+Psyn+Plna+Pfilr+Pifa+Padc;
57 Pc=Pct+Pcr
58 d=[0.1:1:30] %If using this loop, the d should be singleton
59 % d=100
60 Gd=G1.*(d.^k).*Ml
61 for x=1:length(d)
62 Ea(x,:)=((1+a).*(4/3).*((Nf.*psd.*((2.^b)-1))./b).*(log
    ((4.*(1-(2.^((-b)./2)))))./(b.*Pb))).*Gd(x)) + (((Pc.*Ton)
    + (2.*Psyn.*Ttr))./L);
63 Eat(x,:)=((1+a).*(4/3).*((Nf.*psd.*((2.^b)-1))./b).*(log
    ((4.*(1-(2.^((-b)./2)))))./(b.*Pb))).*Gd(x));
64 end
65 Eadb=10*log10(Ea.*1000);
66 Eatdb=10*log10(Eat.*1000);
67 [Mi,In]=min(Eadb. ');
68 optton=Ton(In);
69 optton2=optton. ';
70 for x=1:length(d)
71 optEa(x)=Ea(x,In(x));

```

```

72 end
73 optEa2=optEa.';
74 optEa2db=10*log10(optEa2.*1000) %If using this loop, use
    Ea2db(index)
75 optb=b(In);
76 optb2=optb.';
77 r=L./(optton);
78 r2=r.';
79 % end %If using this loop, the d should be singleton
80 % optEa3db=optEa2db.';% If using this loop, the d should be
    singleton
81 % figure(1)
82 % plot (Ton/T,Eadb)
83 % % ylim([-30 50])
84 % xlim([0 1])
85 % legend('d=1 m','d=5 m','d=30 m','d=100 m')
86 % xlabel('Ton/T')
87 % ylabel('Energy per information bit in dBmJ')
88 % figure(2)
89 % plot(b,Eadb)
90 % legend('d=1 m','d=5 m','d=30 m','d=100 m')
91 % xlabel('b')
92 % ylabel('Energy per information bit in dBmJ')
93 % figure(3)
94 % plot(b,Eadb(2,:),b,Eatdb(2,:))

```

A.2 MFSK Matlab code

```

1 % This program simulates energy per bit utilization of a THz
    band

```

```
2 % transceiver operating using MFSK modulation. the programm
3 % considers channel to be an AWGN. The architecture used is
    multiplier
4 % based semiconductor system
5 %
6 %
7 %
8 %
9 clear all
10 close all
11 clc
12 fc=300e9;
13 % n=0-(1/19);%If using this loop, the d should be singleton
14 % for index=1:20 %If using this loop, the d should be
    singleton
15 %      n=n+((1/19))
16 k=2.8;
17 B=1e9;
18 Pmix=30.3e-3;
19 Plna=20e-3;
20 Pmaxt=250e-3;
21 Ttr=5e-6;
22 T=1.07;
23 Nfdb=5.2;
24 Nf=10^((Nfdb)/10);
25 n=0.42;
26 psddb=-174;
27 psd=10^((psddb)/10)*1e-3
28 L=200e6;
```

```

29 Psyn=50e-3;
30 Pifa=3e-3;
31 Pfilt=2.5e-3;
32 Pfilr=2.5e-3;
33 Mldb=31;
34 Ml=10^((Mldb)/10);
35 Pbex=3;
36 Pb=10^(-Pbex);
37 G1db=30;
38 G1=10^((G1db)/10);
39 Vdd=3;
40 Lmin=0.5e-6;
41 n1=10;
42 n2=10;
43 fcor=1e6;
44 Io=10e-6;
45 Cp=1e-12;
46 % Beta=1; % not used in MFSK
47 % Pdac=Beta*((0.5*Vdd*Io*((2^n1)-1))+(n1*Cp*(2*B+fcor)*(Vdd
    ^2)))% not used in MFSK
48 Padc=(3*(Vdd^2)*Lmin*(2*B+fcor))/(10^((( -0.1525)*n2)+4.838))
49 % Ton=0.01:0.001:0.1% not used in MFSK
50 % Ton=0.01;% not used in MFSK
51 z=1;
52 a=(z./n)-1
53 % b=L./(B.*Ton)% not used in MFSK
54 b=1:6
55 M=2.^b
56 Ton=(L.*(2.^b))./(2.*b.*B)

```

```

57 Pct=Psyn+Pfilt
58 Pcr=Pmix+Psyn+Plna+Pfilr+Pifa+Padc
59 Pc=Pct+Pcr
60 d=[0.1:1:30] %If using this loop, the d should be singleton
61 % d=100
62 Gd=G1.*(d.^k).*Ml
63 for x=1:length(d)
64 Ea(x, :)=((1+a).*(4).*((Nf.*psd)./b).*(log((((2.^(b-2)))./(Pb
        )))).*Gd(x)) + (((Pc.*Ton) + (2.*Psyn.*Ttr))./L)
65 Eat(x, :)=(1+a).*(4).*((Nf.*psd)./b).*(log((((2.^(b-2)))./(Pb
        )))).*Gd(x)
66 end
67 Eadb=10*log10(Ea.*1000)
68 Eatdb=10*log10(Eat.*1000)
69 [Mi, In]=min(Eadb. ');
70 optton=Ton(In);
71 optton2=optton. ';
72 for x=1:length(d)
73 optEa(x)=Ea(x, In(x));
74 end
75 optEa2=optEa. ';
76 optEa2db=10*log10(optEa2.*1000)%If using this loop, use Ea2db
        (index)
77 optb=b(In);
78 optb2=optb. ';
79 r=L./(optton);
80 r2=r. ';
81 % end %If using this loop, the d should be singleton
82 optEa3db=optEa2db. ';

```

```
83 % figure(1)
84 % plot (b,Eadb)
85 % ylim([-50 -10])
86 % % xlim([1 6])
87 % legend('d=0.1 m','d=1 m','d=3 m','d=5 m','d=10 m','d=20
      m')
88 % xlabel('b')
89 % ylabel('Energy per information bit in dBmJ')
90 % figure(2)
91 % plot(b,Eadb(2,:),b,Eatdb(2,:))
```

Appendix 2

MATLAB CODE FOR CAPACITY AND COVERAGE ANALYSIS OF FD-MIMO BASED CHANNEL

B.1 FD-MIMO Matlab code

```
1 %% Main code
2 clear all
3 close all
4 clc
5
6 %Constants
7 c=299792458; %speed of light
8
9 %transmitter parameters
10 f=300e9;%transmitter frequency
11 Bc=4.48e9;%coherence BW from my paper
12 wdb=-70;%dBm white noise
13 w=db2pow(wdb)*1e-3;
14 ptdb=0; %in dBm
15 pt=db2pow(ptdb)*1e-3; %power in watts
16
17 n=2.0737;% path loss exponent
18 fconst=c/(4*pi*f);% path loss
19 %Antenna
20 dthetadeg=0.05;
```

```

21 dthetarad=degtorad(dthetadeg);
22 thetadeg=-180:dthetadeg:180;
23 % theta=linspace(-pi,pi,1000);
24 theta=degtorad(thetadeg);
25 gaindbtx=26; % gain in db
26 gaintx=10^(gaindbtx/10);%gain
27 gaindbrx=26; % gain of receiver in db (-8.65@180 degrees)
28 gainrx=10^(gaindbrx/10);%gain
29 bxydeg=12;%HPBW in degrees
30 bxy=degtorad(bxydeg);
31 amfdeg=360;%antenna movement freedom in degrees
32 dtheta=0.1;
33 dthetafine=0.0528;
34 dispjump=12700;%25 for dtheta=0.5 320,1270 for dtheta=0.1 to
    set how frequent plot should be displayed
35 gwdeg=[-11:dtheta:0-dtheta 0:dthetafine:3-dthetafine 3:dtheta
    :180];%guard width in degrees(-11:dtheta:180)
36 gw=degtorad(gwdeg);
37 func=sinc((0.8854/bxy)*theta).^2;
38 Gdb=gaindbtx+(10*log10(func));
39 zetaangledeg=4;% angles in degrees included in smoothing for
    either side
40 zeta=zetaangledeg*(1/dthetadeg);
41 func_smooth=smooth(sinc((0.8854/bxy)*theta).^2,zeta);
42 fdb_smooth=10*log10(func_smooth);
43 Gdb_smooth=gaindbtx+fdb_smooth;
44 thetazeroI=find(thetadeg==0);
45 theta180I=find(thetadeg==180);
46 grx=[thetazeroI theta180I];

```

```

47 dr=1;
48 r=[0.1 0.5 1 1.5 2:dr:30];% distance of UE
49 %*****
50 %values at which we need graph
51 gwdeg_graph=1.32;
52 r_graph=0.5;
53 % Plotting of the antenna radiation pattern
54 figure(1)
55 polarplot(theta,max(Gdb,-20)+20,'LineWidth',1.5)
56 set(gca,'RTick',[0 10 20 30 40 50])
57 set(gca,'RTickLabel',[-20 -10 0 10 20 30])
58 dim = [0.38 0.048 0.3 0.3];
59 str = {'HPBW: 12\circ','FNBW: 27\circ','Main: 26 dB @ 0\circ',
        'Back: -8.97 dB @ 180\circ'};
60 annotation('textbox',dim,'String',str,'FitBoxToText','on','
        BackgroundColor','white','FontWeight','bold','FontSize',9)
        ;
61 % Plotting of the smooth antenna radiation pattern
62 figure(2)
63 polarplot(theta,max(Gdb_smooth,-20)+20,'LineWidth',1.5)
64 set(gca,'RTick',[0 10 20 30 40 50])
65 set(gca,'RTickLabel',[-20 -10 0 10 20 30])
66 dim = [0.38 0.04 0.3 0.3];
67 str = {'HPBW: 12\circ','Main: 26 dB @ 0\circ','Back: -8.97 dB
        @ 180\circ','\zeta: 4\circ'};
68 annotation('textbox',dim,'String',str,'FitBoxToText','on','
        BackgroundColor','white','FontWeight','bold','FontSize',9)
        ;
69 %% Main loop

```

```

70
71 for kgrx=1:length(grx)
72
73
74     for kr=1:length(r)
75
76         %*****
77         k=1;% UE being evaluated; it should remain because of
              cond.i ~= k or can be made random based on value
              of N
78         l=ceil(length(gw)*0.5);%GW index at which the
              radiation plot will be evaluated
79         gwpatdeg=gwdeg(l);
80         for j=1:length(gwdeg)
81             %Users
82             N(j)=floor(amfdeg/(bxydeg+gwdeg(j)));%number of
              independent beams
83             int=0;%interference initialization: by each user
84             %SINR
85             %interested user
86             % Gi=gain*sinc((0.8854/bxy)*thetaik).^2;
87             interference=0;%total interference initialization
88             count=0;
89             for i=1:N(j)
90                 if i ~= k
91                     count=count+1;
92                     [funcmin,funcmin_I]=min(abs(theta-(i-k)*(
              bxy+gw(j)))));
93                     thetaintdeg=radtodeg(wrapToPi((i-k)*(bxy+

```

```

        gw(j)));
94         thetaintI=thetazeroI+round(wrapToPi((i-k)
            *(bxy+gw(j)))/dthetarad);
95         int(count) = (fconst)^n*(pt)*(1/r(kr))^n
            * gainrx*func_smooth(grx(kgrx)) *
            gaintx*func_smooth(thetaintI);
96         interference=interference+int(count);
97     end
98     %         if (mod(j,dispjump*dtheta)==0) && (r(kr)
        ==0.5) && (grx(kgrx)==thetazeroI)
99         if (gwdeg(j)==gwdeg_graph) && (r(kr)==r_graph
            ) && (grx(kgrx)==thetazeroI)
100             figure (3)
101             %                 polarpattern(thetadeg+(i-k)*(bxydeg+
        gwdeg(j)),max(Gdb,-20),'TitleTop',[ 'antenna distribution
        with GW = ',num2str(gwdeg(j))])
102             polarplot(degtorad(thetadeg+(i-k)*(bxydeg
                +gwdeg(j))),max(Gdb_smooth,-20)+20,'
                LineWidth',1.1)
103             drawnow
104             hold on
105             %
106             %                 title(sprintf('GW=%d degrees',gwdeg(j))
        );
107
108     end
109
110 end
111 if (gwdeg(j)==gwdeg_graph) && (r(kr)==r_graph) &&

```

```

        (grx(kgrx)==thetazeroI)
112 %         if mod(j,dispjump*dtheta)==0
113 %             pause;
114             set(gca,'RTickLabel',[-20 -10 0 10 20 30])
115             hold off
116         end
117
118         interferencedb(j,kr,kgrx)=pow2db(interference);
119         signal(kr,kgrx)=(((fconst)^n)*(pt)*((1/r(kr))^n)
            * gainrx*func_smooth(grx(kgrx)) * gaintx);
120         signaldb(kr,kgrx)=pow2db(signal(kr,kgrx));
121         SINR(j,kr,kgrx)=(signal(kr,kgrx))/(w+interference
            );
122         SINRdb(j,kr,kgrx)=pow2db(SINR(j,kr,kgrx));
123         guardwidth=gwdeg(j);
124         user_C(j,kr,kgrx)=Bc*log2(1+SINR(j,kr,kgrx));
125         % C=Bc*log2(1+db2pow(47.72))
126         maxnetwork_C(j,kr,kgrx)=user_C(j,kr,kgrx)*N(j);
127         intercell_interference(j,kr,kgrx)=signal(kr,kgrx)
            +interference;
128         intercell_interferencedb(j,kr,kgrx)=pow2db(
            intercell_interference(j,kr,kgrx));
129         ASE(j,kr,kgrx)=maxnetwork_C(j,kr,kgrx)/(4*pi*r(kr)
            )*Bc);
130     end
131 end
132 end
133 %% Distances and Guard width at which plots will be evaluated
134 UE_dist=0.1;

```

```

135 UE_dist2=0.5;
136 UE_dist3=1.5;
137 UE_dist4=5;
138 gwdeg_start=0;
139 % index values of these parameters
140 UE_distI=find(r==UE_dist);
141 UE_distI2=find(r==UE_dist2);
142 UE_distI3=find(r==UE_dist3);
143 UE_distI4=find(r==UE_dist4);
144 gwdeg_startI=find(gwdeg==gwdeg_start);
145
146 %% Plotting interference seen by user (K=1) facing the TX(
    kgrx=1) at 5 different distances (r=0.1,0.5,1.5,5)
147
148
149 style1='-.'; style2=': '; style3='—'; style4=': '; style5='-';
    style6='-.';
150 color1='k'; color2='m'; color3='r'; color4='b'; color5='k';
    color6='r';
151 marker1='o'; marker2='*'; marker3='+'; marker4='s'; marker5='
    x'; marker6='v';
152 gaps=10*(1/dtheta);
153 %*****dummy points for legend
154 x3=-20;
155 y3=1;
156 %*****
157 figure (4)
158 %dummy plots to cater for legend
159 h1=plot(x3,y3,[color1 style1,marker1]);

```

```

160 hold on;
161 h2=plot(x3,y3,[color2 style2,marker2]);
162 h3=plot(x3,y3,[color3 style3,marker3]);
163 h4=plot(x3,y3,[color4 style4,marker4]);
164 h5=plot(x3,y3,[color5 style5]);
165 % main line plotting
166 plot(gwdeg,interferencedb(:,UE_distI,1),[color1 style1], '
    LineWidth',1.1);
167 plot(gwdeg,interferencedb(:,UE_distI2,1),[color2 style2], '
    LineWidth',1.1);
168 plot(gwdeg,interferencedb(:,UE_distI3,1),[color3 style3], '
    LineWidth',1.1);
169 plot(gwdeg,interferencedb(:,UE_distI4,1),[color4 style4], '
    LineWidth',1.1);
170 plot(gwdeg,(wdb)*ones(length(gwdeg)),[color5 style5], '
    LineWidth',1.1);
171 % Scatter plot for marks
172 plot(gwdeg(1:gaps:end), interferencedb(1:gaps:end,UE_distI,1)
    ,[color1 marker1], 'markers',6);
173 plot(gwdeg(1:gaps:end), interferencedb(1:gaps:end,UE_distI2
    ,1),[color2 marker2], 'markers',6);
174 plot(gwdeg(1:gaps:end), interferencedb(1:gaps:end,UE_distI3
    ,1),[color3 marker3], 'markers',6);
175 plot(gwdeg(1:gaps:end), interferencedb(1:gaps:end,UE_distI4
    ,1),[color4 marker4], 'markers',6);
176 xlabel('Guard width ( $\gamma_{\text{circ}}$ )','FontSize',13)
177 ylabel('Interference (dBm)','FontSize',13)
178 axis tight;
179 axis([-10 180 -130 -30])

```

```

180 legend([h1,h2,h3,h4,h5],{'r=0.1 m','r=0.5 m','r=1.5 m','r=5
    m','White noise (N_0)'},'FontSize',9,'Location','
    northeast');
181 hold off
182 %% Finding optimum Guard width value and index for maximum
    user capacity and determine max number of users
183 % to enumerate maximum network capacity for the distance of
    0.1 m
184 [maxc,maxcI]=max(user_C(:,UE_distI,1));
185 optgw_user=gwdeg(maxcI)
186 num_d=N(maxcI)
187 network_C=user_C(maxcI,1)*N(maxcI)
188 [maxNC,maxNCI]=max(maxnetwork_C(:,UE_distI,1))
189 num_maxNd=N(maxNCI)
190 optgw_network=gwdeg(maxNCI)
191 %% Curve fitting
192 %cftool
193 %use custom equation 'a*exp(-b*x)+c'
194 %Click 'exclude outlier' and exclude some points to improve
    the shape
195 gwdegnew=gwdeg(maxNCI:5:end);
196 user_Cnew=user_C(maxNCI:5:end,1,1)*1e-10;
197 user_C_curve=(-6.124*exp(-0.01049*gwdeg)+7.443)*1e10;
198 user_C_curve_normalized=user_C_curve/user_C_curve(maxcI);
199 maxnetwork_C_new=maxnetwork_C(maxNCI:5:end,1,1)*1e-11;
200 abc=5.207;
201 maxnetwork_C_curve=(abc*exp(-0.045*gwdeg)+1.037)*1e11;
202 maxnetwork_C_curve_normalize=maxnetwork_C_curve/
    maxnetwork_C_curve(maxNCI);

```

```

203 %% Not useful , Plot to optimize paramters of cureve fitting
204
205 % figure
206 % plot(gwdeg,user_C(:,1,1))
207 % hold
208 % plot(gwdeg,user_C_curve)
209 %% Not useful , Plot to optimize paramters of cureve fitting
210
211 % figure
212 % plot(gwdeg,maxnetwork_C(:,1,1))
213 % hold
214 % plot(gwdeg,maxnetwork_C_curve)
215 %% Plotting capacity of User(K=1) facing the TX(kgrx=1) at 5
    different distances (r=0.1,0.5,1.5,5)
216 figure(5)
217 % subplot(2,2,1)
218 %dummy plots to cater for legend
219 h1=plot(x3,y3,[color1 style1 ,marker1]);
220 hold on;
221 h2=plot(x3,y3,[color2 style2 ,marker2]);
222 h3=plot(x3,y3,[color3 style3 ,marker3]);
223 h4=plot(x3,y3,[color4 style4 ,marker4]);
224 h5=plot(x3,y3,[color5 style5]);
225 % main line plotting
226 plot(gwdeg,user_C(:,UE_distI,1)*1e-9,[color1 style1], '
    LineWidth',1.1);
227 plot(gwdeg,user_C(:,UE_distI2,1)*1e-9,[color2 style2], '
    LineWidth',1.1);
228 plot(gwdeg,user_C(:,UE_distI3,1)*1e-9,[color3 style3], '

```

```

        LineWidth',1.1);
229 plot(gwdeg, user_C(:, UE_distI4,1)*1e-9,[color4 style4], '
        LineWidth',1.1);
230 plot(gwdeg, user_C_curve*1e-9,[color5 style5], 'LineWidth'
        ,1.5);
231 % Scatter plot for marks
232 plot(gwdeg(1:gaps:end), user_C(1:gaps:end, UE_distI,1)*1e-9,[
        color1 marker1], 'markers',6);
233 plot(gwdeg(1:gaps:end), user_C(1:gaps:end, UE_distI2,1)*1e-9,[
        color2 marker2], 'markers',6);
234 plot(gwdeg(1:gaps:end), user_C(1:gaps:end, UE_distI3,1)*1e-9,[
        color3 marker3], 'markers',6);
235 plot(gwdeg(1:gaps:end), user_C(1:gaps:end, UE_distI4,1)*1e-9,[
        color4 marker4], 'markers',6);
236 % title(['Capacity of user = ',num2str(k)])
237 xlabel('Guard width (\gamma\circ)', 'FontSize',13)
238 ylabel('User capacity (Gbps)', 'FontSize',13)
239 axis tight;
240 axis([-10 180 0 120])
241 legend([h1,h2,h3,h4,h5],{'r=0.1 m', 'r=0.5 m', 'r=1.5 m', 'r=5
        m', 'Curve fit'}, 'FontSize',9, 'Location', 'northwest');
242 hold off
243 %% not useful
244 % figure(4)
245 %% subplot(2,2,2)
246 % plot(N, user_C(:,1,1)*1e-9)
247 % title(['Capacity of user = ',num2str(k)])
248 % xlabel('Number of users (N)')
249 % ylabel('Capacity (Gbps)')

```

```

250 %% xlim([0 50])
251
252 %% Plotting GW vs Number of users
253 figure(6)
254 plot(gwdeg,N,[color1 style1], 'LineWidth',1.1);
255 hold on
256 plot(gwdeg(1:gaps:end), N(1:gaps:end),[color1 marker1], '
    markers',6);
257 xlabel('Guard width (\gamma\circ)', 'FontSize',13)
258 ylabel('Number of users (N)', 'FontSize',13)
259 axis tight;
260 axis([-10 180 0 100])
261 hold off
262 %% Plotting Network capacity facing the TX(kgrx=1) at 5
    different distances (r=0.1,0.5,1.5,5)
263 % abc=5.707;
264 % maxnetwork_C_curve=(abc*exp(-0.08382*gwdeg)+1.537)*1e11;
265 figure (7)
266 % subplot(2,2,3)
267 %dummy plots to cater for legend
268 h1=plot(x3,y3,[color1 style1,marker1]);
269 hold on;
270 h2=plot(x3,y3,[color2 style2,marker2]);
271 h3=plot(x3,y3,[color3 style3,marker3]);
272 h4=plot(x3,y3,[color4 style4,marker4]);
273 h5=plot(x3,y3,[color5 style5]);
274 % main line plotting
275 plot(gwdeg,maxnetwork_C(:,UE_distI,1)*1e-12,[color1 style1],
    'LineWidth',1.1);

```

```

276 plot(gwdeg, maxnetwork_C(:, UE_distI2, 1) * 1e-12, [color2 style2],
        'LineWidth', 1.1);
277 plot(gwdeg, maxnetwork_C(:, UE_distI3, 1) * 1e-12, [color3 style3],
        'LineWidth', 1.1);
278 plot(gwdeg, maxnetwork_C(:, UE_distI4, 1) * 1e-12, [color4 style4],
        'LineWidth', 1.1);
279 plot(gwdeg, maxnetwork_C_curve * 1e-12, [color5 style5], '
        LineWidth', 1.5);
280 % Scatter plot for marks
281 plot(gwdeg(1:gaps:end), maxnetwork_C(1:gaps:end, UE_distI, 1) *
        1e-12, [color1 marker1], 'markers', 6);
282 plot(gwdeg(1:gaps:end), maxnetwork_C(1:gaps:end, UE_distI2, 1)
        * 1e-12, [color2 marker2], 'markers', 6);
283 plot(gwdeg(1:gaps:end), maxnetwork_C(1:gaps:end, UE_distI3, 1)
        * 1e-12, [color3 marker3], 'markers', 6);
284 plot(gwdeg(1:gaps:end), maxnetwork_C(1:gaps:end, UE_distI4, 1)
        * 1e-12, [color4 marker4], 'markers', 6);
285 % title('Capacity of the Network (NC-{\Re^2})')
286 xlabel('Guard width (\gamma\circ)', 'FontSize', 13)
287 ylabel('Network capacity (Tbps)', 'FontSize', 13)
288 axis tight;
289 axis([-10 180 0 0.6])
290 legend([h1, h2, h3, h4, h5], {'r=0.1 m', 'r=0.5 m', 'r=1.5 m', 'r=5
        m', 'Curve fit'}, 'FontSize', 9, 'Location', 'northeast');
291 hold off
292 %% not useful
293 % figure (7)
294 %% subplot(2,2,4)
295 % plot(N, maxnetwork_C(:, 1, 1) * 1e-12)

```

```

296 % title('Capacity of the Network w.r.t. users')
297 % xlabel('Number of user (N)')
298 % ylabel('Capacity (Tbps)')
299 %% xlim([0 50])
300
301 %% Plotting of normalized curves
302 figure (8)
303 %dummy plots to cater for legend
304 h1=plot(x3,y3,[color1 style1,marker1]);
305 hold on;
306 h2=plot(x3,y3,[color3 style3,marker3]);
307 % main line plotting
308 plot(gwdeg,user_C_curve_normalized,[color1 style1],
        'LineWidth',1.1);
309 plot(gwdeg,maxnetwork_C_curve_normalize,[color3 style3],
        'LineWidth',1.1);
310 % Scatter plot for marks
311 plot(gwdeg(1:gaps:end), user_C_curve_normalized(1:gaps:end),[
        color1 marker1], 'markers',6);
312 plot(gwdeg(1:gaps:end), maxnetwork_C_curve_normalize(1:gaps:
        end),[color3 marker3], 'markers',6);
313 xlabel('Guard width (\gamma\circ)', 'FontSize',13)
314 ylabel('Normalized Capacity', 'FontSize',13)
315 axis tight;
316 axis([-10 180 0 1])
317 legend([h1,h2],{'User capacity curve','Network capacity curve
        '}, 'FontSize',9,'Location','southeast');
318 hold off
319 %% not useful

```

```

320 % figure (9)
321 % rjumps=5;
322 % djumps=50;
323 % surf(r(1:rjumps:end),gwdeg(1:djumps:end),
        intercell_interferencedb(1:djumps:end,1:rjumps:end,1))
324 % title('intercell interference')
325 % zlabel('intercell interference')
326 % xlabel('User distance (meters)')
327 % ylabel('Guard Width (degrees)')
328 % hold on
329 % surf(r(1:rjumps:end),gwdeg(1:djumps:end),
        intercell_interferencedb(1:djumps:end,1:rjumps:end,2))
330 % text(10,1.6,-105,'\downarrow for closest receiver facing
        opposite direction')
331 % text(10,1.6,-70,'\downarrow for farthest receiver facing
        same direction')
332 % xlim([0 18])
333 %% Generating 3D graph that shows How much minimum intercell
        distance is required
334 figure(9)
335 rjumps=1;
336 djumps=25*(1/dtheta);
337 surf((2/3)*r(1:2*rjumps:end),gwdeg(gwdeg_startI:djumps:end),
        intercell_interferencedb(gwdeg_startI:djumps:end,1:2*
        rjumps:end,1))
338 % title('Inter-cell interference')
339 zlabel('Inter-cell interference (dBm)')
340 xlabel('Inter-cell distance (m)')
341 ylabel('Guard width (\gamma\circ)')

```

```

342 hold on
343 surf(2*r(1:rjumps:end),gwdeg(gwdeg_startI:djumps:end),
      intercell_interferencedb(gwdeg_startI:djumps:end,1:rjumps:
      end,2))
344 text(11.5,1.6,-80,{ 'neighbouring cell RX_{2,2} facing', '
      directly towards the local TX', '\downarrow'}, '
      HorizontalAlignment','center','FontWeight','bold')
345 text(11.5,1.6,-105.2,{ 'neighbouring cell RX_{1,2} facing', '
      away from the local TX', '\downarrow'}, 'HorizontalAlignment
      ','center','FontWeight','bold')
346 xlim([0 16])
347 zlim([-120 -50])
348 set(gca,'YTick',[0 100 180])
349 set(gca,'XTick',0:2:14)
350 set(gca,'YDir','Reverse')
351 view(7,5)
352 hold off
353 %% Generating 3D graph to find Area spectral efficiency using
      the receiver facng towards the Tx
354
355 figure(10)
356 rjumps=2;
357 djumps=(maxNCI-1)/2;%6.15*(1/dtheta);
358 hf8=surf((2/3)*r(1:rjumps:end),gwdeg(1:djumps:end),min(6,ASE
      (1:djumps:end,1:rjumps:end,1)))
359 % title('Area Spectral Efficiency')
360 xlabel('ASE (b\\s\\Hz\\m^2)','interpreter','tex')
361 % text1=('ASE (b\\s\\Hz\\')
362 % text2=('m^{2}'))

```

```

363 % xlabel([text1 text2])
364 xlabel('Inter-cell distance (m)')
365 ylabel('Guard width ( $\gamma$ )')
366 xlim([0 7])
367 ylim([-20 180])
368 set(gca,'YTick',[-20 0 50 100 150 180])
369 zlim([0 5])
370 view(57,16)
371 set(gca,'YDir','Reverse')
372 % set(hf8,'edgecolor','none')
373 %% not useful
374 %% hold on
375 % figure(12)
376 % surf(2*r(1:rjumps:end),gwdeg(1:djumps:end),ASE(1:djumps:end
    ,1:rjumps:end,2))
377 % title('Area Spectral Efficiency')
378 % xlabel('ASE (b/s/Hz/m^2)')
379 % ylabel('Inter-cell distance (meters)')
380 % ylabel('Guard Width (degrees)')
381 % xlim([0 6])

```

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