Joonas Kokkoniemi

NANOSCALE SENSOR NETWORKS: THE THz BAND AS A COMMUNICATION CHANNEL

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NANOSCALE SENSOR NETWORKS:
THE THZ BAND AS
A COMMUNICATION CHANNEL

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Abstract

This thesis focuses on THz band channel modeling and characterization. This vast frequency band spans from 100 GHz to 10 THz. The approximately 10,000 GHz bandwidth together with advances in THz capable electronics have made this band highly potential for many future applications, e.g., imaging and nanodevice-to-nanodevice communications. The latter is the reference application of this thesis and it focuses on the communication among very small and simple devices. The main focus of the thesis is on the THz channel characterization. Therefore, the channel models presented herein are also suitable for communications at macroscopic scale.

The THz band offers opportunities, but has many problems as well. One of these is molecular absorption, which causes frequency selective fading to signals. The fading is caused by the signals’ energy absorption in the resonance frequencies of the molecules in the communication medium. Based on the conservation of energy, the absorption is understood to cause a new type of noise in the THz links: transmission induced noise. This noise component is analyzed from multiple physical viewpoints. The THz signals have short enough wavelengths to theoretically allow scattering on aerosols in the atmosphere. Scattering causes frequency dependent loss of the signals, but also a signal spread in time over multiple scattering events. It is shown here that in some specific atmospheric conditions the scattering causes signal loss and time spread. In addition to the theoretical channel models, measurements on a variety of propagation phenomena are conducted and analyzed. These include penetration losses, rough surface reflections and scattering, and diffraction. Through the measurements, it can be shown that the THz band communications is feasible in non-line-of-sight (NLOS) conditions in spite of the above phenomena.

In the last part of this thesis, stochastic geometry is applied to the THz band in order to estimate the mean interference power and outage probabilities in dense networks formed from nanodevices. Because of the large losses in the channel, large interference levels require large numbers of devices. Stochastic geometry offers perfect tools to estimate the mean interference, and also in the case of directional antennas, which are most likely implemented in all the THz band devices due to large power losses in the channel.

Keywords: nanodevice communications, stochastic geometry, THz channel modeling, THz communications
Kokkoniemi, Joonas, Nanoskaalan sensoriverkot: THz-taajuusalue tiedonsiirto-kanavana.
Oulun yliopiston tutkijakoulu; Oulun yliopisto, Tieto- ja sähkötekniikan tiedekunta; Centre for Wireless Communications; Oulun yliopisto, Infotech Oulu
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Oulun yliopisto, PL 8000, 90014 Oulun yliopisto

Tiivistelmä


Työn viimeisessä osassa sovelletaan stokastista geometriaa THz-tauiksille keskimääräisen häiriötehon ja toimintakyvyttömyyden selvittämiseksi tiheissä nanolaitteiden muodostamissa verkoissa. Isojen kanavahäiriöiden takia suuri häiriötasoa vaatii suuremman määrän laitteita. Stokastinen geometria antaa täydelliset työkalut häiriöton eristäminen. Tätä voidaan myös hyödyntää suuntaavien antennien tapauksessa, joita tullaan suurella todennäköisyydellä käyttämään kaikissa THz-laitteissa johtuen suurista signaalihäviöistä kanavassa.

Asiasonat: nanolaitteiden tietoliikenne, stokastinen geometria, THz-kanavamallinnus, THz-tiedonsiirto
To my parents, my siblings,
and my goddaughter Elli
Preface

This doctoral thesis and the research work for it was conducted at the Centre for Wireless Communications (CWC), at the University of Oulu, in Finland, during a time period from 2012 to 2017. The whole doctoral degree process has taught me a great deal. I have been lucky to find a research subject that involves many fields of science. This has significantly helped me to learn and develop my thinking process towards seeing the big picture from multiple angles, to see things from “outside the box”.

I would like to express my gratitude to my principal supervisor, Adjunct Professor, Dr. Janne Lehtomäki. He first took me under his wings to do my Master’s thesis on the spectrum occupancy measurements and analysis and later also for nano and THz communications. He has also greatly helped me to established connections with fellow researchers throughout the world. I would also like to thank my second supervisor and the head of my research group, the Communication Signal Processing group, Professor Markku Juntti, for his excellent advice and help throughout my doctoral training. He has an excellent experience and knowledge on a wide variety of research areas, which has helped me a lot. He is also very well known for his excellent writing skills, and this has improved my own writing skills significantly.

I wish to thank the reviewers of the thesis, Professor Antti Räisänen from Aalto University, Finland, and Associate Professor Eduard Alarcón from Technical University of Catalunya, Spain, for their valuable comments and suggestions regarding the thesis.

I would also like to thank my colleagues at CWC for all the help and insightful conversations that helped a lot with my work. I would like to thank my friend, band mate, and office colleague, Markus Leinonen for very helpful conversations on the mathematical aspects on various subjects throughout the years (“Brothers in Arms”). I would also like to thank my University of Oulu Graduate School (UniOGS) follow-up group, Adjunct Professor Dr. Harri Saarnisaari (head of my follow-up group), Adjunct Professor Dr. Marian Codreanu, and Dr. Juha-Pekka Mäkelä for their contributions and comments that helped me to achieve the doctoral degree in the recommended four–year time period. A special thanks goes to Associate Professor Dr. Kenta Umebayashi and Professor Yasuo Suzuki from Tokyo University of Agriculture and Technology, who hosted me for three months during 2013 in their laboratory (Suzuki Lab.).
I would like to thank all the funding bodies throughout my career at CWC. My first job was to do my Master’s thesis in the Cognitive Radio Trial Environment (CORE) project, led by Dr. Saarnisaari. On that, I would like to thank my Master’s thesis supervisors, Professor Matti Latva-aho and Dr. Lehtomäki. Professor Latva-aho along with Dr. Jouko Leinonen are the ones to thank for believing in me and giving me a job on the TIVIT Internet of Things (TIVIT IoT) project after my graduation to work on moving relay nodes in the Long Term Evolution (LTE) systems. Related to this work, I would like to thank my friends Petri Luoto and Simon Scott for their help with “The Simulator”. This step of my career was crucial in the process of moving to conduct my doctoral training under Dr. Lehtomäki in the Small and Nanoscale Networks (SAND) project. Later on, I got a three–year position and funding from Infotech Oulu Doctoral Programme, for which I am particularly thankful as it gave me a great deal of stability by securing most of the required funding for my doctoral training phase. Along with Infotech Oulu Doctoral Programme, I would also like to thank Professor Juntti and his Academy of Finland–project, Sensing, Compression, Communications and Data Fusion in Wireless Sensor Networks (SeCoFu) for the additional funding for my salary, as well all of the necessary work trips. I would also like to thank UniOGS and Infotech Oulu Doctoral Programme for the travel grants that allowed me to see the world from Japan to Canada as a doctoral student. Last, but not least, I would like to thank the following foundations for their financial support in the form of personal grants throughout the years: Riitta ja Jorma J. Takasen säätiö, Walter Ahlströmin säätiö, Tauno Tönningin säätiö, Ulla Tuomisen säätiö, and the Nokia Foundation.

I would also like to thank my family and especially my parents, Arvi and Tiina Kokkoneni, for all the support, financial and other support, throughout my Master’s degree studies, as well as my doctoral training during the years I have been living here in Oulu. My father is the reason I am in the communications engineering field, and this choice has been a good one for me. Finally, I would like to thank all my friends, colleagues throughout the world, and everyone else who has contributed to my life during the past decade.

I shall begin this doctoral thesis with the words from Albert Einstein:

“The value of a college education is not the learning of many facts but the training of the mind to think”

Oulu January 24th, 2017

Joonas Kokkoneni
### Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>2D</td>
<td>two-dimensional</td>
</tr>
<tr>
<td>3D</td>
<td>three-dimensional</td>
</tr>
<tr>
<td>ACM</td>
<td>Association for Computing Machinery</td>
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<tr>
<td>BEP</td>
<td>bit error probability</td>
</tr>
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<td>CIA</td>
<td>collision induced absorption</td>
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<tr>
<td>EM</td>
<td>electromagnetic</td>
</tr>
<tr>
<td>FEPA</td>
<td>Federation of the European Producers of Abrasives</td>
</tr>
<tr>
<td>FSPL</td>
<td>free space path loss</td>
</tr>
<tr>
<td>GEISA</td>
<td>gestion et etude des informations spectroscopiques atmosphériques</td>
</tr>
<tr>
<td>GEV</td>
<td>generalized extreme value (distribution)</td>
</tr>
<tr>
<td>HITRAN</td>
<td>high-resolution transmission molecular absorption database</td>
</tr>
<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>IoT</td>
<td>internet of things</td>
</tr>
<tr>
<td>ISI</td>
<td>inter-symbol interference</td>
</tr>
<tr>
<td>ISO</td>
<td>International Organization for Standardization</td>
</tr>
<tr>
<td>JPL</td>
<td>Jet Propulsion Laboratory</td>
</tr>
<tr>
<td>LOS</td>
<td>line-of-sight</td>
</tr>
<tr>
<td>LTE</td>
<td>local thermodynamic equilibrium</td>
</tr>
<tr>
<td>MAC</td>
<td>medium access control</td>
</tr>
<tr>
<td>MIMO</td>
<td>multiple-input multiple-output</td>
</tr>
<tr>
<td>NLOS</td>
<td>non-line-of-sight</td>
</tr>
<tr>
<td>OOK</td>
<td>on-off keying</td>
</tr>
<tr>
<td>PGFL</td>
<td>probability generating functional</td>
</tr>
<tr>
<td>PPP</td>
<td>Poisson point process</td>
</tr>
<tr>
<td>PSD</td>
<td>probability density function</td>
</tr>
<tr>
<td>rms</td>
<td>root mean square</td>
</tr>
<tr>
<td>RTE</td>
<td>radiative transfer equation</td>
</tr>
<tr>
<td>Rx</td>
<td>receiver</td>
</tr>
<tr>
<td>SINR</td>
<td>signal-to-interference-plus-noise ratio</td>
</tr>
<tr>
<td>SIR</td>
<td>signal-to-interference ratio</td>
</tr>
<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
</tr>
</tbody>
</table>
THz-TDS  terahertz time domain spectroscopy
Tx      transmitter
UWB     ultra-wideband
UHF     ultra-high frequency
WSSUS   wide sense stationary uncorrelated scattering

|·|     absolute value
\Gamma(·) gamma function
\delta(·) Dirac delta function
\Theta(·) unit step function
\mathbb{E}(·) expected value
\mathcal{F}(·) Fourier transform
\mathcal{L}(·) Laplace transform
P(·)   probability
\mathbb{R}^d d-dimensional coordinate space
\text{var}(·) variance

\alpha  path loss exponent
\beta   polarizability of a molecule
\beta_I SIR threshold for outage calculations
\gamma  optical depth
\gamma_\infty optical depth due to extraterrestrial objects
\Delta  difference
\epsilon  time delay
\epsilon_m  mean delay spread
\epsilon_{rms} rms delay spread
\epsilon_s time delay through a sample
\zeta  depolarization ratio
\zeta  a set of interfering nodes
\eta  number of obstacles per unit distance
\theta  receiver angle/azimuth angle (ambiguous)
\theta_1 elevation for the incident radiation
\theta_2 elevation for the scattered radiation
\theta_3 azimuth for the scattered radiation
\theta_{Rx} angles from the origin to the nodes
$\vartheta_{Tx}$ angles to the origin from the nodes
\begin{itemize}
    \item $\kappa_a$ absorption coefficient
    \item $\kappa_{ca}$ continuum absorption coefficient
    \item $\kappa_e$ extinction coefficient
    \item $\kappa_{la}$ line contribution coefficient
    \item $\kappa_{others}$ loss coefficient for other possible loss mechanisms
    \item $\kappa_p$ penetration coefficient
    \item $\kappa_s$ scattering coefficient
    \item $\chi$ fractional amplitude of the specularly reflected component
    \item $\lambda$ wavelength
    \item $\Lambda$ (network) intensity function
    \item $\mu_n$ mean value of the non-central chi-squared distribution
    \item $\mu_s$ mean value of the chi-squared distribution
    \item $\nu$ Fresnel parameter
    \item $\varnothing$ albedo of scattering particle
    \item $\rho$ density of the air
    \item $\rho_n$ density of the network
    \item $\rho_{H_2O}$ relative humidity
    \item $\rho_i$ volume mixing ratio of substance $i$
    \item $\sigma_a$ absorption cross section
    \item $\sigma_s$ scattering cross section
    \item $\sigma_g$ geometric standard deviation of the scattering particles
    \item $\sigma_h$ standard deviation for the surface height variations
    \item $\sigma_{sp}$ standard deviation of the scattering particles
    \item $\tau$ transmittance
    \item $\upsilon$ azimuth/elevation angle(s) of the antenna gain pattern
    \item $\Upsilon$ SNR/SINR/non-centrality parameter
    \item $\phi$ receiver elevation angle (ambiguous)
    \item $\Phi$ elevation angle for antenna pattern
    \item $\Psi$ azimuth angle for antenna pattern
    \item $\omega$ angular frequency
    \item $\Omega$ probability generating functional
\end{itemize}
\begin{itemize}
    \item $a_{slit}$ slit width
    \item $A$ area element of the network/surface
\end{itemize}
\( A_{21} \) Einstein coefficient for spontaneous emission
\( A_{\text{FSPL}} \) free space path loss
\( A_{\text{LOS}} \) LOS path loss
\( A_{\text{mol}} \) molecular absorption loss
\( A_{\text{sca}} \) scattering loss
\( B \) Planck function
\( B_{12} \) Einstein coefficient for absorption
\( B_{21} \) Einstein coefficient for stimulated emission
\( B_c \) coherence bandwidth
\( c \) speed of light
\( c_{d_d} \) volume of a unit sphere in drop space
\( c_{d_s} \) volume of a unit sphere in propagation space
\( C_{p,m} \) isobaric mass heat capacity
\( C_{p,V} \) isobaric volumetric heat capacity
\( C(v), S(v) \) Fresnel integrals
\( d \) distance
\( d_1 \) length from Tx to sample
\( d_2 \) length from sample to Rx
\( d_d \) dimension of the drop space
\( d_s \) dimension of the (propagation) space
\( d_{\text{sample}} \) sample thickness
\( d_{\text{slit}} \) slit separation
\( D \) depolarization term
\( E \) energy (ambiguous)
\( f \) frequency
\( f_i \) interference probability density function
\( F_{\text{m}} \) molecular noise acceptance factor
\( F^i \) the spectral line shape of \( i \)th gas
\( g \) Rayleigh roughness factor
\( g_1 \) degeneracy of state 1
\( g_2 \) degeneracy of state 2
\( G_{2D} \) 2D antenna gain pattern
\( G_{3D} \) 3D antenna gain pattern
\( G_a \) antenna gain pattern
\( G_i \) the antenna gain compared to the isotropic antenna gain
$G_{Rx}$: Rx antenna gain

$G_{Tx}$: Tx antenna gain

$h$: Planck constant

$h_e$: height of the edge

$h(t)$: impulse response

$H_0$: hypothesis for receiving bit 0

$H_1$: hypothesis for receiving bit 1

$H_c$: measured LOS channel transfer function

$H(f)$: channel transfer function

$H_{sc}$: channel transfer function of the sample

$I$: intensity/energy flux/energy density/power (ambiguous)

$\bar{I}$: mean intensity/energy flux/energy density/power

$I_{aggr}$: aggregated interference power

$J_n$: energy density

$k$: angular wavenumber

$k_B$: Boltzmann constant

$l_a$: absorption length

$l_c$: surface correlation length

$l(r)$: path loss at distance $r$

$l_s$: scattering length

$l_x$: edge length in x-axis direction

$l_y$: edge length in y-axis direction

$L$: path gain function

$L_{dB}(m,n)$: additional loss due to non-ideal knife-edges

$M$: number of time/frequency domain samples

$n$: refractive index

$n_1$: lower energy level population in non-LTE

$n_2$: upper energy level population in non-LTE

$\pi_1$: lower energy level population in LTE

$\pi_2$: upper energy level population in LTE

$N$: total number of molecules/particles/nodes

$N^i$: number of quantity $i$

$N_d(x_d)$: number density distribution of the scattering particles

$N_x$: the noise power of source $x$

$p$: probability to transmit
\begin{itemize}
  \item \( p_s \) success probability
  \item \( p(t) \) signal power waveform
  \item \( p'_w \) water vapor saturated partial pressure
  \item \( P \) pressure
  \item \( P_{Rx} \) received power
  \item \( P_{Tx} \) transmitted power
  \item \( q \) size parameter for the scattering particles
  \item \( Q_{fa} \) probability of false alarm
  \item \( Q_{md} \) probability of missed detection
  \item \( r \) distance between Tx and Rx
  \item \( r_g \) mean geometric diameter of the scattering particles
  \item \( r_{ke} \) separation of two knife-edges
  \item \( r_{sp} \) mean diameter of the scattering particles
  \item \( R_{21} \) the population decay rate
  \item \( R \) total distance traveled by a photon
  \item \( s \) location of a point in space
  \item \( S(f) \) signal power in frequency domain
  \item \( S_c \) measured LOS received power
  \item \( S'_i \) the spectral line intensity of \( i \)th gas
  \item \( S_{sc} \) measured received power through a sample
  \item \( t \) time
  \item \( T \) temperature
  \item \( T_o \) temperature caused by the extraterrestrial objects
  \item \( T_a \) antenna temperature
  \item \( T_A \) (effective) temperature of the atmosphere
  \item \( T_i \) integration time
  \item \( T_p \) pulse length
  \item \( T_t \) test pulse length
  \item \( V_T \) decision threshold in energy detection
  \item \( W \) bandwidth
  \item \( x \) LOS path length in diffraction
  \item \( x_d \) diameter of a scattering particle
  \item \( x(t) \) transmitted time domain signal
  \item \( X(f) \) transmitted pulse of the measurement device
  \item \( y(t) \) received time domain signal
\end{itemize}
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1 Introduction

Increasing interest in larger data rates has made the higher frequency bands very appealing platforms for realizing future communication systems. The millimetre wave band (30–300 GHz) will be implemented as a part of the 5G small cell applications [1, 2]. This will bring a large improvement in the link capacity over the exhausted ultrahigh frequency band (UHF, 300 MHz – 3 GHz). The reason is the increased spectrum resources offered by the higher bandwidths available at the higher frequencies. The target peak data rates for 5G are at 10 gigabits per second (Gbps) [3]. The most convenient way to reach this target is to utilize the millimetre frequencies.

When aiming for even higher data rates, such as one terabit per second (Tbps) links, the most logical choice is to go further up in the frequencies to the THz frequency band (0.1–10 THz) [4]. These frequencies have been considered for a variety of applications, such as for imaging, where it has good potentiality, e.g., in material recognition and in a non-invasive scanning of goods and humans for contraband [5, 6]. Communications in the THz band has been considered for numerous applications, ranging from nano-scale device-to-device communications (nano-communications) [7–9] to indoor communications and many more [4]. A common feature of the THz applications is the short communication range. The THz frequencies are mostly usable at very short range communication (<10 m) because of the large path loss in the channel and the loss due to inherently small antenna apertures.

The most interesting feature of the THz band is the superior data rate it theoretically offers [10–12]. The theoretical peak data rates can exceed the above mentioned 1 Tbps due the very large available bandwidth of approximately 10,000 GHz. A limiting factor is molecular absorption, which causes frequency selective, but deterministic loss to the signals [10]. The large molecular absorption loss along with the large free space path loss (FSPL) limits the maximum communication distances to a few meters. The molecular absorption can be estimated with the help of databases, such as the high-resolution transmission molecular absorption database (HITRAN) [13, 14]. There are multitude of papers about the theoretical channel models on line-of-sight (LOS) paths, e.g., [10, 15–18], and measurements as well, e.g., [19, 20]. Therefore, molecular absorption is not deeply covered in this thesis. Because of the large losses

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2 The definition of the THz band varies in literature. Some define it as 0.3–3 THz or 0.3–10 THz. However, one very common definition is the above 0.1–10 THz, which will be used in this thesis.
in the channel, directional antennas have been proposed and studied for THz band communications [21, 22]. This is a natural solution for high loss links, especially, in the THz band, where any practical solution will require highly directional links to mitigate the losses. However, in this thesis, the use of isotropic antennas is often assumed to keep the discussion at a general level. The directionality of the links can be taken into consideration by applying proper gain factors, as it will be discussed in this thesis in the case of stochastic models for interference in the THz networks.

Nano-communications is the framework application of this thesis and it refers to communication problems between very small devices called nanodevices. These are envisioned to be small devices sized just a few cubic micrometres capable of simple sensing and actuation tasks [7, 8]. The potential applications for these very simple devices include health care, the environmental, and toxic material sensing and monitoring [8], as well as various other internet of things (IoT) applications [23, 24]. The latter could be established as an autonomous sensor network that periodically report the sensing information to the higher–level devices, e.g., health monitoring systems formed of a large number of nanodevices on humans reporting the health information to the healthcare provider [24].

The size of the devices sets limits on the transmitting power, memory, energy consumption, etc. This limits communication distance as well as data rates. In the case of very small devices, this most likely means the devices will have to harvest their energy from their environments. There are two major research directions in nano-communications; molecular and electromagnetic (EM) communications. Molecular communications considers nanodevice-to-nanodevice information exchange via modified bacteria, or pheromones [7, 25, 26]. Molecular communications suffers from low data rates (up to hours per bit) [27], but offers the possibility to fabricate bio-compatible devices. On the other hand, traditional EM communications offers larger data rates and the possibility to utilize well known transmission schemes for nano-communications. The EM approach suffers from a lack of electronic components required to realize the devices. However, research on the components and novel materials is progressing and has therefore made nanodevices feasible within the next couple of decades.

Advances in materials research have opened possibilities to fabricate THz capable components and have made the THz band an interesting choice for many future applications. The THz band has also been suggested as a communication medium for nanodevices [8]. Some work has been done on the lower frequencies as well, such as [28]. The reason for the interest in high frequencies for nanodevices is due to the
small components and especially the antennas (∼100–1000 nm), which would force the centre frequencies to be very high with conventional materials [10]. However, with novel materials, such as graphene, the centre frequencies can be decreased to the THz frequencies [29]. Because of its unique properties, graphene has been proposed for many components in THz transceivers [30], such as antennas [29, 31–33], and modulators [34].

To date, many solutions have been proposed for THz communications and nano-communications, e.g., receiver architectures [35], energy harvesting [36, 37], routing [38], medium access control (MAC) protocols [39–41], and modulations [42] (mainly on-off keying (OOK) and its variants because of the simplicity of the devices), just to mention a few. Thus, advances in materials, as well as research on the components and the transmission techniques instills some confidence that nanodevices will be implemented in the future and the research on nano-communications will be fully utilized. Still, most of the research is equally valid for the macroscopic research, such as indoor communications, which is also a very interesting subject for near future applications.

This thesis focuses on theoretical and empirical channel modelling as well as on the network level interference modelling. The topics include transmission induced noise modelling; scattering loss modelling on small particles; penetration, surface scattering and reflection, and diffraction measurements; and aggregate interference modelling in the dense THz band networks.

1.1 Theory driven research on THz band propagation

1.1.1 Molecular absorption and transmission induced noise

Molecular absorption is the main loss component in the THz frequencies not present in the lower frequency bands. It is an essential element of theoretical channel models and a core component of all the THz band channel models because molecular absorption causes losses already at very short distances. Molecular absorption is caused by energy absorption at the resonance frequencies of the molecules. The THz frequencies are high enough to have plenty of these resonance frequencies and they cause frequency selective fading to the signals.

With the conservation of energy, molecular absorption of the transmitted signals has been predicted to add an additional noise source to the THz band [10] signals. This is a
very interesting idea because it introduces a new type of noise to communication systems, a self-induced noise, henceforth referred to as a transmission induced noise. It was first introduced in [10] and it has been used in multiple papers, such as [12, 38, 42–44].

The existing transmission induced noise models are based on so called sky noise [45–50], and to be more specific, the molecular absorption noise model included in the sky noise models. It is a simple model of the absorbed energy transfer in the atmosphere and it is present in all absorbing atmospheres. It is based on the Kirchhoff’s law of thermal radiation [51]. Sky noise essentially describes the noise coming from the sky and it consists of energy emitted by the atmosphere on the absorption frequencies and the energy from cosmic sources. In the larger context sky noise is a part of the radio noise picked up by the receivers [50]. Other sources include, e.g., ground noise (the Earth’s or object’s surface emitting energy based on the emissivity of the surface) and man-made noise. In general, the molecular absorption noise of the atmosphere is used as a background noise here. This component is caused by the temperature and the emissivity of the atmosphere.

The difference between molecular absorption noise and transmission induced noise is that the molecular absorption noise is caused by the temperature of the atmosphere, thus, ultimately by Sun (and the Earth). Transmission induced noise, on the other hand, is caused by the transmissions on the radio channel [10].

1.1.2 Small particle scattering and multiple scattering

Besides the molecular absorption related phenomena, the small particle scattering caused by aerosols in the air is modelled and analysed in Chapter 2. Aerosols are small solid or liquid particles suspended in the air by the air molecules [52]. A typical aerosol particle is a water droplet, dust particle or smoke particle. Scattering on small particles, i.e., aerosols in the atmosphere, affects the propagation of EM waves. Rayleigh and Mie scattering describe the scattering processes on small spherical objects, i.e., scattering particle diameters are smaller (Rayleigh theory) or approximately equal (Mie theory) to the wavelength of EM wave [53–55]. Scattering on large objects is described by specular scattering, or geometric scattering [56].

Small particle scattering causes additional loss to the LOS component of the channel because a part of the LOS energy is redirected by the scattering. The utilization of the Rayleigh theories for particle scattering in the THz band was suggested in [8], but was not analysed. The scattering does not only reduce the LOS component similarly to
molecular absorption, but it also enables the signal to reach the receiver after multiple scattering events, corresponding to non-LOS (NLOS) propagation. Multiple scattering is referred to as a sequence of one or more scattering events. There exist two generic channel models for multiple scattering [57, 58]. However, they assume no frequency dependence on scattering or absorption and they cannot be directly applied to modelling the THz band, wherein significant variations occur within the bandwidth of typical pulse shapes proposed for nanodevice communications [8, 11].

1.2 Measurement driven research on the THz band propagation

1.2.1 Measurements on penetration loss

There are a large number of papers about THz penetration, e.g., [59–61]. However, it is very difficult to find suitable papers for a communication specific discussion because most of the papers focus on imaging applications. Still, many papers report good penetration properties for plastics. For instance, [61] provides measurement results on a variety of different plastics that showed good penetration properties. On the other hand, metals do not allow penetration of THz radiation. Otherwise more measurement results are required, especially, for communication specific measurements.

In addition to the penetration loss, different materials introduce different delays dependent on the thickness and characteristics of the materials. The time delay depends on the refractive index of a material. Since THz band pulses are very short, an additional delay on top of the transmission line delay may influence the detection of the transmitted signals. This may be a problem if the symbol separation is small, potentially causing inter-symbol interference (ISI), or if the detection window is short in synchronized communications.

1.2.2 Measurements on reflections and scattering on surfaces

The reflection and scattering related papers are the most popular among the THz band NLOS phenomena related studies in the literature. For instance, [62–69] study the reflections and scattering either from the theoretical, or measurement point of view. Especially, papers [62, 67] focus on rough surface scattering measurements, which are the main interest of the measurements herein. These papers describe measurements at frequencies from 100 GHz to one THz, with the main focus around 300–350 GHz.
These studies have shown that communications through the NLOS path is feasible in the THz band.

More measurement studies on THz surface reflections and diffuse scattering are required in order to serve the future network simulation models for this band. These models require wideband information on the material characteristics. The previous studies have shown that the specular component has the most energy, since it is the directly reflected component of the radiation. Rough surfaces, on the other hand, also provide a diffuse scattering field, i.e., the radiation is partially reflected in random directions from the surface. The diffuse scattering field is rather weak, and the loss through the diffuse scattered paths is tolerable to allow communication over short distances.

1.2.3 Measurements on diffraction effects

Diffraction occurs on the edges and slits in the communication medium. The edges of the objects effectively operate as secondary sources for the radiation, allowing the radiation to reach the shadow regions of the objects. The knife-edge effect has been studied previously in [69, 70] for the lower THz band (0.30–0.31 THz). Those studies showed good agreement in the measurement data to the corresponding theories. However, slit diffraction is a less investigated subject in the communication specific research. The reason is that knife-edge diffraction is far more common in real transmission paths. Diffraction theories [69–74] in general are very well known and can be utilized as they are estimating the signal strengths in the shadows of the obstructions in the communication channel.

1.3 Stochastic geometry for dense THz networks

THz band networks have been studied in a variety of papers [11, 15, 75, 76]. Because of the large path loss and absorption, the theoretical network densities can be very high [11]. The number of users is assumed to vary from tens to several hundred per square meter. Therefore, the network is assumed to be interference limited. The transmission induced noise could theoretically make the dense THz band networks also noise-dependent. However, this noise level is assumed to be small in comparison to the actual transmission powers. The assumption is also supported by the transmission induced noise level study in this thesis, in which it will be shown that this noise is very weak if the temporal
distribution of the noise is considered, or the absorbed energy is considered as being transformed into heat.

Stochastic geometry is a powerful tool for network level problems. It can be used instead of heavy network simulations. Stochastic geometry has been studied in large numbers of papers, e.g., [77–88], with landmark papers by Haenggi et al. (e.g., [77–79]).

The core idea of stochastic geometry is to model transmitter-receiver pairs in a network as a point process(es) [89]. The most common choice is to assume that the nodes are Poisson distributed because of the easy handling of the Poisson point process (PPP). It has been shown over and over again that stochastic geometry is very useful when considering simple enough networks. On the other hand, more complex, and heterogenous networks can be modelled by superimposing multiple PPPs.

There are few works on the THz band mean interference and signal-to-interference-plus-noise ratio (SINR) [11, 15, 75]. Jornet et al. [11, 15] showed results on the stochastic interference power based on the knowledge of the number of users and the probability of the transmissions through symbol length and separation. On the other hand, the work in [75] utilized stochastic geometry. They focused on the Mattern process, which assumes a guard band around the transmitters, ensuring minimum distance between the nodes.

1.4 Scope and objectives of the thesis

This thesis mostly focuses on the THz band channel modelling and a theoretical analysis of a communications network in that frequency range. The channel modelling includes theoretical modelling of the molecular absorption, transmission induced noise, and small particle scattering. Empirical measurements are made for penetration, reflections and scattering on surfaces, and diffraction effects. Fig. 1 shows the considered channel propagation phenomena. Although the framework of this thesis is nano-communications, the channel propagation features are universally valid for any THz band application. As the last part of this thesis, stochastic geometry is used to estimate the mean interference in THz band networks.

One of the main objectives of the theoretical channel modelling is to derive models for the transmission induced noise. This noise component is not very well known and will be considered from multiple different perspectives depending on the possible induction mechanisms behind it. Furthermore, the second objective of the theoretical channel modelling is to derive scattering loss models for small particle scattering.
Subsequently, these can be used to derive wideband multiple scattering models to model the time-delayed signal components due to the scattering. As an application of multiple
scattering, and in general for any THz link, energy detection performance is considered in various atmospheric conditions.

The measurements have a clear objective: to introduce wideband measurements on various NLOS phenomena in the THz band. As mentioned above, these include penetration, reflections and scattering on surfaces, and diffraction effects. The objective herein is to show the feasibility of NLOS communications in the THz band. The measurements are done with terahertz time domain spectroscopy (THz-TDS). This method exploits the large bandwidth of short pulses and as the name suggests, the detection occurs in the time domain. The utilized measurement device, a TeraView TeraPulse 4000, is capable of measuring frequencies from approximately 100 GHz to 4.5 THz in free space, and approximately up to 2 THz in NLOS conditions due to losses cutting the higher frequencies out.

The goal is to analyse the THz band penetration from the communications engineering perspective. The measurements and the subsequent analysis focus on the loss and delay properties of different materials commonly found in indoor environments, i.e., plastic, glass, paper, and hardboard.

The surface reflection and scattering measurements are done for two different cases of scattering materials. One set of samples is composed of sandpapers of different roughnesses. The sandpapers have controlled grit size and are therefore a good choice for studying the scattering effects of variable surface roughness. The second set was composed of various materials also used in the case of penetration measurements; aluminium, glass, plastic, hardboard, and concrete. These provided some insights into the behaviour of realistic indoor materials, which is interesting from the viewpoint of ray tracing simulation models for the THz band.

For the diffraction measurements, two cases are studied: knife-edge diffraction and slit diffraction. Knife-edge diffraction is a very practical problem because of objects in the propagation path may shadow the receiver from the transmitter. Slit diffraction is rarer in the communication channels, but still possible. A slit may be caused, e.g., by a partially open door, etc. The theoretical models for the both cases are given and compared against the measurement data.

The last part of this thesis focuses on the mean interference in dense THz networks. A reasonable assumption of the interference limited network in the presence of large number of nodes enables the use of the main tool to model the mean interference level: stochastic geometry. Using this tool, the objective is to derive models for aggregate interference and outage probability.
Channel power loss is usually assumed to follow simple power law, i.e., \( l(r) = r^{-\alpha} \), where \( l(r) \) is the path loss at distance \( r \), and \( \alpha \) is the path loss exponent [77, 78]. This path loss model provides closed form solutions for the aggregate interference in the network provided that the path loss exponent is large enough [77]. This is not the case here, since the traditional free space path loss model is utilized. This causes some problems with closed form solutions, as it will be shown in Section 4.2.2. Furthermore, the channel is usually assumed to be Rayleigh fading, which provides very useful simplifications, as, for instance, the outage probability can be easily calculated from the closed form solution of the aggregate interference [77].

Because the THz band random networks are modelled with the usual THz channel models, i.e., the molecular absorption and general free space loss, the existing closed form solutions for aggregate interference are no longer applicable. Therefore, all the results are validated with computer simulations.

Putting the whole thesis together, the ultimate objective herein is to increase the knowledge of the channel behaviour of THz band point-to-point links. These are complemented by energy detection and a stochastic network analysis. These issues are critical with respect to upcoming THz band applications. The channel needs to be known before any real–life solutions can be implemented. The work on channel propagation modelling extends beyond this thesis since the THz band is a very new research subject. Thus, more channel studies on different scenarios and materials are required. The ones given here and in various other works around the world provide a good background on the theoretical and empirical modelling for the future research.

1.5 Contributions and outline of the thesis

Besides the basic channel propagation phenomena in the THz band, Chapter 2 focuses on the transmission induced noise and scattering on small particles, or aerosols. This chapter is based on articles and conference papers: [90–93].

Journal article [90] focuses on transmission induced noise. It expands the existing research on this new noise component by introducing several propagation phenomena not previously considered in the context of communication oriented papers. It also shows that transmission induced noise is most likely significantly lower than expected based on the existing research. This is supported by the fact that this noise component has not been measured, whereas the existing research suggests the transmission induced noise
level should be measurable. However, the true noise contribution remains unknown because this noise component has not been measured.

The articles [91–93] focus on aerosol scattering, multiple scattering and energy detection. The LOS scattering model is the first one derived for the THz band. Similarly, the wideband multiple scattering model is new, although, based on the narrow band models derived in [57, 58].

The measurements results are presented in Chapter 3 and were first introduced in [94–98]. The penetration studies are familiar from THz imaging papers. However, the analysis are mostly unsuitable for communication system specific analysis. Because of this, the work herein analyses the frequency domain penetration loss, as well as delay properties of the penetration.

Surface scattering/reflection analysis has been previously done in numerous papers. These mostly focus on the frequencies around 300 GHz, which makes the measurements here very important as the results are given on extremely wide band from approximately 100 GHz to 2 THz. Therefore, they provide an important contribution to the wideband channel behaviour.

The diffraction measurements are very rare in the THz band and slit diffraction especially has not been analysed previously. As in the case of surface scattering measurements, the main novelty comes from the large bandwidth and analysis with respect to the needs of the communications engineering community. The diffraction measurements are compared against the theoretical models and it is shown that the measurements agree with the models very well.

Chapter 4 focuses on the aggregate interference and its applications. This section is based on a submitted journal paper [99] and it utilizes stochastic geometry to estimate the mean interference level and outage in THz band networks. Compared with the other works on the THz band, the approach here gives a far more general derivation of the interference compared to the previous work on the THz band due to utilization of the PPP assumption. The derivation follows closely the landmark papers on the subject, but the difference comes from the channel models, which incorporate molecular absorption in the free space path loss. As it will be shown, the derived models predict the interference levels perfectly.

The general organization of the thesis is as follows. Chapter 2 presents the theoretical works on the channel behaviour, as well as on the basic channel propagation phenomena. These include molecular absorption, transmission induced noise and the scattering loss on small particles in the atmosphere. In Chapter 3, measurements on the THz band
propagation phenomena are given. These include penetration, rough surface scattering, and diffraction. Chapter 4 derives the analytical results of the stochastic geometry for the THz frequencies.

1.6 Author’s contribution

This monograph type thesis is composed of ten original articles [90–99], of which [90, 91, 99] are journal articles ([99] is submitted) and [92–98] are conference papers ([98] is accepted). The author of this thesis had the main role in developing the ideas, deriving the theoretical results, implementing the MATLAB codes, developing and implementing the measurement setups, taking the measurements and doing the subsequent analysis, and writing the papers. The exception is [98], where the measurement setup and analysis was done by the author, but the measurements were conducted by M.Sc. student Perttu Rintanen (the second author of the paper). Other co-authors participated in the developing the ideas for the research subjects and provided help, guidance, and criticism during the research and writing process. The main contributions were given in the previous section.
2 THz band signal propagation: theory

The theoretical aspects of THz band signal propagation are considered in this chapter. This chapter is based on the following papers: [91, 92] [© 2015 IEEE], [93] [© 2015 ACM], and [90] [© 2016 Elsevier]. The most important propagation phenomenon in the THz band is molecular absorption, which is the most interesting and distinguishing feature of the higher frequency bands, compared, e.g., to the UHF channels. Absorption of energy also causes noise, transmission induced noise, which is a new kind of noise for the communication systems. The main focus of this chapter is in the scattering of energy due to the small particles, aerosols, in the atmosphere. Because molecular absorption, as well as multiple scattering have an influence on the time domain behaviour of the received signal, the energy detection performance in the multiple scattering conditions is also considered at the end of this section.

Molecular absorption is briefly presented in Section 2.1 followed by a transmission induced noise discussion in Section 2.2. Frequency domain scattering loss is derived in Section 2.3. Multiple scattering, i.e., the time domain models are discussed in Section 2.4. Energy detection performance in multiple scattering conditions is explored in Section 2.6 and a discussion of the results is presented in Section 2.7.

2.1 Molecular absorption loss

Molecular absorption loss is the main difference between the THz band propagation and the lower frequency bands, such as the UHF band. Molecular absorption is caused by the resonance frequencies of the molecules in the atmosphere. It causes random discrete frequency domain loss. However, because of the large numbers of molecules in a unit volume of the atmosphere, the molecular absorption loss is deterministic in practise. The molecular absorption is ultimately caused by the transmitted EM wave shifting the molecules in the medium to higher energy states. The energy, equivalent to the difference between the higher and the lower energy state of a molecule, determines the absorption energy that is drawn from the EM wave. This has a direct impact on the absorption frequency because the absorbed energy is \( E = hf \), where \( h \) is the Planck constant and \( f \) is frequency [100, Chapter 17]. This process can be described stochastically by using the absorption coefficient \( \kappa_a(f) \). This quantity describes the average effective area of
the molecules per unit volume. Molecular absorption loss can further be divided into two different loss mechanisms, line contribution and continuum absorption.

### 2.1.1 Line contribution

The line contribution refers to the attenuation of EM waves on the resonance frequency lines of the molecules in the communication medium. In this thesis, the line contribution is assumed to be included in molecular absorption, which includes all the mechanisms of molecular absorption. Most of the molecular absorption in the atmosphere is caused by greenhouse gases, such as water vapour, carbon dioxide, oxygen, and ozone [17, 101]. In the THz band, water vapour and oxygen are the major absorbers [17, 102]. The Beer-Lambert law can be used to calculate the fraction of incident EM radiation that propagates through the channel [10, 17, 103]. This fraction is called the transmittance and is defined as [17]

\[
\tau(f, r) = \frac{P_{Rx}(f)}{P_{Tx}(f)} = e^{-\sum_{i} \kappa_{la}^{i}(f)r \sec(\theta)},
\]

where \( f \) is frequency, \( r \) is the distance from the transmitter (Tx) to the receiver (Rx), \( P_{Tx}(f) \) and \( P_{Rx}(f) \) are transmitted and radiated power, respectively, \( \kappa_{la}^{i}(f) \) is the absorption coefficient of the \( i \)th absorbing species at frequency \( f \) and \( \theta \) is angle of the incident wave. For instance, in a plane parallel atmosphere, the distance is effectively increased by the factor \( 1/\cos(\theta) = \sec(\theta) \) as the wave propagates in the vertical direction [102]. In the further calculations, the angle of the incident wave is omitted due to the assumption of a spherically symmetric environment, i.e., the environment is uniform in all directions and the distance \( r \) fully defines the geometry of the environment. Also, the total summed absorption coefficient is denoted as \( \kappa_{la}(f) \), i.e., \( \kappa_{la}(f) \equiv \sum_{i} \kappa_{la}^{i}(f) \).

The absorption coefficient for the line contribution at frequency \( f \) depends on the pressure, temperature and molecular composition of the channel through

\[
\kappa_{la}^{i}(f) = N^{i} \sigma_{la}^{i}(f),
\]

where \( N^{i} \) is the number density and \( \sigma_{la}^{i} \) is the absorption cross section of the \( i \)th absorbing species. The number density for absorbing species \( i \) can be obtained from ideal gas law and by weighting each absorbing species (and their isotopologues) with corresponding fractional abundance \( \rho_{i} \), as presented, e.g., in [10, 17].
The natural abundances for atmospheric molecules (along with other important parameters for line-by-line absorption calculations) can be found, e.g., in HITRAN [13, 14]. There are also other well–known line catalogues, which can be utilized by the line-by-line models, such as Gestion et Etude des Informations Spectroscopiques Atmosphériques (GEISA) database [104] and Jet Propulsion Laboratory (JPL) catalogue [105]. These catalogues give the abundances for dry air. Therefore, the volume mixing ratio of water vapour should be used to scale the abundances of dry air components. The relative humidity is used in all the calculations in this work. The volume mixing ratio of water vapour can be estimated with the Buck equation [106]

\[ p_w^* = 6.1121(1.0007 + 3.46 \times 10^{-6}P) \exp \left( \frac{17.5027}{240.97 + T} \right), \]  

(3)

where \( p_w^* \) is saturated water vapour partial pressure under pressure \( P \) and temperature \( T \). In this equation, \( P \) is given in hectopascals and \( T \) is given in degrees Celsius. The volume mixing ratio of water vapour in relative humidity \( \rho_{H_2O} \) (in percent) is calculated as

\[ \rho_{H_2O} = \frac{\rho_{H_2O}^* p_w^*}{100 P}, \]  

(4)

where \( \rho_{H_2O}^* p_w^*/100 \) is the partial pressure of water vapour. Thus the abundances for moist air components are \( \rho_{air,moist} = \rho_{air,dry}(1 - \rho_{H_2O}) \) (excluding the water vapour).

The absorption cross section \( \sigma_i(f) \) is a product between spectral line intensity \( S_i(T) \) and spectral line shape \( F_i(P, f, T) \), i.e., \( \sigma_i(f) = S_i(T)F_i(P, f, T) \). The absorption cross section describes the effective area for absorption for a single particle. Furthermore, spectral line intensity describes the strength of the spectral lines and spectral line shape tells us the width and shape of the spectral lines. The exact calculations, as well as, the scaling of the line intensity and the line shape can be done as in [10, 17, 107].

2.1.2 Continuum absorption

It has been found in measurements that the total spectral absorption is larger than the total line contribution. The continuum absorption is defined as excess absorption and not counted in line contribution [17, 108, 109]. The spectral shape of continuum absorption is a smooth curve over the frequencies. It is caused, e.g., by collisions of molecules, momentarily changing their properties. However, the explicit reason for continuum absorption is not known, as it composed of all the possible phenomena affecting the absorption of the energy [17, 109].
As in the case of the line contribution, water vapour causes most of the continuum absorption [17, 108]. Some parts of the continuum absorption can be calculated theoretically, but due to uncertainty in the source of continuum, the continuum absorption calculations also rely on experimental data. The continuum absorption coefficients can be calculated, e.g., with the \textit{am} atmospheric model [17]. This model uses the MT\_CKT continuum model for calculation of the water vapour continuum [110]. The MT\_CKT continuum model utilizes theoretical models and experimental data. Since the \textit{am} atmospheric model only accounts for the continuum caused by collisions between two molecules, accounting for the absorption on lines forbidden for the line contribution [17], continuum absorption is henceforth referred to as collision induced absorption (CIA) as in [17]. However, the term continuum absorption is not fully omitted as the CIA is a part of the total continuum spectrum.

The \textit{am} model produces binary absorption coefficients for the CIA, i.e., the collisions between two molecules have been accounted for in the total continuum absorption. However, higher order collisions are omitted. The previously presented absorption coefficient has unit cm$^{-1}$. The unit of the continuum absorption coefficient in the \textit{am} model is cm$^3$, thus the continuum absorption coefficient has to be multiplied twice with the corresponding number densities. For instance, the total H$_2$O continuum absorption coefficient becomes [17]

$$\kappa_{\text{ca}}(f) = N^{\text{H}_2\text{O}} (N^{\text{H}_2\text{O}} \kappa_{\text{ca}}^{\text{H}_2\text{O}}(f) + N^{\text{air}} \kappa_{\text{ca}}^{\text{air}}(f)), \quad (5)$$

where $N^{\text{H}_2\text{O}}$ is number density for water molecules, $N^{\text{air}}$ is number density for dry air molecules, $\kappa_{\text{ca}}^{\text{H}_2\text{O}}(f)$ and $\kappa_{\text{ca}}^{\text{air}}(f)$ are binary continuum absorption coefficients from the \textit{am} model, namely the self and the foreign continuum absorption coefficients for water vapour. The self continuum is caused by the H$_2$O – H$_2$O collisions and the foreign continuum is caused by the intermolecular collisions, with H$_2$O being the other colliding partner. The other continuum absorption coefficients are treated similarly. For more information on the explicit calculation of the continuum absorption coefficient, please see the \textit{am} model documentation [17, Chapter 4].

In the following, notation $\kappa_{\text{a}}(f)$ is used to include the total line contribution $\kappa_{\text{la}}(f)$ and the total CIA $\kappa_{\text{ca}}(f)$ at frequency $f$, i.e.,

$$\kappa_{\text{a}}(f) \equiv \sum_{f} \kappa_{\text{la}}^{f}(f) + \sum_{k} \kappa_{\text{ca}}^{f}(f), \quad (6)$$
where index \( k \) refers to the source of CIA, e.g., \( \text{H}_2\text{O} \) – \( \text{H}_2\text{O} \) collisions. It is shown in Section 2.1.4 that the overall impact of CIA is not significant, but it is included for accurate channel modelling.

2.1.3 Free space path loss

The main attenuation mechanism in static channel models is the isotropic expansion of the radio waves in the medium [10, 111]. If isotropic antennas are assumed, the FSPL is

\[
A_{\text{FSPL}}(f) = \frac{(4\pi f r)^2}{c^2},
\]

(7)

where \( c \) is the speed of light. Then, assuming FSPL and molecular absorption loss, the total loss in the LOS path becomes

\[
A_{\text{LOS}}(f) = A_{\text{FSPL}}(f) A_{\text{mol}}(f),
\]

(8)

where \( A_{\text{mol}}(f) = \exp(\kappa_a r) \) is the inverse of the transmittance due to molecular absorption loss.

2.1.4 Numerical examples of the molecular absorption phenomena

Besides the FSPL, the main attenuation mechanism in the THz band is molecular absorption. These losses (on the dB-scale) are shown in Fig. 2 as a function of frequency and distance and this figure is essentially the same as the one presented in [10]. Due to the focus on nanodevice communications, the results are given at a distance range from 1 to 100 cm. The values in this figure are calculated at: temperature 303 K, pressure 101.3 kPa and relative humidity 100%, corresponding to approximately 4% volume mixing ratio for water vapour at the mentioned pressure and temperature. The calculation is based on (8) with necessary variables from HITRAN database. Therefore, this figure illustrates the output of the existing and common frequency domain channel model [10]. The maximum attenuation has been limited to 120 dB (similarly as in [10]). This way the molecular absorption characteristics are easy to observe because otherwise the attenuation would be extremely high on the lines on which the transmittance approaches zero. This figure shows the high frequency selectivity of the molecular absorption. Roughly speaking, the exponentially degrading features as a function of distance and frequency are caused by FSPL whereas sharp loss features are due to molecular absorption.
Fig. 2. The LOS path loss due to FSPL and molecular absorption in the THz band in the distance range from 1 cm to 100 cm [© 2015 IEEE].

An additional loss is introduced by CIA. Fig. 3 gives this loss (in dB) with the same parameters as in Fig. 2. It can be seen that the CIA induces additional loss to
the received signal. However, the effect is at worst less than 2 dB at a distances below one meter. Similarly to molecular absorption, the CIA is strongly dependent on the amount of water vapour in the atmosphere as the water vapour is the main cause for CIA. Therefore the water vapour plays an important role in the maximum transmission distances for the nanodevices. Although the CIA is small, it is causing additional loss. Thus, with respect to the accurate THz band channel modelling, CIA should be included regardless of the small effect on the nanodevice-to-nanodevice links.

2.2 Transmission induced noise

Transmission induced noise has been predicted to exist because of the conservation of the absorbed energy in the atmosphere. Transmission induced noise introduces a new type of a radiative noise to communication systems in THz band systems. However, the exact behaviour of this noise is still not completely understood.

The current transmission induced noise models are based on the sky noise, which includes contributions from molecular absorption noise and the noise from cosmic sources [50]. The molecular absorption noise is the main point of interest here and it is caused by reradiation of the absorbed energy. That is, the emissivity of the medium at a certain temperature. The emissivity is directly dependent on the level of absorption as it will be shown in the next section. Since the emissivity is an important part of the molecular absorption noise, and thus the sky noise as well, it has been used as a base model in the existing transmission induced noise models. The reason lies behind the similar absorption phenomenon as in the case of molecular absorption. However, the main difference is in the source. Molecular absorption noise is caused by the temperature of the atmosphere (or any propagation medium), whereas transmission induced noise is caused by the signal from the transmitter.

Because the transmission induced noise is most likely to be very weak, its validation using measurements is difficult. This section presents the possible propagation phenomena and other phenomena behind transmission induced noise, including the propagation delay, emission rate of the absorbed energy, and the heat approximation. These do not provide an answer to the total contribution of the transmission induced noise, but they provide a basis for future studies by introducing some phenomena that the molecular absorption noise approach does not take into account.

The radiative transfer equation is very important piece of the transmission induced noise and is presented in Section 2.2.1. Because the molecular absorption noise model
is the basis of the transmission induced noise, it is presented in Section 2.2.2. The propagation delay caused by the limited propagation speed in the medium is derived in Section 2.2.4. The emission rate of the absorbed and thereafter re-emitted radiation is given in Section 2.2.5. If all the absorbed energy is transformed into heat, another approach is used. This is given in Section 2.2.6.

2.2.1 Radiative transfer equation

All the molecular absorption noise models, as well as the existing transmission induced noise model are ultimately based on the radiative transfer equation (RTE). Assuming a constant atmosphere, i.e., the absorption coefficient is constant in the path of the radiation, the RTE can be presented as [53, Section 1.7]

\[
I(d, f) = I(0, f)e^{-\kappa_a(f)d} + \int_0^d B(T, f)\kappa_a(f)e^{-\kappa_a(f)s}ds,
\]

(9)

where \( I(d, f) \) is the intensity of the radiation at distance \( d \), \( I(0, f) \) is the intensity of the transmitted signal and \( B(T, f) \) is the source term, i.e., the Planck function

\[
B(T, f) = \frac{2h\pi f^3}{c^2} \left( e^{\frac{hf}{k_B T}} - 1 \right)^{-1},
\]

(10)

where \( k_B \) is the Boltzmann constant. The above equation is multiplied with \( \pi \) in order to transform the unit of the Planck function from W/Hz/cm\(^2\)/sr to W/Hz/cm\(^2\) [112]. Then (10) corresponds to spectral exitance of the surface of the black body [112]. The first term on the right-hand side of (9) is the transmitted energy from the Tx and the second term is the radiative energy of the atmosphere. In the general case, it is described by the Planck law because the Planck law determines the radiation emitted by an object at temperature \( T \) (black body radiation). The integral in the RTE describes the noise intensity at point \( d \) given the entire atmosphere contributes to the noise at point \( d \). That is, the contribution of all the points \( s \) in the atmosphere are summed. This is clear from the definition of RTE, as the absorption term \( \exp(-\kappa_a(f)s) \) describes the transmittance at the surface of a sphere with radius \( s \).

2.2.2 Molecular absorption noise

Molecular absorption noise is caused by the temperature of the absorbing atmosphere/medium, causing the medium to be an effective black body radiator, or a grey
body radiator in non-homogenously absorbing medium (in frequency domain). The molecular absorption noise is therefore known as a background noise and it is independent of the transmitted signals. This noise component is an important part of the sky noise in the absorbing frequencies. The sky noise also contains cosmic sources, such as cosmic background radiation at approximately 2.7 K [50], but other possible extra-terrestrial sources as well. There are a multitude of papers describing sky noise, such as [45–50]. It is present, e.g., in the Earth stations in satellite communications and it is commonly defined by the antenna noise temperature. Earth’s atmosphere is a dynamic medium, with a roughly speaking decreasing temperature and pressure as a function of elevation. Therefore, the antenna noise temperature can be defined as an integral over the entire atmosphere [46, 48]

\[ T_a(f) = \int_0^m T(s) \kappa_a(s, f) e^{-\gamma(s, f)} ds + T_\infty e^{-\gamma(f)}, \]  

(11)

where \( T_a(f) \) is the antenna noise temperature, \( T(s) \) is the temperature of the atmosphere as a function of the distance \( s \), \( \kappa_a(s, f) \) is a distance dependent absorption coefficient, \( \gamma(s, f) = \int_0^s \kappa_a(s', f) ds' \) is the optical depth of the medium [46, 48] and the term \( T_\infty e^{-\gamma(f)} \) accounts for the radiation coming from outside the atmosphere, e.g., the cosmic background noise at 2.7 K [50]. This last term is henceforth assumed to be small, since at the THz frequencies the molecular absorption noise is quite strong. In the general case, the absorption coefficient is a function of distance as in the above equation. This due to variable temperature and pressure in the atmosphere. If the distance is small, or the atmosphere is otherwise homogenous, \( \kappa_a(s, f) = \kappa_a(f) \). Assuming an isothermal atmosphere, (11) reduces to (neglecting the possible other terrestrial/extra-terrestrial sources) [45, 47–50]

\[ T_a(d, f) = T_A(1 - e^{-\kappa_a(f)d}) \]  

(12)

where \( T_A \) is the temperature of the atmosphere (or the effective temperature of the atmosphere) and \( d \) is the thickness of the atmosphere.

The antenna orientation has an impact on the antenna noise temperature [45, 46, 49, 50]. This due to the thickness of the atmosphere changes with the elevation angle. The effect of the elevation angle has been estimated to be [46]

\[ T_a(d, f, \Theta) = T_A(1 - e^{-\kappa_a(f)csc(\Theta)}) \]  

(13)

where \( csc(\Theta) \) is a cosecant of the elevation angle \( \Theta \) with 90 degree angle at the zenith. This has been estimated to be quite accurate down to elevation angles from 10 to 15
degrees [46]. It can be seen that the antenna noise temperature is highest when the antenna is pointed in a horizontal direction and lowest when pointed at the zenith [45, 46, 50]. This is due to the fact that the length of atmosphere is shortest in the zenith direction.

On top of the sky noise, if the receiver sees the ground through the main lobe or the side lobes, it also has an impact on the total antenna noise temperature. This is dependent on the temperature and emissivity of the ground, and if observed from a space station down to the Earth, also on the reflection coefficient of the atmosphere.

Since the existing models for transmission induced noise are based on the molecular absorption noise, the ground noise and other possible radio noise sources are not considered in the following discussion. Given the antenna temperature, assuming no other losses, the molecular absorption noise can be approximated as [47]

\[ N_{\text{sn}}(T_A, d, f) = k_B T_A W F_a (1 - e^{-\kappa_a(f)d}). \]  

where \( W \) is the bandwidth of the system and \( F_a \) is a molecular absorption noise acceptance factor of the receiver (an aperture term for the noise). In the general case, \( k_B T \) should be replaced by the Planck law (10). However, there is a difference in the unit of these two terms. The unit of \( k_B T \) is J or W/Hz. Integrating it over the bandwidth of the system yields \( W \). The Planck law (10) has a unit of W/Hz/cm\(^2\). Integrating (10) over the bandwidth yields W/cm\(^2\). This is because Planck law is a general radiative function of the surface of the black body. Thus, the antenna aperture term is required if one wants to describe it as power.

The most interesting feature here is the general behaviour of the noise in the absorbing medium. Therefore, the molecular absorption noise will be described by Planck law (10) and RTE (9). Assuming a highly absorbing medium, such as the THz band, the molecular absorption noise reduces to the Planck function

\[ N_{\text{sn}}(T_A, d, f) = \int_0^d B(T_A, f) \kappa_a(f)e^{-\kappa_a(f)d}ds \]

\[ = B(T_A, f)(1 - e^{-\kappa_a(f)d}) \approx B(T_A, f) \]

assuming large \( d \) (the whole atmosphere contributes to the molecular absorption noise). This is a good approximation as it can be shown that the antenna temperature is approximately constant in the THz band (by utilizing large \( d \)). As a consequence, in this approximation, the atmosphere radiates according to Planck law, which is utilized as
background radiation and the transmission induced noise is an additional noise source, caused by the actual transmissions. As stated above, the molecular absorption noise is not caused by the transmissions, but by the atmosphere. Since it is approximated with the Planck function, the unit of the molecular absorption noise becomes W/Hz/cm$^2$. It can be further approximated by taking into account the (ideal) antenna aperture term $c^2/(4\pi f^2)$:

$$N_p(T_A, f) = \frac{c^2}{4\pi f^2} B(T_A, f), \quad (16)$$

where $N_p(T_A, f)$ is the molecular noise power spectral density (PSD) due to the Planck function. Using the antenna aperture, the unit becomes W/Hz. The Planck function without the antenna aperture is plotted in Fig. 4. The antenna aperture was discarded in this figure since most of the transmission induced noise results are given in W/Hz/cm$^2$ due to interest in the pure radiative behaviour of this noise component.

The next three sections consider transmission induced noise. First, the existing transmission induced noise model is derived. Next, the propagation delay of the transmission induced noise is derived assuming instantaneous emission of the absorbed energy in a random direction. Finally, the emission delay due to the spontaneous and the stimulated emissions are considered.

### 2.2.3 General frequency dependent transmission induced noise

Whereas molecular absorption noise is caused by the atmosphere, transmission induced molecular absorption noise requires a different approach. Since a point source (a transmitter) is considered, the integral in the right-hand side of (9) vanishes and the source term (the Planck function) must be replaced by an appropriate transmitting energy function. Furthermore, it is noticed that

$$\kappa_a(f) e^{-\kappa_s(f)r} r = \frac{d}{dr} \left( 1 - e^{-\kappa_s(f)r} \right). \quad (17)$$

This derivative is the transmission induced noise density of a point source with unit transmitting power at distance $r$. This is very easy to verify by calculating the derivative. On the other hand, this can also be verified by looking at the purpose of the derivative in the above equation. The difference in the absorbed energy between two points, infinitesimally close to each other, indicates the energy that was absorbed between these points. There is also another way to think about the correctness of this approach: the fraction $(1 - \exp(-\kappa_s(f)r))$ gives a cumulative fraction of the energy absorbed.
This is clear when the distance is taken to the limit: all the energy is absorbed and
\[ \lim_{r \to \infty} (1 - \exp(-\kappa_s(f) r)) = 1, \] unless the absorption coefficient is zero. The derivative of
a cumulative distribution is a probability density function. Then the above equation
is an exponential random variable with a mean given by an inverse of the absorption
coefficient. This is the point of interest: an energy density function of the absorbed
energy, i.e., the spatial distribution of the absorbed energy. Thus, the transmission
induced noise energy at point \( r \) depends on the derivative of the complement of the
transmittance.

There is an additional proof of the transmission induced noise being proportional to
(17), which gives the fraction of the energy absorbed at the surface of a sphere with
radius \( r \). This is very easy to verify by stating that the total energy \( E(f) \) at frequency
\( f \) is either transmitted a distance \( r \) or is absorbed along the path from zero to \( r \). This

corresponds to a transmittance at distance \( r \) plus the integral over the path of propagation:

\[
E(f) e^{-\kappa_s(f) r} + \int_0^r E(f) \frac{d}{ds} \left( 1 - e^{-\kappa_s(f)s} \right) ds = E(f) e^{-\kappa_s(f) r} + E(f) \left( 1 - e^{-\kappa_s(f) r} \right) = E(f).
\] (18)

As it can be seen, this is valid for all \( r \). This verifies the absorbed energy density at point
\( r \) to be

\[ E_p(r, f) = \frac{P_{tx}(f) \left( 1 - e^{-\kappa_s(f) r} \right)}{4\pi r^2} = \frac{P_{tx}(f) \kappa_s(f) e^{-\kappa_s(f) r}}{4\pi r^2}, \] (19)

where \( P_{tx}(f) \) is a transmitting signal PSD, \( 1/(4\pi r^2) \) accounts for the spreading loss.

From the above expression for the energy density, the existing transmission induced noise models \[10, 42\] can be derived. Without the time dependency, the energy flux through the surface of a sphere with radius \( r \) is obtained by

\[
\int_0^r P_{tx}(f) \frac{d}{ds} \left( 1 - e^{-\kappa_s(f)s} \right) ds = P_{tx}(f) \left( 1 - e^{-\kappa_s(f) r} \right).
\] (20)

By dividing by the area of the sphere, the existing transmission induced noise models for the receiver at distance \( r \) from the transmitter become

\[ N_m(r, f) = \frac{P_{tx}(f)}{4\pi r^2} \left( 1 - e^{-\kappa_s(f) r} \right). \] (21)

There are some variations in the existing models \[10, 12, 38, 42-44\] of how to take into
account the spreading loss and the antenna aperture, but this is the core model. For
instance, taking into account the antenna aperture term in (21) leads to the transmission induced noise models presented in [42, Eq. (16)], [43, Eq. (8)], and [38, Eq. (12)].

An example of this kind of a noise is given in Fig. 4. The transmission induced noise is given for the transmitting power $P_{\text{Tx}} = 2 \text{ pW/Hz}$ and the distance $r = 10 \text{ cm}$. For reference purposes, the Planck function according to (10) at temperature $T = 300 \text{ K}$ is also shown and it corresponds to the molecular absorption noise. It can be seen that with the above assumptions of the noise, i.e., no spatial or time dependencies, the transmission induced noise is a potential source of interference in the THz band. However, because of neglecting the spatial and temporal features, the correctness of the model can be questioned. The spreading loss is a way to limit the noise that would otherwise increase with distance. This, however, also suggest that the transmission induced noise at the receiver is directly transmitted by the transmitter, although it is transmitted by the surrounding medium. Therefore, some other aspects should be considered when modelling the transmission induced noise.

![Fig. 4. The transmission induced noise according to (21). A reference is provided by the Planck function representing the molecular absorption noise [© 2016 Elsevier].](image-url)
2.2.4 Propagation delay for the transmission induced noise

The above approach did not take into account that the absorbed energy has a propagation delay dependent on the absorption properties of the atmosphere. The spatial energy density is obtained from (19) assuming a spherically symmetric space. It can be noticed that the representation of the noise density is the same as the average path length in the multiple scattering models [58, 111]. Therefore, utilization of the multiple scattering channel models is proposed for the transmission induced noise under the assumption of no emission delay. It should be noticed that in reality there is a delay between absorption of the energy and its release. Just for modelling purposes, this delay is assumed to be small. Because of this assumption, the whole propagation process is similar to the scattering of energy. The difference is that the scatterers are not particles but molecules, which do not scatter at THz frequencies. The reason is the overly long wavelength compared to the size of the molecules. Instead, the "scattering" here refers to the release of the absorbed energy to a random direction.

Multiple scattering models [58, 111] are also used in Section 2.3 to model the wideband multiple scattering in the presence of small particles and they will be better reviewed there. Here, the three-dimensional model given by Paasschens in [58] is used due to the interest on emissions on individual frequencies. The multiple scattering theory comprises the LOS and the NLOS parts of the radiation. The NLOS is the interesting part of the theory, because it gives the delayed tail response, or the transmission induced noise. The LOS component is always the desired signal component. The multiple scattering model for the transient response per unit power is given by [58]

\[
I(r,t,f) = \sum_{N^c=1}^{\infty} \kappa_a(f)^3 e^{-\kappa_a(f)ct} \frac{\Gamma\left(\frac{3}{2}N^c + \frac{3}{2}\right)}{\pi^{3/2} \Gamma\left(\frac{1}{2}N^c\right)} \times \left(1 - \frac{r^2}{c^2t^2}\right) \frac{1}{N^c-3} \Theta(r-ct),
\]

where \( I(r,t,f) \) is the energy flux per hertz in units of W/Hz/cm\(^2\) (J/s/Hz/cm\(^2\)) at the receiver at the distance \( r \) from the isotropic transmitter, \( N^c \) is the number of collisions, i.e., the absorptions before a photon reaches the receiver, \( \Gamma(\cdot) \) is the gamma function and \( \Theta(x) \) is a step function, which one for \( x > 0 \) and zero elsewhere. The transmitting power spectral density is required, since the theory in [58] is for unit transmitting power.

Figure 5 shows the behaviour of the intensity according to (22). An impulse-like signal in the time domain with the power spectral density \( P_{Tx}(f) = 2 \) pW/Hz is assumed.

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The distance between Tx and Rx is assumed to be $r = 10 \text{ cm}$. It can be seen that the transient caused by the continuous absorption-emission process is strongly frequency dependent and locally rather long. This is due to differences in the absorption properties from frequency to frequency. This causes variations in the effective path length before the next absorption occurs, which is given by an inverse of the absorption coefficient. It can further be seen that the noise level is quite small compared to the one in Fig. 4. The main reason is because the noise energy is distributed over time in contrast to the case in Fig. 4. For these reasons, the contribution of the transmission induced noise is small compared with the molecular absorption noise due to the temperature of the atmosphere.

### 2.2.5 Emission rate in THz band

In this section, the purely radiative properties of the radiative decay of the energy through the spontaneous and the stimulated emissions are considered. The absorbed energy is assumed to remain at the absorption frequencies and the emissions are only caused by the radiative effects.
The problem of the absorbed and thereafter re-emitted energy requires a non-local thermal equilibrium (non-LTE) assumption since the LTE condition is broken by the absorbed energy from the transmission. This phenomenon is purely related to the transmission induced noise, theoretically locally disturbing the thermal equilibrium. There are two good books explaining non-LTE phenomena [113, 114] with a similar approach to the problem. The LTE assumption cannot be fully forgotten, since the full non-LTE problems may be very demanding.

The two-level absorption/emission model can be used to describe energy transitions in the absorption process, which can be explained through Einstein coefficients. The idea behind the two-level model is that each energy transition is seen as an energy transition between lower and upper energy states. There are a total of $N$ molecules per unit volume of the medium. As a consequence, there are $N$ possible energy transitions in the unit volume per frequency. These $N$ energy transitions populate either the upper or the lower energy state of each molecule at frequency $f$. Knowing the Einstein coefficients, the time dependent energy state populations can be derived as it is done in Appendix 1:

\begin{align}
  n_2(t, f) &= \pi_2(f) + (n_2(0, f) - \pi_2(f))e^{-R_{21}(f)t}, \quad (23) \\
  n_1(t, f) &= \pi_1(f) - (n_2(0, f) - \pi_2(f))e^{-R_{21}(f)t}, \quad (24)
\end{align}

where $n_1(t, f)$ and $n_2(t, f)$ are the time dependent population of the lower and higher energy states, respectively, $\pi_1(f)$ and $\pi_2(f)$ are the equivalent for the LTE situation, $R_{21}(f) = B_{21}(f)B(T, f) + A_{21}(f)$ is the rate of decay of population in $n_2(t, f)$, where $A_{21}(f)$ is the Einstein coefficient for the spontaneous emission and $B_{21}(f)$ is the Einstein coefficient for the stimulated emission. The detailed derivation of the time dependent populations, as well as the related issues, with a discussion on the assumptions is given in Appendix 1. Knowing the time dependent populations $n_1(t, f)$ and $n_2(t, f)$, the additional energy density at point $r$ due to absorption can be estimated and as it was shown in Appendix 1:

\begin{equation}
  J_a(T, r, t, f) = B(T, f) \frac{n_2(t, f) \pi_1(f)}{n_1(t, f) \pi_2(f)} \frac{1 - \exp \left(-\frac{hf}{kT}r\right)}{1 - \frac{g_1(f)n_2(t, f)}{g_2(f)n_1(t, f)}} - B(T, f), \quad (25)
\end{equation}

where $g_1(f)$ and $g_2(f)$ are the degeneracies of states 1 and 2 respectively. The dependence of $J_a(T, r, t, f)$ on $r$ comes through $n_1(t, f)$ and $n_2(t, f)$ as shown in Appendix 1. It should be noticed that the Planck function in this equation is given as the energy density as in (134) in Appendix 1 instead of as an energy flux as in (10). The subtraction with
the Planck law is due to interest in the excess energy in the system, i.e., the transmission induced noise. This energy approximation is true for the energy decay in a single point \( r \) in the space. The validity of this model is discussed in Appendix 1 and it can be shown that the model is valid at the boundaries \( (t = 0 \) and \( t = \infty)\). At \( t = 0 \), with simple manipulations, it can be shown that (154) is equivalent to the absorbed energy density at the distance \( r \). On the other hand, (154) is zero at \( t = \infty \). Therefore, the model is correct at the boundaries. However, the validity of the decay rate is not easy to show in the absence of measurement data.

Figure 6 shows the decay of the energy density according to the above non-LTE theories assuming a transmitting power \( P_{\text{Tx}} = 2 \) pW/Hz. The energy decay is rather slow, causing a large temporal distribution of the energy. Therefore, the energy of the noise is quite low. Furthermore, the energy decay rate varies with the frequency. This is a direct consequence of the frequency selective absorption as it is (indirectly) shown in Appendix 1.
2.2.6 Heat approach

If all the absorbed energy is assumed to be transformed into heat, one more approach to the transmission induced noise is obtained. In this case, the heat capacity of the medium must be considered. The heat capacity in general terms is the amount of energy to be transferred to a substance in order to change its temperature by one degree [74]. It is usually given as specific heat capacity, or specific heat, which describes the heat capacity per unit mass of the substance. The energy density of the absorbed energy is known from (19). If all this energy is assumed to be transformed into heat, the temperature increase of the medium due to the absorbed energy can be calculated. Assuming constant pressure, tabled values (e.g., [115]) can be used for the isobaric mass heat capacity $C_{pm}$. Since the energy density of the absorbed energy is known, the isobaric volumetric heat capacity $C_{PV}$ is required in order to calculate the temperature increase due to the absorbed energy density. It can be obtained from the density of the air. Then, $C_{PV} = \rho C_{pm} \text{J/cm}^3/\text{K}$, where $\rho$ is the density of the air, which can be taken from the tabled values, e.g., [116].

The temperature increase due to the absorption is given by [74, Chapter 17.5]

$$\Delta T = \frac{Q(r)}{C_{PV}},$$

(26)

where

$$Q(r) = \int_{W} P_{Tx}(f) \frac{d}{\pi r} \left(1 - e^{-\kappa_s(f)r}\right) T_p df$$

(27)

is the absorbed energy density integrated over the bandwidth $W$ of the system. The pulse length $T_p$ is required in order to describe the energy over the bandwidth $W$, i.e., the heat warming the system. The temperature increase according to (26) is shown in Fig. 7 for one pulse. The parameters for this figure were as follows: the atmospheric density is $\rho = 1.191 \times 10^{-6} \text{ kg/cm}^3$ at 294 K and 101.3 kPa [116] and the isobaric mass heat capacity is $C_{pm} = 1,011.7 \text{ J/kg/K}$ at 293 K, 101.325 kPa and a relative humidity of 50% [115]. Furthermore, the transmitting power ranges from $P_{Tx}(f) = 10^{-12} \text{ W/Hz}$ to $P_{Tx}(f) = 10^{-9} \text{ W/Hz}$, the bandwidth is $W = 9.87 \text{ THz}$ and the pulse length is $T_p = 100 \text{ fs}$. This gives a pulse energy $E_p = P_{Tx} W T_p = 0.987 \text{ pJ}$. It can be seen that the temperature increase as a consequence of one transmitted pulse is modest. Even with a billion pulses per second the temperature increase would be small. This is even when assuming a stationary environment, i.e., the temperature is not allowed to disperse, e.g., due to the Brownian motion. However, the above model is valid only for one point in space.
The contribution of the entire space to the receiver is subject to future research. That is, to investigate the aggregate radiative noise due to all points in space. Each point experiences a temperature increase due to transmitted power and each point in space radiates according to the Planck function. Then it would be possible to model the entire space. This requires further investigation and is therefore subject to the future work. However, it is possible that the noise contribution with the heat approximation will remain very small due to the modest effect of the small temperature increase to the Planck function. Especially, since the THz band is not at the peak of the Planck function at temperatures around 300 K. Therefore, the impact of the temperature increase is larger on the frequencies above 10 THz. This is easy to test and see with the Planck function.

![Temperature increase due to the absorption of the signal energy as a function of distance with several transmit powers](image)

**Fig. 7.** Temperature increase due to the absorption of the signal energy as a function of distance with several transmit powers [© 2016 Elsevier].

### 2.3 Small particle scattering

#### 2.3.1 LOS loss on small particles in air

In order to model the loss to the received signal due to molecular and particle scattering, a scattering coefficient is utilized [54]. The scattering coefficient is the same for the scattering loss, and what the absorption coefficient is for the absorption loss, i.e., the
effective area of the molecules/particles per unit volume. Therefore, the attenuation caused by the scattering $A_{\text{sca}}(f)$ on the LOS component can be modelled with and inverse of the Beer-Lambert law since the transmittance for scattering is defined similarly as the transmittance for absorption [54]. Thus,

$$A_{\text{sca}}(f) = \exp(\Sigma_j \kappa^s_j(f)r),$$

(28)

where the sum is over the scattering species, i.e., a sum over molecular species and small particle species. The scattering coefficient $\kappa^s_j(f)$ for scattering species $j$ is

$$\kappa^s_j(f) = N^s_j \sigma^s_j(f),$$

(29)

where $N^s_j$ is the number density of scattering species $j$ and $\sigma^s_j(f)$ is the scattering cross section for species $j$. Notation $\kappa_s(f)$ is used for the total summed scattering coefficient, i.e., $\kappa_s(f) \equiv \Sigma_j \kappa^s_j(f)$. Knowing the scattering loss,

$$A_{\text{mol}}(f)A_{\text{sca}}(f) = e^{(\kappa_{\text{ae}}(f) + \kappa_s(f)r)} = e^{\kappa_s(f)r} = A_{\text{ext}}(f)$$

(30)

is the total extinction loss with an extinction coefficient $\kappa_s(f)$. Indeed, the scattering “shadows” a part of the radiation by redirecting the radiation, causing an additional loss to the LOS component. It should be noticed that scattering causes losses on the LOS path due to the Beer-Lambert law. Therefore, the average amount of scattered energy is not received, but merely lost. The total received LOS power at the receiver, including absorption loss and scattering loss, can be calculated with

$$P_r = \int_{W} P_{\text{Tx}}(f) \frac{e^{-\kappa_s(f)r}}{(4\pi f r)^2} df,$$

(31)

where $P_{\text{Tx}}(f)$ is the PSD of the transmitted signal in W/Hz and $W$ is a bandwidth over which the power is calculated.

### 2.3.2 Scattering coefficient

In order to calculate the total extinction coefficient $\kappa_s(f)$, the scattering cross section must be calculated first. The type of the scattering (Rayleigh, Mie, or geometric scattering) is defined by the size parameter $q = \pi x_d / \lambda$, where $\lambda$ is a wavelength and $x_d$ is the diameter of scattering particle [117]. Since the wavelengths in the THz band are long in comparison with the average particle size (diameters $\sim 1 \mu\text{m} - 5 \mu\text{m}$), Rayleigh
scattering is dominant. In contrast, at visible light frequencies, the size of the particles is comparable to the wavelengths, thus Mie scattering [55] becomes dominant. Mie scattering causes strong forward scattering, whereas Rayleigh scattering has a nearly circular/spherical scattering phase function. When the size parameter \( q \gg 1 \), geometric scattering should be utilized [55, 56]. Geometric scattering may also be accompanied by diffusion scattering as the EM wave is partly scattered in random directions on a rough surface [67]. There exists a time domain channel model specifically for 300 GHz indoor communication [118], which is models the geometric scattering on macroscale objects. In this thesis, the scattering characteristics are approximated with Rayleigh scattering due to the interest in molecular/aerosol scattering in THz frequencies.

The characteristics for Rayleigh scattering are given by the Rayleigh scattering cross section. There are multiple ways to evaluate the scattering cross section. First of all, the classical Rayleigh scattering cross section for hard spherical scatterers can be calculated with [53]

\[
\sigma_s(f) = \frac{128\pi^5\beta^2}{3\lambda^4},
\]

where \( \beta \) is polarizability of a molecule. The polarizability can be approximated with [119]

\[
\beta = \frac{n(f)^2 - 1}{2n(f)^2}\left(\frac{x_d}{2}\right)^3,
\]

where \( n(f) \) is a refractive index of the medium at frequency \( f \). In the absence of refractive index measurement data in the THz band, \( n(f) = 1.6 \) is used over all frequencies, which is approximately an average refractive index of mineral dust at visible light frequencies [120]. Also, this refractive index is quite close to the refractive index of polystyrene at visible light and near-infrared [121, 122], a material commonly used in artificial micro-spheres. The scattering cross section for hard spheres with particle diameter dependency becomes

\[
\sigma_s(f) = \frac{2\pi^5x_d^6}{3\lambda^4}\left(\frac{n(f)^2 - 1}{n(f)^2 + 2}\right)^2.
\]

There also exists a more accurate method for the Rayleigh scattering cross section evaluation for molecules [54], i.e.,

\[
\sigma_s(f) = \frac{24\pi^3}{\lambda^4(N)^2}\left(\frac{n(f)^2 - 1}{n(f)^2 + 2}\right)^2\frac{6 + 3f}{6 - 7f},
\]

where \( f \) is a depolarization ratio for a molecule and ratio \( (6 + 3F)/(6 - 7F) \equiv D \) is depolarization factor for the air molecules (excluding water vapour) and it can be
approximated as [54]

$$D(\text{air}) = \frac{78.084D(N_2) + 20.946D(O_2) + 0.934 + 1.15C_{CO_2}}{78.084 + 20.946 + 0.934 + C_{CO_2}}, \quad (36)$$

where $D(N_2) = 1.034 + 3.17 \times 10^{-4} \lambda^{-2}$ is the depolarization ratio of Nitrogen, $D(O_2) = 1.096 + 1.385 \times 10^{-3} \lambda^{-2} + 1.448 \times 10^{-4} \lambda^{-4}$ is the equivalent for Oxygen and $C_{CO_2}$ is a concentration factor for carbon dioxide for standard air, so the $C_{CO_2} \approx 0.036$.

For particle scattering, aerosols are assumed to be spherical particles with a diameter $x_d$. This allows particle scattering to be treated by (34). It is also assumed that the scattering on the aerosols is elastic, i.e., the wavelength of the EM wave does not change as result of scattering event. However, the scattering particles are not equal in size, but the number density distribution $N_d(x_d)$ (usually) follows a log-normal distribution [123]. Also, gamma distributions have been used for water droplets [123, 124]. The log-normal distribution is defined as [125, 126]

$$N_d(x_d) = \frac{dN}{dx_d} = \frac{N}{x_d \sqrt{2\pi \ln(\sigma_g)}} e^{-\frac{(\ln(x_d)-\ln(r_g))^2}{2(\ln(\sigma_g))^2}}, \quad (37)$$

where $r_g$ is the mean geometric diameter of the scatterers and $\sigma_g$ is the corresponding standard deviation. To calculate the total scattering coefficient for aerosol particles, (37) must be multiplied with (34). Integrating over the number density distribution, the scattering coefficient for aerosols becomes [123]

$$\kappa_s(f) = \int_0^\infty \frac{2\pi^5 x_d^5 N_d}{3\lambda^4 \sqrt{2\pi \ln(\sigma_g)}} \left( \frac{n(f)^2 - 1}{n(f)^2 + 2} \right)^2 e^{-\frac{(\ln(x_d)-\ln(r_g))^2}{2(\ln(\sigma_g))^2}} \, dx_d. \quad (38)$$

On the other hand, it is possible to buy spherical particles for laboratory experiments. Then it is possible that the number density distribution becomes Gaussian. When assuming a normal distribution for the scattering particles, the number density distribution becomes

$$N_d(x_d) = \frac{N}{\sigma_{sp} \sqrt{2\pi}} e^{-\frac{(x_{sp}-x_d)^2}{2\sigma_{sp}^2}}, \quad (39)$$

where $r_{sp}$ is the mean diameter of the scattering particles and $\sigma_{sp}$ is the standard deviation of the diameters. By multiplying the number density distribution with the scattering
cross section and integrating over the diameters \( x_d \), the scattering coefficient for normally distributed particles becomes

\[
\kappa_s(f) = \int_0^\infty \frac{2\pi^5 f^4 x_d^5 N_s}{3c^4 \sigma_p \sqrt{2\pi}} \left( \frac{n(f)^2 - 1}{n(f)^2 + 2} \right)^2 e^{-\frac{(x_d - x_p)^2}{2\sigma_p^2}} \, dx_d.
\] (40)

The total number density \( N \) of the aerosol particles is usually counted as a maximum of hundreds of particles per cm\(^3\) [123, 127]. The number of aerosol particles may be as high as one million per cubic centimetre [128]. However, these large number densities for aerosols are produced by very small particles, which do not cause significant scattering in the THz band. Therefore, atmospheric conditions in which the scattering particles are larger, but fewer in numbers are considered. The size of an average larger aerosol particle, e.g., water droplet, is circa 1–10 \( \mu \)m [123]. The variety of size distributions, e.g., for sea salt aerosols can be found in [127], or for fog particles in [129].

The aerosol particles are assumed to be perfect scatterers (the imaginary part of the refractive index is zero, i.e., absorption coefficient for aerosol particles is zero [130]). Regardless, the scattering has a contribution to the LOS loss, since the total transmittance of the LOS component is \( \exp(-\kappa_a(f) + \kappa_s(f))r \). Summing the absorption, molecular scattering and particle scattering cross sections, the total transmittance becomes

\[
\tau_{\text{tot}}(f) = \exp[-r(\Sigma_i \kappa_{a_i}(f) + \Sigma_k \kappa_{c_k}(f) + \kappa_s(f))].
\] (41)

Furthermore, inverse of (41) is equal to (30).

### 2.4 Multiple scattering theory

If a very wide pulse is assumed (e.g., a 10 THz wide pulse transmitted to the channel), the transmitted time domain signal is approximately an impulse. Yet, an impulse will not be received in the THz band due to the frequency selective molecular absorption and scattering. The received signal can be evaluated by the convolution between the channel impulse response and the transmitted waveform. Thus, the output of the channel is

\[
y(t) = h(t) * x(t),
\] (42)

where \( * \) denotes a convolution and

\[
h(t) = \mathcal{F}^{-1} \left\{ \sqrt{(A_{\text{FSPL}}(f)A_{\text{mol}}(f)A_{\text{sca}}(f))^{-1}} \exp(-i2\pi f^2 r^2/c) \right\}
\] (43)
is the amplitude impulse response of the channel according to the inverse Fourier transform reciprocal of (8), where \( \exp(-i2\pi fr/c) \) is a linear phase term, where \( r \) is the distance between the Tx and the Rx. A similar treatment to the LOS path is given in [42]. The reciprocal of (8) is telling the frequency response of the channel at the receiver and an impulse response calculated this way is valid for the LOS component of the channel. Due to the frequency selectivity, a wideband signal will have a transient even without NLOS components. The time domain representation of the LOS channel in (43) does not allow scattered energy to reach the receiver. The multiple scattering channel models are considered next, which allow the scattered energy to propagate, thus reach the receiver after one or more scattering events.

2.4.1 Photon diffusion model

A time domain channel model given in [57] is based on diffusion of photons in two-dimensional (2D) space. This model utilizes a random walk of photons in an environment with randomly distributed particles. When a photon encounters a particle, it changes direction randomly, or is absorbed based on the albedo \( \varpi \) of a particle. The albedo can be estimated by the difference between absorption and scattering cross sections with \( \varpi = \sigma_s / (\sigma_a + \sigma_s) \). Thus, in mostly scattering environment \( \varpi \approx 1 \) and in mostly absorbing environment \( \varpi \approx 0 \).

This channel model can be represented as [57]

\[
I(r,R) = \frac{e^{-\eta r}}{2\pi r} \delta(R-r) + \sum_{N=1}^{\infty} \frac{\varpi^N}{2\pi} \frac{2\sqrt{\pi} \eta^N}{2^N 1^{(N+1)} 1^{(N/2)}} \times e^{-\eta \delta (R^2 - r^2)} \Theta(R-r),
\]

(44)

where \( I \) is the probability that photon reaches a point at distance \( r \) after traveling the total distance \( R \equiv ct \) (or an energy flux per unit transmitting power), where \( N^c \) is the number of collisions in the path \( R \). \( \eta \) is the number of obstacles in the path of photon per unit distance (effectively \( \eta = 1/l_a + 1/l_s = \kappa_a + \kappa_s \), where \( l_a = 1/\kappa_a \) is the absorption length and \( l_s = 1/\kappa_s \) is the scattering length), \( \delta(\cdot) \) is Dirac’s delta function. The model is presented here as a combination of ballistic signal component, referring to the directly propagated term for which \( N^c = 0 \), and the components for which \( N^c > 0 \). The equation is also multiplied with factor \( \varpi^N / \eta \), in which \( \varpi^N \) introduces the absorption of photons.
over $N^c$ collisions and $\eta$ is related to a probability of a photon to undergo a collision in a certain point. This term vanishes on the ballistic component [57].

### 2.4.2 Radiative transfer solution

Paasschens [58] provides an exact solution to the radiative transfer equation in two and four dimensional spaces. He derives the models from the Boltzmann diffusion equation, which leads to similar thinking as in the case of random walk. He then derives the transfer equation for three dimensional space by interpolation from two and four dimensional spaces. The three–dimensional (3D) equation is represented as

$$
I(r,t) \simeq \frac{e^{-ct(\frac{1}{2} + \frac{1}{\pi})}}{4\pi r^2} \delta(r - ct) + \sum_{N^c=1}^{\infty} \frac{e^{-ct(\frac{1}{2} + \frac{1}{\pi})}}{\pi l_s^3} \frac{\Gamma(\frac{3}{4}N^c + \frac{1}{2})}{\sqrt{\pi N^c} \Gamma(\frac{3}{4}N^c)}
$$

$$
\times \left(\frac{ct}{l_s}\right)^{N^c-3} \left(1 - \frac{r^2}{c^2t^2}\right)^{\frac{1}{2}N^c-1} \Theta(ct - r). \tag{45}
$$

As in the case of photon diffusion model, the ballistic component has been combined with the components for which $N^c > 0$ and multiplied the equation with an absorption term $e^{-ct}$. It can be seen that the absorption term is identical to the transmittance of the channel due to molecular absorption. Besides the ballistic component and the absorption term, this equation was also used to estimate the propagation delay of the transmission induced noise in Section 2.2.4. There, the absorption coefficient was used as a free space path length measure, whereas here the same measure is the actual scattering coefficient.

### 2.4.3 Narrowband multiple scattering channel model for the THz band

In [58] a 2D diffusion model is also given, which is in fact identical to the 2D diffusion model presented in [57]. The proof is provided in Appendix 2. The reason, why both models are required, is a validation issue. By showing that both 2D models are identical, although they based on different derivations, it can be verified that the models are correct from the theoretical point of view. Thus, a 3D channel model for pulse based communications based on these two original papers can be shown. This is done by modifying the impulse theory in [58] similarly as was done in [57]. The correctness of the procedure is provided in the Appendix 2, where the 2D model of [58] is modified.
according to [57] and then the models are shown to be equal. Similarly, the 3D diffusion theory in [58] is modified according to the pulse signal presentation in [57] in order to achieve 3D propagation model for arbitrary input signal waveform.

The impulse terms are replaced with the pulse representation by integrating the equation over \( ct \), i.e., over all the possible distances travelled by the incidence radiation in time \( t \). Moreover, there are two terms: the radiation flux during the time the pulse is on \( (r/c \leq t < r/c + T_p) \) and after the pulse has passed the receiver \( (t \geq r/c + T_p) \) [57]. Otherwise the flux is zero. Then a model for pulse based radiative transfer in a 3D environment is achieved as

\[
I(r,t) = \begin{cases} 
  I_1(r,t) & r/c \leq t < r/c + T_p, \\
  I_2(r,t) & t \geq r/c + T_p,
\end{cases}
\]

where

\[
I_1(r,t) = \begin{align*}
&= e^{-r(\frac{1}{4\pi r^2})} f\left(t - \frac{r}{c}\right) \\
&+ \int_r^{ct} \sum_{N=1}^{\infty} f\left(t - \frac{R}{c}\right) \frac{e^{-R(\frac{1}{4\pi r^2})}}{\pi l_s^2} \frac{\Gamma(\frac{3}{2}N_c + \frac{3}{2})}{\sqrt{\pi N_c! \Gamma\left(\frac{3}{4}N_c\right)}} \\
&\times \left(1 - \frac{r^2}{R^2}\right)^{\frac{3}{2}N_c - 1} dR,
\end{align*}
\]

\[
I_2(r,t) = \begin{align*}
&= \int_{ct-T_p}^{ct} \sum_{N=1}^{\infty} f\left(t - \frac{R}{c}\right) \frac{e^{-R(\frac{1}{4\pi r^2})}}{\pi l_s^2} \\
&\times \frac{\Gamma(\frac{3}{2}N_c + \frac{3}{2})}{\sqrt{\pi N_c! \Gamma\left(\frac{3}{4}N_c\right)}} \left(1 - \frac{r^2}{R^2}\right)^{\frac{3}{2}N_c - 1} dR,
\end{align*}
\]

where \( f(\cdot) \) is the power envelope of the transmitted signal. In order to specify this model for the THz band, THz specific parameters need to be utilized for this channel model. The scattering coefficient is directly included in \( l_s \), the absorption coefficient is directly included in \( l_a \) and the free space path loss in the 3D environment is \((4\pi r^2)\) (without the antenna aperture term). Thus, the above equations now take into account all the attenuation terms in (8). This model works for an arbitrary input signal waveform. The integrals in (47) and (48) are in fact convolutions between the radiation flux of the channel and the input signals power envelope.

It has been shown that the 2D impulse models in [57] and [58] are identical. At the same time it was shown that the pulse theory in [57] can be achieved with modifications
to the 2D model in [58] (in Appendix 2). Furthermore, the 3D model [58] is expected to be accurate due to interpolation from two exact solutions of radiative transfer theory. Therefore it is assumed that (46) is very accurate for modelling the 3D radiative transfer. In the numerical results in Section 2.5, it will be shown that this model agrees very well with Monte Carlo simulations based on following the individual photons through a vast number of collisions.

2.4.4 **Wideband multiple scattering channel model for the THz band**

The multiple scattering channel model presented above (46) performs well in a situation when the channel frequency response is flat, such as for narrowband signals. In the case of wideband signals, the non-flat frequency response of the channel has to be taken into consideration. In order to account for multiple scattering, the multiple scattering radiative flux, which was introduced in (45) and derived in [58], is utilized. The problem is that the impulse response represents the radiative transfer in a flat channel, i.e., the frequency domain variables need to be constant. In order to solve this problem, a new approach to calculation of the impulse response is required.

The starting point is the radiative flux $I(r,t)$ presented in (45). Since in the wideband case $l_a$ and $l_s$ are frequency dependent, this equation gives the time-frequency function for the radiative flux. When the distance $r$ is fixed, the resulting time-frequency function is denoted as $I(t,f)$. Because of the vast number of scatterers, the average received power is a summation over the powers of the different paths [57]. This assumption relies on the fact that the random polarization of the received photons is averaged out as the number of scattering events is large [57]. Furthermore, it can assumed that the scattering events are uncorrelated. Thus, this is a wide sense stationary uncorrelated scattering (WSSUS) channel [57, 131].

The philosophy behind the wideband multiple scattering channel model is that, based on the time-frequency grid $I(t,f)$, one can calculate a channel transfer function for each time instant $t = t_i$. This explicitly gives an impulse response corresponding to the frequency response. The procedure can be done for each time instant described by (45). Thus, the result is a WSSUS channels time-varying impulse response, which is a function of time and delay. By summing all the corresponding power delay profiles, a single power delay profile, or impulse response for the entire THz band, is obtained. The procedure is as follows. By utilizing the time-frequency function $I(f,t)$ of the radiative
flux, the channel transfer function for each time instant $t_i$ can be calculated
\[ H(f, t = t_i) = \sqrt{I(f, t = t_i)}e^{-i2\pi ft_i}, \]
(49)
where zero value is assumed for $t_i$ on the arrival of the first signal component, or the
time instant $r/c$. This is not a crucial issue, as it only rotates the phase of the transfer
function and eventually a power summation is considered. Therefore, the phase of the
signal disappears and only the time shift of the phase in the upcoming inverse Fourier
transform is important. $H(f, t)$ is the time-varying transfer function according to the
WSSUS theory [131]. As a consequence, the time-varying impulse response $h(\epsilon, t)$ can
be obtained by an inverse Fourier transform from $H(f, t)$, where $\epsilon$ is the delay. The
linear phase term $\exp(-i2\pi ft_i) = \exp(-i\omega t_i)$, where $\omega$ is the angular frequency, in
(49) is important, as it gives each impulse response an explicit time dependency. Thus,
the time-varying impulse response depends on time and delay. From the time-varying
impulse response, one can obtain the delay cross-power density $h(\epsilon, \Delta t)$ by taking an
autocorrelation with respect to $t$. The channel power delay profile can be obtained by
setting the time-difference $\Delta t$ to zero [131]. On the other hand, it is straightforward to
write the power delay profile as
\[ h(\epsilon) = \int |h(\epsilon, t)|^2 dt, \]
(50)
which is equivalent to taking the autocorrelation of $h(\epsilon, t)$ with respect to $t$ and setting
$\Delta t$ to zero.

Could the same thing be done by just directly summing (45) over all the frequencies,
which would effectively give a power summation? The answer is no. It is known that the
frequency response is highly frequency selective in the THz band. This is true also in
the presence of the scattering particles. However, direct summation over the intensities
given by (45) would be an impulse (in the absence of scattering particles), which is
strictly incorrect in a frequency selective channel. This impulse behaviour of (45) is
caused by the assumption of the flat frequency response, which is not generally the
case in the THz band, except for the narrowband signals. When the summation is done
as was done above, in the absence of the scattering particles, the impulse response
becomes similar to (43). The only difference is that the impulse response becomes
a power delay profile. This partly validates the model. However, the full validation
requires further measurements in the future. Still, it can be shown that (50) produces
exactly the same power delay profile as (45) if the absorption and scattering parameters
are flat. This further validates the model, but real measurements are necessary in order
to fully validate the model. After the power delay profile has been calculated, the output of the channel can be found as a convolution between the power delay profile and the signal power envelope.

### 2.4.5 Antenna aperture in the time domain models

One important observation should be made concerning the above time domain theories: none of the above time domain channel models have an antenna aperture term in the path loss. The absence of this term does not have an impact on the accuracy of the theories due to the change of unit of the result. The path loss with the antenna aperture term indicates the received power in W/Hz. Without the antenna aperture term, the unit of the result is the radiation flux in W/Hz/cm$^2$.

All the frequency domain models were given with the frequency dependent antenna aperture term ($4\pi(f/c)^2$) following the well-known Friis transmission equation. However, in the case of multiple scattering, the link to the Friis equation is no longer evident. Since (50) can be utilized in all atmospheric conditions, one should choose the antenna aperture term based on the demanded accuracy. The simplest case is the frequency dependent aperture term ($4\pi(f/c)^2$). In reality, in wideband applications the antenna gains should also be frequency dependent, and thus have to be included in the aperture term. Including the aperture term in the impulse response has to be done in the time domain, requiring convolution between impulse response and antenna impulse response. One could also multiply antenna aperture at the early stage of the derivation, e.g., to (49) by $H(f, t = t_i) = \sqrt{\hat{R}(f, t = t_i)c^2/(4\pi f^2)}e^{-i2\pi ft_i}$. However, the antenna impulse response allows more sophisticated treatment of the aperture term, e.g., in ultra-wideband (UWB) communications, as it was shown in [42, 132]. The main interest here is in the radiative transfer, thus the ideal aperture term is utilized, or no aperture at all (radiative flux at distance $r$).

### 2.5 Numerical results for the small particle scattering

In this section, the numerical examples of the various molecular phenomena presented will be shown. These include molecular absorption, LOS scattering loss, CIA, and multiple scattering.
Table 1. Aerosol number and size distribution parameters for the three cases used in this thesis.

<table>
<thead>
<tr>
<th>Case</th>
<th>$r_e$ (µm)</th>
<th>$\sigma_e$ (µm)</th>
<th>$N_s$ (1/cm$^3$)</th>
<th>Ref.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case 1</td>
<td>2.81</td>
<td>2.40</td>
<td>13.4</td>
<td>[127]</td>
</tr>
<tr>
<td>Case 2</td>
<td>3.20</td>
<td>2.30</td>
<td>484.4</td>
<td>[129]</td>
</tr>
<tr>
<td>Case 3</td>
<td>4.10</td>
<td>2.64</td>
<td>17.89</td>
<td>[127]</td>
</tr>
</tbody>
</table>

2.5.1 Scattering loss on small particles

Three cases are utilized for the realistic scattering parameters and the log-normal size distribution for the scattering particles. The corresponding cases with the scattering parameters are given in Table 1. These cases are for outdoor measurements, but several papers suggest similar values for indoor aerosol particles, e.g., [133–135]. The measurements made in [133, 134] show strong correlations between outdoor and indoor aerosol particle size distributions. However, very few papers give explicit values for required size distributions. For instance, [133] gives the mean and standard deviation for indoors, which gives a similar performance to the case 3 parameters. The LOS scattering losses for the cases 2 and 3 are given in Figs. 8 and 9, respectively. The scattering makes a clear contribution to the total loss in the THz band in both cases, but especially with the case 3 parameters. This shows that the scattering loss is very sensitive to the average size of the scattering particles, but not so much to the number density of the scattering particles. This is a logical result, as the scattering cross section is dependent on the sixth power of the diameter of a scattering particle. Therefore, the scattering loss increases strongly as the average diameter of the scattering particles increases.

Due to strong frequency dependency, scattering loss increases as a function of frequency, but also as a function of distance through Beer-Lambert law. Standard air molecules make a minimal contribution to the total scattering loss. Therefore, the scattering loss in practise is not affected by temperature, pressure, or relative humidity, given these environmental aspects are not considered to have an effect on the existence of the aerosols. This means that the opportunity for multiple scattering increases as absorption is decreased. This is due to the general albedo of the atmosphere: the more energy is absorbed, the less energy there is for scattering. As it has been shown, the scattering should not be neglected from the attenuation calculations. On the other hand, in a standard atmosphere without any specific scattering particles, scattering can be neglected due to insignificant scattering on molecules. The effect of scattering
Fig. 8. The LOS scattering loss in the THz band with the case 2 scattering particle parameters from Table 1. White lines are illustrating the contours for the scattering loss as a function of distance and frequency [© 2015 IEEE].

Fig. 9. The LOS scattering loss in the THz band with the case 3 scattering particle parameters from Table 1. White lines are illustrating the contours for the scattering loss as a function of distance and frequency [© 2015 IEEE].
at the higher frequencies has been studied in [124]. The author provided attenuation calculations for frequencies from 10 GHz to 1 THz. The effect of the scattering was found to be very small, a result that the study here confirms. At the lower end of the THz band, the effect of the scattering is not significant. However, at the higher end of the THz band the scattering may cause significant losses. The white lines in Figs. 8 and 9 correspond to the scattering loss contours indicated in the figures. These contours give the loss as a function of distance and frequency, below which the scattering loss is smaller than a given value. It can be seen that in the presence of scattering particles, the lower part of the THz band is safe. However, at the higher end, the scattering loss rapidly increases. When power limited nanodevices are considered, the received power is expected to be very small due to the low transmitting power. In this case even a small additional loss on top of the molecular absorption may take the system below the minimum operation threshold. Thus, scattering modelling is required in order to fully estimate the performance of the nanodevice-to-nanodevice links.

2.5.2 Validation of the narrowband multiple scattering channel model

The validation of the narrowband multiple scattering channel model in (46) is given in Fig. 10. The figure is achieved by utilizing the absorption and scattering coefficients $\kappa_a = \kappa_s = 0.015$ cm$^{-1}$. The distance $r$ is approximately 33 cm. The narrowband model is validated with a Monte Carlo simulation, which was conducted by following a randomly wandering photon in a scattering environment. The mean free path length was determined by exponential distribution [57] with the mean value $1/(\kappa_a + \kappa_s)$. Furthermore, $N_c$ collisions absorb $\sigma N_c$ part of the radiation flux. By following the photon through a vast number of collisions and by repeating this to a vast number of photons, the flow of photons through a unit area at distance $r$ can be calculated. For pulse signals, this procedure has to be repeated multiple times, depending on the pulse length. That is, delayed photons are sent to correspond to the length of the pulse. Scaling the result with the time resolution of the simulation and taking into account the FSPL, the result of the Monte Carlo simulation is presented in Fig. 10. The Monte Carlo simulation agrees very well with the theoretical model narrowband signalling. This validates the narrowband channel model.
2.5.3 The effect of the wideband multiple scattering to the coherence bandwidth of the channel

A wideband multiple scattering channel model is given in (50). To be more specific, this equation gives the power delay profile of the channel. Fig. 11 gives the power delay profiles for different scattering parameter cases given in Table 1 at 10 cm distance. One LOS scattering case, i.e., the case without multiple scattering, has been included, which has been calculated with (43) without the antenna aperture in FSPL. Fig. 11 shows that the tail of the signal gets longer as the scattering coefficient is increased (as the case 3 is giving highest scattering coefficient). This is also shown by the delay spread of the power delay profiles. The root mean square (rms) delay spread can be calculated with [136]

$$\varepsilon_m = \frac{\int \varepsilon h(\varepsilon) d\varepsilon}{\int h(\varepsilon) d\varepsilon},$$ \hspace{1cm} (51)

$$\varepsilon_{rms} = \sqrt{\frac{\int (\varepsilon - \varepsilon_m)^2 h(\varepsilon) d\varepsilon}{\int h(\varepsilon) d\varepsilon}},$$ \hspace{1cm} (52)

Fig. 10. Validation of the narrowband multiple scattering model © 2015 IEEE.
where $\epsilon$, $\epsilon_m$, and $\epsilon_{\text{rms}}$ are delay, mean delay spread and rms delay spread, respectively. Based on the above equations, the rms delay spread and coherence bandwidth values for a 10 cm distance are presented in Table 2 following the usual assumption that the coherence bandwidth $B_c \approx 1/\epsilon_{\text{rms}}$. In order to filter out the non-significant signal components, the signal components below 30 dB of the peak power were filtered out. A similar procedure was done in [137] for ultra-wideband channel modelling. It can be seen that even the case 1 parameters have an effect on the delay spread and the coherence bandwidth, although the overall effect on the tail term is small. However, it is significantly larger than in the LOS case for the case 1 parameters, although Fig. 11 shows that the transients are on top of each other (multiple scattering versus LOS case). The LOS scattering transient term is in practise zero after a short ripple caused by the molecular absorption, while the multiple scattering case has several orders of magnitude higher values at the transient. Regardless, the transient term in the multiple scattering curve for case 1 parameters is small in comparison with the other two scattering cases.

It can also be seen that the case 3 parameters for LOS scattering produces a slightly longer delay spread than the other two cases. This is due to the large attenuation at the higher frequencies, effectively narrowing the frequency response, thus causing a longer delay spread. It should be noticed that for the LOS cases the below peak power threshold could be higher than 30 dB since without the multiple scattering the transient term vanishes quickly. However, in the case of multiple scattering the case 3 scattering parameters requires a rather low threshold due to the long tail. Thus, by using a 30 dB threshold for all the cases, the values are comparable. The actual coherence bandwidths for the LOS cases are not as high as Table 2 suggests. Regardless, it can be seen that even the LOS cases have delay spread. This is due to the frequency selectivity of the molecular absorption, causing a finite coherence bandwidth, which leads to a delay spread. This may lead to ISI if symbol separation is not sufficiently large. The symbol-to-symbol separation should be much larger than the delay spread of the channel in order to counteract the errors due to the ISI. It is clear that the probability of ISI is significantly increased if the multiple scattering is taken into consideration.

### 2.6 Energy detection in a multiple scattering environment

This section considers effect of the multiple scattering to the energy detection. Molecular absorption itself causes energy distribution over time due to frequency selective loss.
Table 2. Delay spread and coherence bandwidth of the power delay profile for 10 cm distance for the scattering parameter cases given in Table 1. The LOS cases are included. The values are given for a threshold of 30 dB below the peak power.

<table>
<thead>
<tr>
<th>Case</th>
<th>$\varepsilon_{\text{rms}}$ (ps)</th>
<th>$B_c$ (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case 1</td>
<td>87.1</td>
<td>11.5</td>
</tr>
<tr>
<td>Case 2</td>
<td>480.1</td>
<td>2.1</td>
</tr>
<tr>
<td>Case 3</td>
<td>2362.9</td>
<td>0.4</td>
</tr>
<tr>
<td>Case 1 LOS</td>
<td>0.64</td>
<td>1.55</td>
</tr>
<tr>
<td>Case 2 LOS</td>
<td>0.64</td>
<td>1.55</td>
</tr>
<tr>
<td>Case 3 LOS</td>
<td>0.81</td>
<td>1.23</td>
</tr>
</tbody>
</table>

The multiple scattering further increases the tail response as shown earlier. Therefore, the energy detection performance is considered in this section.
2.6.1 Theory of detection of deterministic unknown signal

The landmark paper by Urkowitz [138] describes the detection of unknown deterministic signals. Its framework fits the scope here perfectly, since the interest is on the energy detection of OOK modulated signals. From the definition of the power delay profile \( h(t) \) in (50), the received energy can be determined as

\[
E(T_i) = \int_0^{T_i} [p(t) * h(t) + N_n(t)] dt,
\]

(53)

where \( T_i \) is the integration time, \( p(t) \) is the signal power waveform, \( N_n(t) \) is the noise power and \( * \) denotes convolution. Since the signal and the power delay profile are defined for power, the convolution is equivalent to the frequency domain multiplication of the PSDs of the signal and the channel. \( E(T_i) \) can also be called as a detection variable, determining whether a received symbol is detected as “1”, or “0”. The above equation assumes that the starting point of the integration is at the arrival instant at the received signal. Since OOK modulation is assumed, \( E(T_i) \) is just an integration over the noise for bit “0”. The noise power spectral density is defined traditionally, i.e.,

\[
N_0 = k_B T_a,
\]

(54)

where \( N_0 \) is a one-sided baseband per hertz noise power. It can be assumed that the integrated noise power follows a chi-squared distribution with \( 2T_iW \) degrees of freedom, where \( T_i \) is the integration time and \( W \) is the bandwidth of the signal/detector [138]. The signal-plus-noise energy follows a non-central chi-squared distribution with \( 2T_iW \) degrees of freedom and with a non-centrality parameter \( \Upsilon \) [138]. As a consequence, the energies are

\[
E_n \sim \chi^2_{2T_iW},
\]

(55)

\[
E_y \sim \chi^2_{2T_iW}(\Upsilon),
\]

(56)

where \( E_n \) is the noise energy, \( E_y \) is the signal plus-noise-energy and the non-centrality parameter is defined as

\[
\Upsilon = \frac{\int_0^{T_i} p(t) * h(t) dt}{\int_0^{T_i} W N_0 dt} = \frac{E_s(T_i)}{T_i W N_0},
\]

(57)

where \( E_s \) is the symbol energy. Since the unit of \( p(t) \) is W, \( h(t) \) is W/cm\(^2\) and \( N_0 \) is W/Hz, the integrated signal energy becomes J/cm\(^2\) and the unit of the noise energy
is J. Thus, the non-centrality parameter here is equivalent to the SNR per antenna aperture (SNR/cm²). The optimal decision threshold $V_T$ lies at the intersection of the two chi-squared distributions. This way most of the probability mass of the noise and the signal lies as far away as possible from the threshold. The worst case would be the SNR approaching zero. Then the detection would be a coin flip: there is approximately 50% probability to receiving a bit, since the threshold would lie at the mean of the chi-square distribution.

The chi-squared distribution is highly asymmetric at the lower degrees of freedom. However, at large degrees of freedom, the chi-square distributions have an approximately Gaussian shape. Thus, the intersection point can be estimated to be the average of the mean values of the chi-squared distributions

$$V_T = \frac{\mu_s + \mu_n}{2} = 2TW + \frac{1}{2}Y,$$

(58)

where $\mu_s$ and $\mu_n$ are the mean values for the two chi-squared distributions. Furthermore, the mean for the chi-square distribution is $2TW$ and the mean for the non-central chi-square distribution is $2TW + Y$. Thus, producing the right-hand size term. This threshold is utilized regardless of the degrees of freedom, or the optimality of the approach.

The decision is made based on the above threshold and the corresponding error probabilities are

$$Q_{fa} = P(E_n > V_T | H_0) = P(\chi^2_{2TW} > V_T),$$

(59)

$$Q_{md} = 1 - P(E_y > V_T | H_1) = 1 - P(\chi^2_{2TW}(Y) > V_T),$$

(60)

where $Q_{fa}$ and $Q_{md}$ are the probability of a false alarm and the probability of a missed detection, respectively, $P(\cdot)$ denotes probability and $H_0$ and $H_1$ are the hypotheses for receiving bit “0” or “1”, respectively. Also, $1 - Q_{md}$ is a probability of detection. In the case of ISI, the hypothesis $H_0$ is also drawn from non-central chi-square distribution assuming the signal energy leaks to the adjacent symbol slot. For instance, in the case of bit combination “11”, the ISI helps the detection of the second symbol. However, in the case of bit combination “10”, the ISI increases the probability of false alarm. The situation is more difficult in the case of the ISI reaching over multiple symbol slots. In the numerical examples the focus is on the analysis of one symbol and further analysis is left to future work. Given the above error probabilities, the bit error probability (BEP) for uniform data distribution can be calculated as

$$BEP = \frac{1}{2}Q_{fa} + \frac{1}{2}Q_{md}.$$
Table 3. Aerosol number and size distribution parameters for the four scattering cases.

<table>
<thead>
<tr>
<th>Case</th>
<th>$\mu_{sp}$ (µm)</th>
<th>$\sigma_{sp}$ (µm)</th>
<th>$N_s \times 10^7$ (1/cm$^3$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case A</td>
<td>2.07</td>
<td>0.15</td>
<td>21.5</td>
</tr>
<tr>
<td>Case B</td>
<td>2.50</td>
<td>0.16</td>
<td>12.2</td>
</tr>
<tr>
<td>Case C</td>
<td>2.87</td>
<td>0.40</td>
<td>8.1</td>
</tr>
<tr>
<td>Case D</td>
<td>3.87</td>
<td>0.25</td>
<td>3.3</td>
</tr>
</tbody>
</table>

2.6.2 Numerical examples of energy detection under multiple scattering conditions

Many parameters affect the received signal such as aerosol particle size distribution (scattering properties), atmospheric temperature and pressure (absorption properties) and the integration time at the energy detector (received SNR). The scattering parameters used here are given in Table 3. Particle parameters found on the web site of Bangs Laboratories, Inc.$^2$ have been used. Also, other manufacturers for microspheres exist, such as Polysciences, Inc.$^3$ By using the scattering parameters of commercially available particles, the results are easier to compare against future laboratory experiments. The number densities were approximated by assuming 0.1% concentration of the scattering particles, with a known mean diameter and spherical shape. That is,

$$N^0 = \frac{0.1\%}{\frac{4}{3}\pi(\mu_{sp}/2)^3},$$

(62)

where $\mu_{sp}$ is given in centimetres unlike in Table 3. The number densities become quite high with the utilized concentration. On the other hand, a large number of such small particles are required to fill even a small volume.

The atmospheric parameters, i.e., pressure, relative humidity and temperature are given in Table 4. The atmospheric parameters describe the absorption coefficient, which depends strongly on the humidity of the atmosphere as the water vapour is efficient absorber in the THz band. Table 4 also gives the signal properties. $T_s = 0.5$ ps long rectangular pulse is used, resulting in approximately $2/T_s$ double sided bandwidth for the signal. Utilizing constant 1 mW transmitting power over the symbol length $T_s$ results in 0.5 fJ of energy for one pulse. The above parameters are kept fixed. The integration time is variable, as a consequence, so is the SNR as it was shown in (57).

$^2$http://www.bangslabs.com

$^3$http://www.polysciences.com
Table 4. Signal and atmospheric parameters.

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pulse energy</td>
<td>0.5 fJ</td>
</tr>
<tr>
<td>Pulse length</td>
<td>0.5 ps</td>
</tr>
<tr>
<td>Pulse power</td>
<td>1 mW</td>
</tr>
<tr>
<td>Pulse width</td>
<td>4 THz</td>
</tr>
<tr>
<td>Pressure</td>
<td>1 atm</td>
</tr>
<tr>
<td>Relative humidity</td>
<td>50%</td>
</tr>
<tr>
<td>Temperature</td>
<td>296 K</td>
</tr>
</tbody>
</table>

Figure 12 gives the scattering losses as function of frequency for the four considered scattering parameters in Table 3. The scattering loss is not extremely high with the assumptions made here. However, it was shown earlier that even a small scattering loss may lead to a significant transient response.

Figures 13 and 14 show the SNR as a function of integration time $T_i$. The SNRs were obtained with (57). Fig. 14 is a close up of Fig. 13 in order to show the SNR in the vicinity of the pulse length. It can be seen that the optimal integration time is closely
following the pulse length. This is well expected as a consequence because most of the power is at the actual pulse, even with increased transient due to the scattering. The LOS path component is actually smaller in scattering conditions. This is simply due the Beer-Lambert law causing additional scattering loss on the LOS component. The gain in multiple scattering is solely due to the NLOS components.

The SNR performance is less dependent on the integration time when the scattering coefficient increases. This is expected, since the scattering allows more power to be received due to the NLOS response. This is the mechanism that increases the transient in the first place. Although the scattering increases the received signal energy, it also increases the probability of inter-symbol interference. Figure 13 especially shows that the symbol separation should be rather large in a highly scattering environment when compared, e.g., a with non-scattering case in order to avoid the ISI. This could be competed against by setting the detection threshold high enough to omit the ISI. However, this would then decrease the detection probability of the desired signal. Therefore, there is always a trade-off.

Figures 15 and 16 show the bit error performance figures with the SNRs obtained above as a function of integration time. Because of the interest in the probabilistic error
Based on the SNRs, expected error performance curves can be seen, with the best performance represented by the highest scattering case due to the largest received energy. However, even with the largest transient, the error quickly approaches the boundary of 0.5. This is due to the chi-squared distributions starting to overlap strongly as the SNR decreases. This does not encourage the use of very long integration times in the nanonetworks utilizing extremely high bandwidths. The reason is the quickly aggregating noise power, overwhelming the small transmitting power provided by the simple nanomachines.

These figures show that although the multiple scattering could theoretically increase the energy at the receiver, the overall error performance favours an integration time around the pulse length. A strong transient may increase the required integration time for maximum energy detection. The LOS response is clearly dominant due to additional absorption and scattering losses on the delayed signal components. Therefore, most of
the energy is received during the pulse duration. This explains why the minimum error is achieved with the integration times close to the pulse length. On the other hand, multiple scattering increases the probability of ISI. Thus, the scattering is undesirable, although it may theoretically give an advantage in some situations, i.e., results in more energy at the receiver. This is also due to the above-mentioned reduction of the LOS path component, which is very important component in the frequency bands where NLOS propagation reduces the signal levels significantly as it will be shown in Chapter 3.

2.7 Discussion

The main difference between the THz band and the lower bands is the strong molecular absorption. This causes deterministic frequency selective fading in the signals. This effect is very well known in the literature. A less known side effect has been predicted: transmission induced noise.

Several different approaches to transmission induced noise modelling were given above. However, the true nature of the noise component remains unrevealed. This is due to the highly complex nature of real atmospheres. Different physical phenomena behind
the absorption and emission were considered. The existing models do not describe the transmission induced noise correctly. The true noise contribution is a combination of the phenomena presented in Sections 2.2.3–2.2.6 without forgetting the traditional noise mechanism, such as radio noise and thermal noise, of the receiver. One important task for future work is to unify the models into one THz band noise model.

Another focal area in this chapter were the scattering losses in the THz band. Two multiple scattering channel models for diffusion propagation in a scattering environment were reviewed. By combining these channel models a narrow-band nanonetwork channel model for 3D environment was shown. This model was then easy to validate with Monte Carlo simulations. In addition, the time domain characteristics of the THz band were derived in the absorbing medium and further in the presence of small particles. Based on the WSSUS channel models and the existing multiple scattering models, a wideband channel model was derived. As a result, a comprehensive free space channel model for nanonetworks was obtained. Furthermore, the effect of the channel delay spread and energy detection were analysed.
Fig. 17. The probability of false alarm as a function of integration time [© 2015 ACM].

Fig. 18. The probability of missed detection as a function of integration time [© 2015 ACM].
When the scattering (LOS, or multiple scattering) is added on top the molecular absorption, the THz band is truly an interesting channel. Even in full LOS, the complex frequency dependent fading mechanisms will cause time spread to the transmitted signals. This may cause ISI if the symbols are closely paced, but this is especially the case in the presence of enough many small particles. In indoors, it is highly unlikely that the air would be dirty enough to enable scattering. However, in some outdoor applications it may occur. The best way to fight the possible ISI is to increase the symbol separation, an easy trick on which the THz band applications can afford due to large theoretical bit rates.

The increased signal transient has both pros and cons with respect to the link performance. On the other hand, multiple scattering allows more power to be received as the LOS response is summed with the NLOS components. This may be an advantage if the symbols are sufficiently far apart. However, this increases the probability of ISI if the symbols are closely separated as mentioned above. Interesting part of the multiple scattering channel models is that they are, in theory, also suitable for multipath propagation modelling. This is what they essentially are and it was shown in [57] that the multiple scattering model can be used to model a real densely obstructed environment. The only problem is that the model requires in average a uniformly obstructed environment. Regardless of this, in theory, the multiple scattering channel models presented in this thesis can be utilized in a macroscopic multipath channel.
3 THz band signal propagation: measurements

The THz band measurements are the focus area of interest in this chapter. This chapter is based on the following papers: [94–98] [© 2016 IEEE]. The measurements were conducted with a TeraView TeraPulse 4000 THz–TDS measurement device. The LOS response is easy to model based on well-known theories on absorption. Therefore, the interesting part of the measurements is to investigate the NLOS phenomena and especially the strength of the various effects, such as penetration, surface reflection/scattering, and diffraction losses.

The measurement device and the general measurement setup are depicted in Section 3.1. The NLOS measurement setups, related theories, and numerical results are presented for penetration loss, reflection/scattering loss, and diffraction loss in Sections 3.2–3.4, respectively. Finally, a discussion of the results is given in Section 3.5.

3.1 General measurement setup

The TeraView TeraPulse 4000 measurement device is based on THz-TDS. This device is originally intended for imaging applications, as well as material characterization. Still, it is perfectly suitable for channel measurements as a channel sounder. The transmitted pulses cannot be modulated, or controlled in any way. This, however, is an irrelevant aspect with respect to the channel behaviour.

As the name suggests, the measurements are made in the time domain and the frequency domain behaviour of the received signal is obtained by a Fourier transform. The transmitter and receiver antennas are excited with extremely short laser pulses (circa 100 fs). This causes the transmitter to emit a short pulse that lies in the THz frequency band. The receiver works on the same principle: while excited by the laser pulse, it is capable of receiving THz pulses in a short time window. Because the reception is limited to a short time interval, a delay line is used to capture the entire time domain pulse. This effectively means several thousand pulses to be sent for each received THz pulse. More specific parameters are given below for different measurements scenarios.

A downside of this measurement setup is the lack of phase, or polarization information of the measured pulses. The phase can be obtained with the Fourier transform,
which, on the other hand, can be manipulated by the manual adjustment of the optical delay, which has an effect on the linear phase shift seen by the Fourier transform. The initial optical delay has to be adjusted in order to capture a maximum amount of energy from the pulse, e.g., when increasing the measurement distance, the pulse shifts off the detection window and has to be manually adjusted based on the additional optical delay caused by the increased distance.

A good feature of this measurement technique is the vast bandwidth it offers. The device is capable of taking measurements from approximately 100 GHz up to 4.5 THz depending on the distance and atmospheric conditions, giving it a bandwidth of approximately 4000 GHz. This makes it possible to accurately see the frequency domain behaviour of various NLOS propagation mechanisms.

The measurements can be done with two setups: low-resolution or high-resolution (in frequency). The low-resolution setting is ideal for real-time monitoring of the data. It comprises approximately 11 fs time resolution and 4096 time samples per pulse with the setup used in this work. These numbers correspond to approximately 45 ps time window for the detected pulse and 22.2 GHz frequency resolution. The high-resolution setting used in this work has a time resolution of 100 fs and 2000 samples per measured pulse, i.e., the detection window is 200 ps long. These numbers result in a better frequency resolution of 5 GHz, but with a cost of longer measurement time because of the longer time window for detection. The high-resolution setting is in general better in material characterization because of the better resolution, but the low-resolution setting allows more averaging; thus, less noisy measurements can be made. With the setup here, low-resolution measurements are averaged over 1000 measurements and the high–resolution measurements are averaged over 50 measurements.

Fig. 19 shows the received pulse shape. It shows the vast bandwidth of the device, but also the exponentially decaying shape towards the higher frequencies. The peak response is at approximately 500 GHz. The decaying response towards the higher frequencies limits the measurement bandwidth, because the NLOS losses also increase as a function of frequency. Fig. 19 also shows the molecular absorption lines, which indeed are where the theories predict them to be at.

The measurement heads and the guidance optics for the THz beams can be seen in Fig. 20. The optics are used to gather the energy into an almost perfectly straight beam in the middle of the plates. The beam width at the sample location is approximately 2 cm. This increases the energy through the samples as the spreading loss can be minimized, but makes the adjustments more difficult. The gold–plated mirrors are very
efficient reflectors in the THz band, and thus, are common in THz band optics. This kind of a system is bulky, but allows highly accurate three-dimensional calibration due to adjustable Tx and Rx angles, as well as adjustable mirrors in the path of radiation. On the other hand, the settings of the optics are crucial to the amount of received energy. The optics add another 20 cm of link distance per side, totalling the path length in all the measurements here to 50 or 60 cm, depending on the setup. Most of the measurements are done with 20 cm separation between the Tx/Rx plates to ensure enough space to rotate the plates, or the sample. A difference to this is one of the two measurements done for the penetration loss. In the first paper [94], a 10 cm distance was used, but in the second one [97], the plate separation was increased to 20 cm in order to allow more space to rotate the samples. Otherwise all the measurements are done at 20 cm distance between the Rx and Tx plates.

As said earlier, the measurements are done either in low or high-resolution settings. The parameters for these were fixed to ones given earlier. It would be possible to adjust the resolution of the device, but for the sake of tractability, the utilized parameters
offered the best compromise between accuracy and averaging. The accuracy of the results is not affected by the utilized setup, only the resolution of the results.

The main interest of the measurements is in the path loss caused by the various NLOS phenomena. This can be calculated as:

$$\frac{S_{sc}(f)}{S_{c}(f)} = \frac{H_{sc}(f)X(f) + N_{sc}}{H_{c}(f)X(f) + N_{c}} = \frac{H_{s}(f)H_{c}(f)X(f) + N_{sc}}{H_{c}(f)X(f) + N_{c}} \approx H_{s}(f) + \frac{N_{sc}}{H_{c}(f)X(f)},$$

where $S_{sc}(f)$ is the received signal power with the sample in place and the channel, $S_{c}(f)$ is the LOS reference signal power, i.e., the straight path measurement, $H_{sc}(f)$ is the channel power transfer function of the sample and the channel, $H_{c}(f)$ is the power transfer function of the LOS reference channel, $N_{sc}$ is the noise in the sample measurement, $N_{c}$ is the noise of the reference measurement, $X(f)$ is the unknown signal power envelope, and $H_{s}(f)$ is the sample’s power transfer function. In the case of the last noise term, the noise power $N_{c}$ is assumed to be small in comparison to the reference signal power (see, e.g., Fig. 19) as a consequence of the full LOS path having significantly more power in comparison to the NLOS paths. This causes a visible noise colouring in the result figures. Therefore, this last term causes the false path gain increase in the higher frequencies. In reality, the path gain cannot be estimated outside the smooth part of the path gain. The point where the noise component starts to significantly affect the results (and render them unreliable) is easy to see due to the random white noise-like fluctuations from the 2nd term in (63) as opposed to smooth behaviour of the 1st term in (63).

3.2 Penetration on materials

3.2.1 Measurement setup for the penetration on materials

The penetration measurements were studied in two conference papers [94, 95]. The main difference in the measurements was that the first one [94] focused on the general penetration loss in the THz and was done at a low–resolution setting and with a 10 cm distance (between the plates), whereas the latter paper [95] modelled penetration coefficients of the materials. The measurements were remade for the penetration coefficient measurements with a high–resolution setting and with a 20 cm distance between the plates allow rotation of the sample. The losses were identical between the
Fig. 20. The measurements heads of TeraView TeraPulse 4000 measurement device. The figure shows the emitter (right-hand side) and the receiver (left-hand side), as well as the THz beam guidance optics [© 2016 IEEE].

Two measurements as it could be expected. However, in the latter case the penetration loss was also considered as a function of the penetration angle to the sample.

Figures for the measurement setup are given in Figs. 21 and 22. In Fig. 21 the signal path is shown for the THz radiation and the sample location. The sample is at a zero-degree angle to the Rx and Tx. Fig. 22 shows the measurement setup through the glass sample at a 50-degree angle to the Tx.

Four different materials were used for the penetration measurements: plastic, paper, hardboard, and glass. These were chosen since all are common materials in office environments as well as in any indoor location. The utilized plastic sample was styrene-acrylonitrile due to its easy availability as a regular material for plastic sheets. The paper was regular printing paper and the measurement samples were composed of a number of individual paper sheets. The density of the used paper was 80 g/m², with one sheet being approximately 100 μm thick. By utilizing 25 sheets as one sample, a sample interval of approximately 0.25 cm was achieved. The exact properties of the hardboard or glass are unknown. The hardboard is commonly used, e.g., in the supporting structures of bookshelves. The glass on the other hand was regular non-coated glass commonly found, e.g., in windows.

Table 5 gives the material thicknesses and the thickness ranges for the measurements. In the case of larger measurement thicknesses, several samples were clamped together,
Table 5. Samples used in the measurements (10 cm measurements distance).

<table>
<thead>
<tr>
<th>Material</th>
<th>Thickness per sample [cm]</th>
<th>Thickness range [cm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plastic</td>
<td>0.20 and 0.60</td>
<td>0.20 – 1.20</td>
</tr>
<tr>
<td>Paper</td>
<td>0.01</td>
<td>0.25 – 1.00</td>
</tr>
<tr>
<td>Hardboard</td>
<td>0.25</td>
<td>0.25 – 0.75</td>
</tr>
<tr>
<td>Glass</td>
<td>0.16</td>
<td>0.16 – 0.48</td>
</tr>
</tbody>
</table>

Table 6. Samples used in the penetration coefficient calculations (20 cm measurements distance).

<table>
<thead>
<tr>
<th>Sample</th>
<th>Thickness per sample [cm]</th>
<th>Thickness range [cm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plastic</td>
<td>0.20</td>
<td>0.20, 0.40, and 0.60</td>
</tr>
<tr>
<td>Hardboard</td>
<td>0.25</td>
<td>0.25, 0.50, and 0.75</td>
</tr>
<tr>
<td>Glass</td>
<td>0.16</td>
<td>0.16, 0.32, and 0.48</td>
</tr>
</tbody>
</table>

as it is also visible in Fig. 21 in the case of hardboard. A very careful clamping in the case of multiple samples was required, since air cavities between the samples would affect the results significantly. This is most likely because of the reflections on the boundaries of the samples, redirecting a part of the radiation. In Fig. 21 is the 0.75 cm sample of the hardboard, i.e., three sheets of hardboard clamped together. All the measurements were done by placing a sample orthogonally to the beam in the middle of the plates, as also visible in Figs. 21.

The penetration coefficient measurements were made for plastic, hardboard, and plastic samples and the utilized thickness ranges are given in Table 6. Three different thicknesses per sample were chosen to estimate the linearity of the path loss as the sample thickness is increased. The linear decrease in the signal power is important when estimating the path loss per unit distance. It can be calculated through derivatives of the path losses of the different material thicknesses. For the actual penetration coefficient calculation, zero-degree angle measurements were utilized, but the angular results for the path gains are also given. In the angular measurements, the orientation of the samples was varied from 0 to 60 degrees with 10 degree intervals.

Additionally, the signal shape has an effect on the lower frequencies and reliable results are obtained above 0.1 THz. Therefore, the results are given from 0.1 THz to approximately 2 THz (dependent on the material).
Fig. 21. Measurement setup showing the THz beam path from Tx to Rx. The figure also shows the 0.75 cm hardboard sample [© 2016 IEEE].

Fig. 22. The measurement setup with a 0.16 cm glass sample in 50 degree angle to the Rx [© 2016 IEEE].
3.2.2 Measurement results for penetration on materials

With the penetration measurements, there are two interesting questions for the communications engineering: 1) what is the average loss per unit distance, and 2) what is the delay through the material over unit distance. These properties are given in Table 7. The average delay was approximated from the time shift in the peak response. This is visible in Fig. 23 for the plastic samples. As the thickness of the material increases, the delay increases linearly due to the approximately constant refractive index inside the material. The refractive indices in Table 7 do not depict the general frequency dependent refractive indices, but here they describe the delay properties of the materials through

\[ n = 1 + \frac{c\varepsilon_s}{d_{\text{sample}}}, \]  

(64)

where \( n \) is the refractive index, \( c \) is the speed of light in vacuum, \( \varepsilon_s \) is the delay through the dielectric sample material (an additional delay due to the sample), and \( d_{\text{sample}} \) is the sample thickness. This equation can be derived from the speed of light in a medium found, e.g., in [74, Chapter 32.3] and based on knowledge of the delays. A more comprehensive analysis on the frequency dependent index of refraction is subject to future work. It can be seen that the delay aggregates quite quickly compared to the short THz pulses and may even slide the pulse out of the detection window. This may cause ISI, e.g., if the Rx is momentarily in NLOS to the Tx. On the other hand, the signal may be lost altogether as a consequence of the delay. For instance, the detection window of the measurement device was about 46 ps long. Therefore, in the case of thicker materials the detection window should be offset based on the delay in order to capture the signal. In the case of real communication between two terminals, it may be difficult predict the delay. This is especially the case if the environment is changes over time, e.g., a door sometimes blocks the signal. This issue would require constant monitoring of the channel in order to predict the delay and capture the signal.

The importance of the delay is irrelevant if the THz radiation is not capable of efficiently penetrating through objects made of different materials. However, as it can be seen in Table 7, THz radiation is perfectly capable of penetrating through all the samples in the experiments. For instance, plastic objects allow quite a tolerable power loss. On the other hand, communication through glass is more difficult, but not impossible if the communication distance is otherwise small.

The measurement results on the penetration loss can be seen in Fig. 24. The loss has been calculated as the difference of the time domain energy of the samples to
Table 7. Average refractive indices, losses, and delays of the materials.

<table>
<thead>
<tr>
<th>Material</th>
<th>Refractive index</th>
<th>Loss [dB/cm]</th>
<th>Delay [ps/cm]</th>
<th>Reflection [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plastic</td>
<td>1.63</td>
<td>12.47</td>
<td>21.16</td>
<td>5.36</td>
</tr>
<tr>
<td>Paper</td>
<td>1.50</td>
<td>15.82</td>
<td>16.59</td>
<td>7.76</td>
</tr>
<tr>
<td>Hardboard</td>
<td>1.57</td>
<td>24.47</td>
<td>19.16</td>
<td>7.71</td>
</tr>
<tr>
<td>Glass</td>
<td>2.60</td>
<td>35.99</td>
<td>53.50</td>
<td>9.15</td>
</tr>
</tbody>
</table>

Fig. 23. Time domain amplitude and delay through the plastic samples [© 2016 IEEE].

the reference energy. Knowing the time domain signal, number of samples, and time resolution, the energy of the sample can be calculated as

\[ E = \Delta t \sum_{n=1}^{M} |s(n)|^2, \]  

(65)

where \( M = 4096 \) is the number of time domain samples, \( \Delta t = 11.15 \) ps is the time resolution, and \( s(n) \) is the time domain signal. The unit of the time domain data is unknown (not provided by the manufacturer of the measurement device). However, most of the results can be given relative to the reference signal, which eliminates the need for the exact unit of the amplitude. Here, the reference signal is measured without the samples, thus, representing the LOS path. Therefore, Fig. 24 shows the results for
$10\log_{10}(E_r/E_s)$, where $E_r$ is reference signal energy and $E_s$ is the sample signal energy (see for instance Fig. 23). Looking Fig. 24 shows the aggregated loss through thicker samples is nearly linear. Small deviations can be caused by several reasons, such as small defects in the samples, or a slight non-uniformity of the samples. Comparing the loss results in Table 7 and Fig. 24 clearly shows quite large differences in the penetration properties between the materials. It should be noticed that the total loss is dominated by the lower frequencies, which have the most of the pulse energy.

One of the interesting effects is show in Fig. 24 and Table 7: there is an additional loss at the boundary of the sample. Some part of the effect is possibly caused by reflection and also by backscattering inside the material. However, distinction between all the possible loss mechanisms is not easy with the measurement setup in hand. Thus, all the additional losses are categorized under reflection loss. It can be seen in Fig. 24 that the first sample experiences a larger loss than the other sample thicknesses. It is easy to calculate the average loss through the materials per unit distance as done in Table 7. Based on this knowledge and the loss at the boundary of the sample, the average reflection loss can be calculated. This is visible in Table 7. It is not surprising that the plastic suffers least from reflection loss as it allows the lowest penetration loss of all the samples here. Also, the glass as the highest attenuator also reflects the most energy.
However, as the hardboard and paper samples show, the reflection loss cannot be directly
deduced from the penetration loss. The hardboard causes a larger loss, but the reflection
loss is quite identical to the paper samples. The reason for these can only be guessed at
this stage. However, since the reflection loss here is just a combination of all the possible
other losses, e.g., the internal structure of the material has an effect on the scattering,
and the surface properties, such as roughness, have effect on the reflections. Therefore,
the two materials may have very different properties with respect to the penetration
performance while having a similar collective loss on the other loss mechanisms.

Furthermore, the frequency responses of the materials with the measured material
thicknesses are shown in Figs. 25 – 28. This provides further details on where the loss
occurs in the frequency domain. These figures are obtained by calculating the frequency
domain energies of the measured samples. These are then normalized with the reference
signal frequency domain energy. This way the figures depict the frequency response
inside the material as normalization with the reference signal equalizes the atmospheric
effects away. However, the atmospheric effects cannot be erased entirely, since the
samples have a finite thickness. This causes small difference in the composition of the
medium between the Tx and the Rx besides the sample itself. Also, the atmospheric
conditions may vary slightly between the measurements.

From the below figures, it can be seen that the penetration loss mainly filters the
higher end of the frequency response. A similar behaviour was reported for polymers
in [61]. As the thickness of the sample is increased, the loss on certain frequencies
increases exponentially, but linearly on a dB-scale because of the approximately similar
loss per sample. As it can be predicted and seen in Fig. 25, plastic allows much higher
frequencies to penetrate in comparison to, e.g., glass in Fig. 28. The same trend is
shown in all figures, as the attenuation per centimetre increases, the general frequency
response characteristic becomes narrower. Again, this behaviour is due to many reasons,
one being that the higher frequencies are more prone to scattering losses because the
scattering loss is strongly frequency dependent as seen earlier and will be shown below
for the surface scattering measurements. One interesting observation can be made on the
paper samples: while providing rather good penetration, paper attenuates the signal
very strongly around 900 GHz. This is clearly visible in the case of thinner samples.
This may be caused by absorption of energy to the material similarly as the molecular
absorption causes absorption lines in the atmosphere. Of course, it is also possible that
paper scatters radiation more efficiently at certain frequencies dependent on, e.g., the
Fig. 25. Frequency responses of the plastic samples [© 2016 IEEE].

Fig. 26. Frequency responses of the paper samples [© 2016 IEEE].

internal structure of the material. However, the exact reason for this behaviour is not known and requires further investigation in the future.
Fig. 27. Frequency responses of the hardboard samples [© 2016 IEEE].

Fig. 28. Frequency responses of the glass samples [© 2016 IEEE].

The frequency dependent loss is somewhat problematic considering the efficient utilization of the higher end of the THz band for communications. There is a risk
of losing the signals at the higher end of the THz. For instance, the lower end of the THz band below 200 GHz is less sensitive to penetration loss. However, this is also the least interesting band in the THz regime. The most interesting part is the above one THz region. It should also be remembered that the measurement device is capable of measurements from around 100 GHz to about 4 THz with an exponentially decaying amplitude towards the higher extreme. Therefore, a perfectly rectangular signal extending over the entire THz band could possibly preserve the high frequency signals better in NLOS conditions. However, they will suffer much more in NLOS compared with the lower end of the THz band as shown above.

Fig. 29 shows the path losses at angles from 0 to 60 degrees at ten–degree intervals. The measurements were made for one layer of each material (see Table 6). Increasing the incident angle increases the path loss through the sample. The reason is the increased path length inside the sample. This issue requires more investigation in future work, since estimating the true path length inside the material requires a purposely built setup for the refractive index measurements, whereas the measurement setup herein focused on general path loss characteristics.
Based on Figs. 25–28, the path gain drops nearly linearly on the decibel-scale when the thickness is increased. This allows an estimation of the penetration coefficients as differences in the path gains. By utilizing this approach, any other possible effects are removed, e.g., reflection loss at the surface of the sample. A wanted feature is the penetration coefficient to be compatible with the Beer-Lambert law, as in the case of molecular absorption and small particle scattering loss. Thus, the penetration coefficient $\kappa_p(f)$ becomes

$$\kappa_p(f) = -\ln \left( \frac{10^{dH_{dB}(f)/10}}{10^{dH_{dB}(f)/10}} \right) / r,$$

(66)

where $dH_{dB}(f)$ is the difference in the path gains on the dB-scale, and $r$ is a thickness of a single sample. The first two thickness values per sample have been utilized to estimate the penetration coefficient in order to increase the bandwidth of the coefficient. The larger the thickness, the narrower the bandwidth of the penetrated signal, and the less information there is at the higher frequencies. The resultant penetration coefficient has a unit of cm$^{-1}$, which can then be directly plugged into the Beer-Lambert law to get path gain, or transmittance $\tau(f)$, at an arbitrary distance $r$ inside the material as

$$\tau(f) = \exp(-\kappa_p(f)r).$$

(67)

The penetration coefficients of the measured materials are given in Fig. 30 and the transmittances through a 0.1 cm thick sample in Fig. 31. There are two observations to be made on these figures: i) the actual losses are in general smaller than the measured values because the reflection loss on the surface of the material is eliminated, and ii) the reflection coefficients are only valid in partial frequency bands, i.e., where there is no noise. For glass and wood, the limit is at approximately 0.6 and 0.7 THz, respectively, and for the plastic it is at around 1.4 THz. The lower bound for all the materials is at approximately 0.1 THz, which comes from the lack of accuracy of the measurement device at the lower frequencies. As it can be seen in (63), the shape of the noisy signal comes from its normalization by the reference signal. Therefore, the actual transmittance above the higher boundaries should be considered to be close to zero. The main observation here is that penetration is possible, but only in the low THz band. Penetration efficiently filters out the high frequency components, as also seen above. A partial reason for the narrow bands is the limited signal bandwidth of the measurement device. Although the signal dependency can be erased, as in (63), in reality it is still there. The reason is that there is no information on the channel where the reference signal approaches the noise level. However, it is clear that the penetration does not
preserve high frequency components, which makes the full THz band utilization difficult in NLOS.

3.3 Reflection and scattering on surfaces

3.3.1 Measurement setup for reflection and scattering on surfaces

Rough surface scattering measurements were made with two different set of samples, one for sandpapers [96], and one for common material found in indoors [97]. The sandpapers were utilized because of the controlled and known grit sizes in the samples. The common materials, on the other hand, were measured to learn the behaviour of some uncontrolled, but common indoor materials.

A measurement setup shown in Fig. 32 was utilized for the sand paper measurements. The measurements were performed by keeping the receiver fixed at a 45–degree angle to the mirror. The transmitter orientation to the mirror was changed with 10 degree intervals. In the reflection measurements, a zero–degree offset is a 45–degree angle of the transmitter to the mirror, i.e., the receiver is straight at the reflected path of the transmitted beam. The offset is increased towards the normal of the surface.
Fig. 31. The transmittances of the three materials through a 0.1 cm thick sample [© 2016 IEEE].

The basic idea behind the measurements was to use a gold plated mirror as a “perfect” reflector, mimicking a perfectly smooth surface. By placing the sandpaper samples on top of the mirror ensures no any unwanted scattering from the background material. The gold plated mirror was proven to be very good reflector at the THz band. Fig. 33 shows the time domain signal through a straight path versus through the reflected path through the mirror (see Fig. 32). The amplitudes are nearly identical. The reflected path has a 180 degree phase difference to the straight path because of the reflection, but otherwise the amplitudes are close to each other. Calculating the energy differences between these paths shows that the mirror reflects approximately 98% of the THz radiation, which is perfect for attempting to use the mirror as a perfectly reflective background for the sandpaper.

Fig. 32 also shows the setup with a rough surface test sample. Sandpapers were chosen due to their well-controlled manufacturing process controlled by standardization. Thus, they have known mean grit sizes given by the International Organization for Standardization (ISO) (and Federation of the European Producers of Abrasives (FEPA)) [139, 140]. The details of the test samples are given in Table 8. The samples were tightly
attached to the surface of the mirror in order to avoid any air cavities between the mirror and the sandpaper. Air cavities were noticed to have quite a large effect on the results.

Table 8. Test sample indexing and sandpaper roughnesses.

<table>
<thead>
<tr>
<th>Test sample #</th>
<th>Sandpaper</th>
<th>Median Grain Size [µm]</th>
<th>Standard</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Reference</td>
<td>n/a</td>
<td>n/a</td>
</tr>
<tr>
<td>1</td>
<td>P2000</td>
<td>10.3</td>
<td>[139]</td>
</tr>
<tr>
<td>2</td>
<td>P1000</td>
<td>18.3</td>
<td>[139]</td>
</tr>
<tr>
<td>3</td>
<td>P320</td>
<td>46.2</td>
<td>[139]</td>
</tr>
<tr>
<td>4</td>
<td>P180</td>
<td>82</td>
<td>[140]</td>
</tr>
<tr>
<td>5</td>
<td>P80</td>
<td>201</td>
<td>[140]</td>
</tr>
<tr>
<td>6</td>
<td>P60</td>
<td>269</td>
<td>[140]</td>
</tr>
</tbody>
</table>

Fig. 34 shows the measurement setup for the common indoor materials. The samples were placed in the zero position between the measurement head plates and adjusted to
be at a perfect 90–degree angle in both horizontal and vertical directions. In Fig. 34, one can see the utilized glass sample in place with the both measurement heads at 45–degree angle to the sample.

The rest of the samples, besides the glass (visible in Fig. 34), are shown in Fig. 35. These samples were chosen because they are common in any indoor or outdoor locations. They included aluminium, glass, plastic, hardboard, and concrete. Furthermore, these samples gave quite a good overview of the different roughnesses found in indoor locations. The aluminium sample was moderately polished unfinished aluminium sheet with a rather smooth surface. The glass sample was the same as earlier. This was the smoothest sample among all the samples along with the plastic sample, which was used previously in the penetration measurements. The second roughest sample was hardboard. Because of being composed of compressed wood fibres, it has a relatively smooth surface with a moderately rough finish. The concrete sample was by far the roughest sample among the considered ones. It had a straight face, but otherwise it was unpolished, non-coated concrete. These samples already give quite good overview of different materials and roughnesses that can be found in indoor, but also in outdoor locations.
The angle of both the Tx and Rx was varied from 35 to 70–degrees. The angle is relative to the tangent of the surface. Any lower angle would not have been possible for all the samples because of the size of the samples limiting the lower angle, but also because of the device’s two centimetre beam width. Therefore, the whole THz beam would have not hit the surface at lower angles. The higher angles were, on the other hand, limited by the size of the measurement head plates. Higher (and also lower) angles would have been possible, but not for all the Tx/Rx angle combinations.

### 3.3.2 Theory on reflection and scattering on surfaces

The rough surface scattering theory is familiar from multiple rough surface scattering papers, such as [62, 64–68]. The theoretical path gain $L(f, \sigma_t, l, \theta_1, \theta_2, \theta_3)$ for a rough
surface with infinite conductivity is given by [66, 67]

\[
L(f, \sigma_h, l_c, \theta_1, \theta_2, \theta_3) = e^g \left( \kappa^2 + \frac{\pi l_c^2 F^2}{A} \sum_{m=1}^{\infty} \frac{g^m}{m!} e^{-v_{xy}^2 l_c^2 / 4m} \right),
\]

where

\[
\kappa = \text{sinc}(v_x l_x) \text{sinc}(v_y l_y),
\]

\[
v_x = k(\sin(\theta_1) - \sin(\theta_2) \cos(\theta_3)),
\]

\[
v_y = k(-\sin(\theta_2) \sin(\theta_3)),
\]

\[
v_{xy} = \sqrt{v_x^2 + v_y^2},
\]

\[
F = \frac{1 + \cos(\theta_1) \cos(\theta_2) - \sin(\theta_1) \sin(\theta_2) \cos(\theta_3)}{\cos(\theta_1)(\cos(\theta_1) + \cos(\theta_2))},
\]

\[
g = k^2 \sigma_h^2 (\cos(\theta_1) + \cos(\theta_2))^2,
\]

where \(l_c\) is the surface correlation length, \(\sigma_h\) is the standard deviation of the variations in surface height, \(\theta_1\) is the elevation of the incident radiation, \(\theta_2\) is the elevation of the scattered radiation, \(\theta_3\) is the azimuth of the scattered radiation, \(g\) is the Rayleigh roughness factor, \(\kappa^2\) is the specularly reflected component, \(l_x\) and \(l_y\) give the fractional area of the surface under investigation (the surface can be broken into a grid of areas \(A = l_x l_y\), with \(l_x, l_y \gg \lambda\), where \(\lambda\) is the wavelength [67]), and \(k\) is the angular wave number. More information on these parameters and rough surface scattering theory can be found in the references [62, 64, 66–68].

Fig. 36 shows the path gain at one THz frequency as a function of elevation angles \(\theta_1\) and \(\theta_2\). The azimuth angle \(\theta_3\) is assumed to be zero. The area on the surface is assumed to be \(l_x l_y = \pi\) with equal \(l_x\) and \(l_y\), which is the approximate surface area of the THz beam. Furthermore, the surface correlation length is assumed to be \(l_c = 0.01\) cm and the surface height distribution \(\sigma_h = 10^{-4}\) cm. Thus, the example surface is a very smooth with a high number of surface variations. Very similar patterns can be achieved with the measurements, e.g., Fig. 41 integrated over the frequencies. The main difference stems from the fact that the measurements have a beam diameter of approximately two centimetres at the sample, which widens the specular response. Also, the exact surface properties are are unknown for the samples for now, as they require further investigation with an electron microscope in the future to accurately predict the corresponding surface height distributions and correlation lengths. Despite the unknown
parameters, the theory matches the measurement results relatively accurately, making the utilization of the above theory in the future perfectly possible in the case of wide band THz scattering measurements.

### 3.3.3 Measurement results for reflection and scattering on surfaces

The main results of the measurements for the sandpapers are given in Fig. 37. The left-hand side column of figures gives the sample energies in arbitrary units for the given offset angles. The middle column of figures shows the energy differences for the case with no sandpaper, only the mirror at the specified offset. The last column of figures shows the sample energies compared to the reflected path for the mirror only and with no offset (corresponding to LOS). This is the extra loss to the signal due to reflection and scattering as compared to the LOS path. The test samples are named as in Table 8.

These energy figures reveal that the behaviour of the scattering is approximately what was expected: the rougher the surface is, the higher is the diffuse scattering contribution, i.e., the contribution coming outside the specularly reflected component increases as the surface roughness increases. However, the energy levels drop sharply as
the specular component is lost. The non-specular directions give roughly four orders of magnitude less power than the specular direction as it can be seen in the left-hand size and right hand size figures. Still, the roughest sandpaper retains energy quite well regardless of the offset. This is caused by the larger scattering at the surface of the sample. The figures also show that when the offset is close to the reflection path, a small amount of scattering increases the received energy, whereas large scattering is undesirable. However, already at 20–degree offset the roughest two samples start to give the highest amount of energy, while sandpaper P1000 gives relatively higher energy figures compared to the neighboring samples. The behaviour of sample #2 is most likely because of better scattering properties on the higher frequencies, as shown in Figs. 38–40: this clearly has a wider frequency response compared with the other samples. For instance, the roughest samples give most of the energy at the lower frequencies. The full reason for the wider response in the case of sample #2 is unknown, but it is possible that the grain size allows more efficient scattering above one THz. A measurement setup
specific problem is unlikely, since the same behaviour was present with all the offsets, different samples, and in an earlier measurement setup. The measurements also showed some unexpected behaviour in the case of 30–degree offset. This offset gave slightly higher energies in general compared to offsets of 20 and 40 degrees. This may be a measurement setup related issue. However, the exact reason is not known since the measurements were made one sample at the time and the anomaly always occurred at the same offset. However, the effect of this anomaly is small compared to the overall results. All in all, the reflected paths are quite good with all the considered samples compared with the diffuse scattering cases (offsets 10–40 degrees). The losses are at a sufficiently small level that communication through the reflections is possible. However, communication through the diffuse scattering field is quite difficult, unless the overall distance is very small, as it was in these measurements.

Figs. 38–40 show the frequency domain behaviour of the sample energies in arbitrary units on the dB-scale with the considered offsets. The figures are given two-dimensional format in order to show the differences more clearly. Two important things can be seen in these figures: i) as the offset is increased, the peak frequency response of the smoother samples moves to higher frequencies, ii) the rougher surfaces start to give more energy at higher angles, but the peak response is at lower frequencies. This is a well expected result, as the scattering is in general strongly frequency dependent, and the size of the scattering particles has an impact on the scattering. For instance, similar behaviour could be expected based on the rough surface model presented, e.g., in [67]. Still, it should be remembered that the energy contribution from diffuse scattering is small compared with the reflection paths.

The measurements on the common materials showed similar behaviour to the ones for the sandpapers. The frequency domain energies as a function of the receiver angle are given in Figs. 41 and 42 for aluminium and concrete, respectively. The Tx angle is fixed at 50 degrees. These samples were chosen for the figures, since they represent the two extremes of the measurements: aluminium was the strongest reflector and concrete was the strongest scatterer. The other samples more or less represent the aluminium response, with some differences as it can be seen in the case of the path gain figures below.

The aluminium gives a quite representative result of what was expected, and what the general observation was of the measurements: the largest contribution comes from the reflected path, with minor response on the diffuse scattering field. This is also true for the concrete. However, it gives far less energy on the reflected path, but also
Fig. 38. Frequency domain energies with zero degree offset to the receiver in arbitrary unit (in dB) with respect to the samples [© 2016 IEEE].

Fig. 39. Frequency domain energies with 20 degree offset to the receiver in arbitrary unit (in dB) with respect to the samples [© 2016 IEEE].
Fig. 40. Frequency domain energies with 40 degree offset to the receiver in arbitrary unit (in dB) with respect to the samples [© 2016 IEEE].

Fig. 41. The measured frequency responses for aluminium at the 50 degree Tx angle as a function of the receiver angle [© 2016 IEEE].
the most energy in the diffuse scattering directions. What can also be seen is that the lower frequencies contribute much more at higher angle offsets than the higher frequencies. The first reason is that the scattering increases with frequency. Thus, the higher frequency components should have a flatter response over the receiver angles. This leads as to the second reason: the frequency response would be lower in any case at the higher frequencies because of the signal shape. This causes weak responses to be easily hidden below the noise level. However, this is a very interesting result, as it shows the THz band encompasses both strong reflected components, as well as the perceptible diffuse scattering field. Theoretically there is no reason why either of these could be used for communications in the NLOS directions.

Figs. 43 to 46 illustrate the path gains of the considered materials at a fixed Tx angle of 35 degrees at the Rx angles of 35, 40, 50, and 65 degrees, respectively. The first two angles have either full, or partial visibility of the Tx at Rx through the reflected paths, whereas the last two figures depict a situation in the diffuse scattering field only. Note the sharp spiked features in the responses are caused by molecular absorption: where the reference signal comes close to the noise level, there are sharp spikes after
Fig. 43. The path gains of the materials as a function of frequency with a Tx angle of 35 degrees and an Rx angle of 35 degrees [© 2016 IEEE].

Fig. 44. The path gains of the materials as a function of frequency with a Tx angle of 35 degrees and an Rx angle of 40 degrees [© 2016 IEEE].
Fig. 45. The path gains of the materials as a function of frequency with a Tx angle of 35 degrees and an Rx angle of 50 degrees [© 2016 IEEE].

Fig. 46. The path gains of the materials as a function of frequency with a Tx angle of 35 degrees and an Rx angle of 65 degrees [© 2016 IEEE].
division. Some spikes are partially caused by division with a close to zero value, but also because of a slight mismatch between absorption features between reference and the sample measurements. The THz band is sensitive to humidity changes because of the absorption of energy in water molecules. The first figure (Fig. 43) shows the reflection properties up to 4 THz and due to the reflected path this relatively gives much energy even at the higher frequencies. The rest of the figures are cut from the 2 THz band as a consequence of the desired signal falling below the noise power. Because the figures are calculated with (63), noise colouring similar to the penetration measurements can be seen in the regime where the desired signal power falls below the noise level.

Some interesting features can be seen in these figures: aluminium is the strongest reflector, but not on all frequencies. The glass and plastic partially exceed the aluminium’s response between 1.5 and 2 THz. The reason for this is most likely the very smooth surface of the glass and plastic in comparison to the aluminium. The aluminium seems to have strong diffuse scattering between 1.5 and 2.5 THz because absorption on the non-coated aluminium is unlikely. This could be caused, e.g., by the refractive index causing weaker reflections at certain frequencies. The refractive index is both frequency dependent, and linked to the reflectivity of the material [74, Chapter 35.4]. Still, the overall energy advantage goes to the aluminium because it does not allow any penetration, whereas the glass and plastic do.

As soon as the reflected components are lost, the concrete is the only significant contributor to the energy. The path loss is very high, but given a short enough distance, communication solely through the diffuse scattering field is theoretically possible. On the other hand, communication through reflected paths is very plausible due to path gain varying between 0 and -20 dB below 2.5 THz for aluminium, glass, and plastic, and being at a somewhat tolerable level on the other two materials also. As it can be seen in Fig. 46, as well as in Fig. 42, very rough surfaces may cause some problems even in the reflected paths, but especially below 0.5 THz the losses are small enough to ensure tolerable path loss.

These results give the same overall view as the sandpaper measurements: the reflected paths have quite reasonable power losses, whereas the diffuse scattering field may theoretically be used for communications at short distances (with rougher materials). This may cause interference in co-existing links, but may also be used to establish connections if LOS path are not available. The reflected path actually has the smallest power loss of all the measured NLOS effects, including the diffracted path that will be discussed in the next section.
3.4  Diffraction

3.4.1  Measurement setups for diffraction measurements

Measurement setup for knife-edge diffraction

The knife-edge effect was measured with the measurement setup shown in Fig. 47. A two millimetre thick aluminium sheet was placed at the centre of the beam, at a distance of ten centimetres from the emitter. The receiver was then moved at five–degree intervals from zero degree to 70 degrees keeping the distance fixed to the centre point. Thus, from the theoretical point of view, angle $\theta$ was variable, whereas the distances $d_1$ and $d_2$ were fixed at ten centimetres (see Fig. 50). A two millimetre thick aluminium sheet was chosen as the knife-edge as it is perfectly opaque at the THz frequencies while being rather thin. The finite thickness of the knife-edge requires adjustments to the corresponding theories and will be discussed in the context of knife-edge theory and the numerical results.

Measurement setup for slit diffraction

The measurement setup for the slit diffraction is shown in Fig. 48. The slits screen was composed of copper plated plastic. The copper plating made sure the material did not allow THz radiation to pass through. The figure shows the two-slit case. Illustrations on the slit screens for the single and double slit measurement cases are also given in Fig. 49. The slit width $a_{\text{slit}}$ was two millimetres in both cases to ensure a good diffraction pattern on the THz band. In the case of two slits, the slit separation $d_{\text{slit}}$ was two millimetres as well. The angle was varied similarly as in the knife-edge measurements, and similarly, the distances from the THz emitter and receiver to the sample were kept fixed at ten centimetres.

3.4.2  Theories on knife-edge and slit diffraction

All the theories behind the diffraction phenomena are very well established. Ideal materials are assumed for the theories. Surprisingly, and as it can be seen in the numerical results in Sections 3.4.3 and 3.4.3, that even these basic level theories offer a very good fit to the experimental results.
Fig. 47. The Measurement setup for the knife-edge measurements. The receiver is at a 30–degree angle to the knife-edge [© 2016 IEEE].

Fig. 48. The Measurement setup for the slit diffraction measurements. The two slit sample is in place for this figure and the receiver is at a 30–degree angle to the sample [© 2016 IEEE].

Knife-edge diffraction

Knife-edge diffraction is caused by an obstacle, and especially the edge of the obstacle operating as a secondary source for the radiation. This allows the radiation to reach the...
shadow region of the obstacle. The knife-edge path gain is described by [69–73]

\[ L(\theta) = \frac{(1 - C(\nu) - S(\nu))^2 + (C(\nu) - S(\nu))^2}{4}, \]  

(75)

where \( \theta \) is the diffraction angle given in Fig. 50, \( C(\nu) \) and \( S(\nu) \) are the Fresnel integrals given by

\[ C(\nu) = \nu \int_0^\nu \cos \left( \frac{\pi s^2}{2} \right) ds, \]  

(76)

\[ S(\nu) = \nu \int_0^\nu \sin \left( \frac{\pi s^2}{2} \right) ds, \]  

(77)

and \( \nu \) is the Fresnel parameter given by

\[ \nu = h \sqrt{\frac{2}{\lambda} \left( \frac{1}{d_1} + \frac{1}{d_2} \right)}, \]  

(78)

where \( h \) is the height of the object measured from the LOS path as shown in Fig. 50 (see Fig. 50). Based on the measurement setup, the length of the LOS path \( x \) can be calculated with the law of cosines (see Fig. 50) as

\[ x = \sqrt{d_1^2 + d_2^2 - 2d_1d_2 \cos(\pi - \theta)}. \]  

(79)

Angle \( \alpha \) can be obtained from the LOS path length as

\[ \alpha = \cos^{-1} \left( \frac{d_1^2 + x^2 - d_2^2}{2xd_1} \right). \]  

(80)
Finally, the height $h_s$ is calculated with

$$h_s = d_1 \tan(\alpha).$$

(81)

In the actual measurement setup a knife-edge of finite thickness is utilized. This causes the above theoretical prediction to differ from the measurement results. The finite thickness leads to more attenuation in the shadow region of the diffraction. Therefore, an additional attenuation introduced by a round obstacle is used. This was noted to rectify the theoretical model towards the measured values, although, the knife-edge in the measurement was rectangular. The edge in the measurements can be thought of as two consecutive infinitely sharp knife-edges beside each other, separated by a distance $2r_{ke}$, where $r_{ke}$ is a diameter of an imagined circle between the edges. Thus, it is a half of the knife-edge thickness. With this assumption the additional loss to the ideal knife-edge on the dB-scale is given by [72]

$$L_{dB}(m, n) = 7.2m^{1/2} - (2 - 12.5n)m + 3.6m^{3/2} - 0.8^2,$$

(82)

for $mn \leq 4$, and

$$L_{dB}(m, n) = -6 - 20\log_{10}(mn) + 7.2m^{1/2}$$

$$- (2 - 17n)m + 3.6m^{3/2} - 0.8^2,$$

(83)

for $mn > 4$, where

$$m = r_{ke} \left( \frac{1}{d_1} + \frac{1}{d_2} \right) / \left( \frac{\pi r_{ke}}{\lambda} \right)^{1/3},$$

(84)

and

$$n = \frac{h_s}{r_{ke}} \left( \frac{\pi r_{ke}}{\lambda} \right)^{2/3}.$$

(85)

Then the total path gain (in linear scale) is given by

$$L_{tot}(\theta) = L(\theta) 10^{-L_{dB}(m,n)/10}.$$  

(86)

**Slit diffraction**

Slit diffraction is a very well-known and familiar in optics. A diffraction pattern is caused by the slit acting as a “virtual” source for radiation (similarly as in the case of the knife-edge, where the edge acts as a source). In the case of multiple slits, multiple sources causes more defined diffraction patterns as it will be shown in the numerical results in Section 3.4.3.
The slit diffraction theory for a single slit is given by [74, Chap. 36.3]

$$\frac{I}{I_0} = \left( \frac{\sin(\beta/2)}{\beta/2} \right)^2,$$  \hspace{1cm} (87)

where $I$ is the received radiation intensity, $I_0$ is the incident radiation intensity, and

$$\beta = \frac{2\pi d_{\text{slit}}}{\lambda} \sin(\theta),$$  \hspace{1cm} (88)

where $d_{\text{slit}}$ is the slit width, and $\theta$ is the diffraction angle (as in above, and in Fig. 50).

The slit diffraction attenuation for a double slit is given by [74, Chap. 36.4]

$$\frac{I}{I_0} = \cos\left( \frac{\phi}{2} \right)^2 \left( \frac{\sin(\beta/2)}{\beta/2} \right)^2,$$  \hspace{1cm} (89)

where

$$\phi = \frac{2\pi d_{\text{slit}}}{\lambda} \sin(\theta),$$  \hspace{1cm} (90)

where $d_{\text{slit}}$ is the slit separation distance.

### 3.4.3 Measurement results for the knife-edge and slit diffraction

The measurement results and comparisons to respective theories are shown in this section. The knife-edge measurements are presented first followed by the slit diffraction measurements. The system geometry related parameters are given in Table 9.
Table 9. The key properties of the measurement samples in diffraction measurements.

<table>
<thead>
<tr>
<th>Quality</th>
<th>Quantity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Distance $d_1$</td>
<td>10 cm</td>
</tr>
<tr>
<td>Distance $d_2$</td>
<td>10 cm</td>
</tr>
<tr>
<td>Radius $r_{ke}$</td>
<td>0.1 cm</td>
</tr>
<tr>
<td>Slit width $a_{slit}$</td>
<td>2 mm</td>
</tr>
<tr>
<td>Slit separation $d_{slit}$</td>
<td>2 mm</td>
</tr>
</tbody>
</table>

**Measurement results for Knife-edge diffraction**

Fig. 51 shows the measured knife-edge path gains (in dB) as a function of frequency and the diffraction angle. The path gain was obtained by comparing the measurement results against a reference signal measured at full LOS (e.g., shown in Fig. 19). The theoretical path gain is given by (86), and shown in Fig. 52 for the parameters discussed in the theory section and reflecting the measurement setup given above. The measurement results agree very well with the theory for the round knife-edge path gain given in (86). Both the theory and the measurements are suppressed to give a minimum path gain of -70 dB in order to focus on the transmission in the shadowed regions. As it was discussed in the last section, the measurement results for the frequency bands below 100 GHz are not utilizable due to the pulse shape. Otherwise the theory and the empirical data converge. Another angle on the results is given in Fig. 53, where the path gains are given for three frequencies as a function of the diffraction angle. Also this figure shows a great fit between the measurement data and the theoretical model.

The excellent fit is an expected result as the theories are fundamental results for any frequency band. At the same time, this is an interesting result from the viewpoint of multiple possible NLOS propagation methods for the THz band communications. Adding to these, a very interesting contribution is provided by the diffraction, offering a possibility of reception in the shadow region, so that the THz band has quite an interesting multi-path environment. However (as expected), the diffraction response quickly approaches the noise level as the angle is increased. Therefore, the expected operational region for communication is at shallow angles behind objects. As it can be seen in Fig. 53, 10 degree angle still gives reasonable path gain, but the response decays exponentially as the diffraction angle is increased.
Measurement results for slit diffraction

The slit diffraction measurement data for the path loss (in dB) in one-slit case is given in Fig. 54, and for the two-slit case in Fig. 56. The corresponding theoretical values are given in Figs. 55 and 57. The figures are limited to a minimum path gain of -50 dB to better show the diffraction patterns. As it can be seen, the measurement data is similar to the theoretical values. The measurement data is noisier and the finite beam diameter distorts the data, because the two–centimetre wide beam allows reception at multiple angles at the same time. The measurement results are effectively obtained after a two–dimensional filter (due to the beam diameter), leading to blurry results. A part of the distortion comes from the non-ideal measurement setup. However, the
corresponding diffraction features can be seen. Furthermore, attenuation at the maxima are almost spot on within the tolerable error range. This can also be seen in Figs. 58 and 59, which show the frequency responses of the single and double slit cases, respectively, for a 40–degree diffraction angle. The high angle was chosen to better illustrate the diffraction pattern, which is quite distorted at the lower angles.

The two-slit case suffers visually a little bit from the interpolation necessary to make the figure informative. This creates a false secondary grid opening from the bottom left to the upper right. In reality this is not in the data, but is an artefact caused by the interpolation (and can be seen in Figs. 58 and 59). A secondary issue with the two-slit case is the measurement data, which looks more like a single slit pattern. The maxima are in their correct places, but the maxima strengths are more uniform when compared to the theoretical values. They show a similar pattern to the single slit case (Fig. 55). This can also be seen in Fig. 59. This requires further investigations in the future. However, the most probable reason lies in the previously mentioned finite beam diameter, hiding the fine structure of the diffraction pattern. The reason may also be due to a similar effect seen in the case of the knife-edge diffraction: the finite depth of the sample causes some unexpected behaviour. However, two–slit diffraction is not probable in real communication systems and the knife-edge effect is the most important with respect to modelling the NLOS paths in the THz band.

The overall picture of slit diffraction indicate lower attenuation in the knife-edge effect, with somewhat reasonable losses throughout the studied band. This is especially true for the single slit case, which has rather narrow minima. The two–slit case widens
the minima and therefore causes a larger loss in the shadow region. Still, both of the slit diffraction cases would allow communication in the shadow region, given the transmitting power is high enough.

3.5 Discussion

The measurement results for the penetration, reflections and scattering from surfaces, and diffraction were presented. All of these NLOS phenomena cause low enough attenuation that they could theoretically be used for communication. The overall message is that the THz band offers a much richer multipath environment than many have anticipated. This also highlights a need for further investigations in terms of the measurements.
Furthermore, the theoretical aspects need to be emphasized in the future work to better describe the channel behaviour in different environments.

Each of these NLOS effects cause time delayed responses at the receiver due the path length differences as well as the dielectric differences in the propagation paths. Therefore, the NLOS effects will cause additional interference to the coexisting links and ISI. This is because of the tight detection window length requirements set by the short pulses in the THz band, causing the NLOS paths to potentially spread the symbol energy over multiple symbol time slots. Still, the contribution of the multipath components depends on several factors, such as, the desired/interfered link in LOS or NLOS conditions, as well the distances, and antenna directivities. The NLOS paths have considerably less energy than the LOS links, therefore decreasing the impact of
the NLOS paths if the LOS path is available. On the other hand, NLOS paths can be used to establish a link between two nodes if LOS path is not present. Based on the measurements here, this can also be done in the THz band, despite the large losses.
4 Stochastic geometry for THz networks

The previous sections focused on individual channel effects. These can be used in modelling the individual links between the users. This chapter will focus on a network of nanodevices or other THz devices. The chapter is based on the following paper: [99] [© 2017 IEEE]. Network level interference can be modelled with stochastic geometry. The main focus here is on the mean interference, but the outage probability is also derived. The channel is assumed to be LOS, without any additional propagation effects. However, they can be included in the models by incorporating them into the absorption coefficient as done earlier.

The system model and the channel used here are given in Section 4.1. The mean interference is derived in Section 4.2. This section also contains the derivations of the outage probability and the directional antennas. Numerical results are given in Section 4.3 and the models and results are discussed in Section 4.4.

4.1 System model

4.1.1 System geometry

An illustration of the system model is given in Fig. 60 with three interfering transmitter-receiver node pairs as an example in addition to the receiver of interest at the origin. Because of the random network, the results hold for any point in space, but placing the studied receiver at the origin makes the notations simpler. The desired transmitter is always at a distance $r$ from the origin and located on the horizontal axis in the positive direction. The interfering nodes are randomly and evenly distributed around the origin.

The positions of the nodes are assumed to follow a Poisson point process that has very important characteristics making it possible to model the interference as a homogenous PPP [78], which

1. has evenly and homogenously distributed nodes.
2. is stationary, which means that the PPP is independent of translations.
3. is simple in the sense that there cannot be multiple points at the same location.
4. is isotropic, which means that the PPP is independent of rotations.
The above properties make the below derivations feasible. For instance, the point at the origin is studied without a loss of generality based on the second item. With the directional antennas, a simulation model is utilized where the coordinate axes are rotated with respect to the transmitting antenna directions. The corresponding stochastic antenna gain is independent of the simulation rotations, because the PPP is independent of rotations. The first point was already mentioned: the interfering nodes are assumed to be evenly distributed around the origin. The third fact mainly says that $N$ random points can be utilized to model an $N$ node network. The Poisson point process further has a couple of interesting features, which, however, are not utilized in this thesis: an independent thinning of a PPP is PPP, and superposing multiple PPPs is PPP\[78\]. Technically the thinning property is used, since an ALOHA network is assumed (similarly as, e.g., in [77] and many others), i.e., all the nodes transmit randomly with a probability $p$. In other words, this is a $p$-fold thinning to the $N$ node network, i.e., there are effectively $pN$ nodes transmitting at the same time. Thus, the density of the network $\rho_n$ is effectively $pp_n$\[77\].
4.1.2 Channel model

The commonly used path loss model in the THz band was shown in (1) and will be denoted in $\mathbb{R}^{d_s}$ space as \cite{10, 17, 141}

$$l(r) = \frac{\exp(-\kappa_a(f)r)}{c_{d_s}d_s^{d_s-1}},$$  \hspace{1cm} (91)

where $d_s$ is the dimension of the space and $c_{d_s}$ is the volume of $d_s$-dimensional unit ball \cite{141}:

$$c_{d_s} = \begin{cases} \frac{\pi^{d_s/2}}{d_s^{d_s/2}}, & d_s \text{ even} \\ \frac{1}{d_s^{d_s/2}} \pi^{d_s/2} 2^{d_s-1} \frac{d_s-1}{2}!, & d_s \text{ odd}. \end{cases}$$ \hspace{1cm} (92)

In practical applications $c_{d_s} = \pi$ in $\mathbb{R}^2$ and $c_{d_s} = \frac{4}{3}\pi$ in $\mathbb{R}^3$. Some other possible losses can be packed into the absorption coefficient, such as the scattering and penetration losses as seen earlier. Then the total absorption coefficient, the extinction coefficient, in (91) becomes a sum over all loss coefficients

$$\kappa_e(f) = \kappa_a(f) + \kappa_{\text{others}}(f),$$ \hspace{1cm} (93)

since the overall propagation effects are multiplied in the frequency domain. This is valid for the loss mechanisms represented by the transmittance given by the Beer-Lambert law, such as the one for molecular absorption: $\exp(-\kappa_a(f)r)$ in (91) and (1).

Knowing the path loss and assuming an interference limited network, the SINR $\Upsilon(f)$ at frequency $f$ can be calculated as

$$\Upsilon(f) = \frac{S(f)}{N_n(f) + I_{agg}(f)} \approx \frac{S(f)}{I_{agg}(f)},$$ \hspace{1cm} (94)

where $S(f)$ is the received desired signal power, $N_n(f)$ is the noise power in the system, and $I_{agg}(f)$ is the aggregated interference power at the Rx. The approximation of a small noise level is a consequence of assuming a dense network. The signal power is given by

$$S(f) = P_{Tx}(f)G_{Tx}(f)G_{Rx}(f)l(r) = P_{Tx}(f)G_{Tx}(f)G_{Rx}(f)\frac{\exp(-\kappa_a(f)r)}{c_{d_s}d_s^{d_s-1}},$$ \hspace{1cm} (95)

where the transmitting power density $P_{Tx}(f)$ is mostly normalized to unity below, and $G_{Tx}(f)$ and $G_{Rx}(f)$ are the transmitter and receiver antenna peak gains at the main lobes. The main interest is on the isotropic antennas, where the antenna gains are
\( G_{\text{Tx}} = G_{\text{Rx}} = 1 \). However, directional antennas are considered in Section 4.2.3. The directional antenna case is also illustrated in Fig. 60, where the antenna gains are calculated from the random angles between the transmitting and receiving antennas.

### 4.2 Derivation of the models

Stochastic geometry is based on utilizing the stochastic average values for the interference power, which is a summation over the interfering transmitters in the network. All the calculations below are based on [77, 78, 83, 84, 89].

The aggregate interference of the network is

\[
I_{\text{agg}}(f) = \sum_{i \in \zeta} P_{\text{Tx}} l(r_i), \tag{96}
\]

where \( \zeta \) is a set of interfering nodes in area/volume \( |A| \) of the network. In the path loss model, fading is not considered as a random process, since the molecular absorption causes deterministic fading given the distance \( r \). Also, the THz transmissions are very sensitive to the existence of the LOS path and as a consequence the usual assumption of Rayleigh fading cannot be used, as it will be discussed in the context of the outage probability analysis in Section 4.2.4.

#### 4.2.1 Modelling principle of interference

In order to evaluate the aggregate interference in a network, modelling the interference as a shot noise process has been widely used (e.g., [77]). The idea in the shot noise process is to model the noise components as Poisson distributed time instants. Considering a spatial random process, the time instants can be replaced with spatial locations of the nodes and the impulse responses associated with the time instants can be replaced by the path loss model. Then (96) can be assumed to describe the interference as a shot noise process. This is a very handy assumption, since it allows the usage of well-known theories to analyse the aggregate interference in the network. Furthermore, if Poisson distributed nodes are assumed, the average number of nodes in any area/volume \( |A| \) can be described with a Poisson intensity parameter \( \rho_n \). Assuming the ALOHA channel access for the nodes has a probability \( p \) of transmitting, the average number of transmitting nodes is \( N = pp_n |A| \). Thus, the transmitting nodes form a Poisson distributed network with intensity \( pp_n \).
4.2.2 Mean interference

The mean interference power can be evaluated at least in two different ways: with the probability generating functionals (PGFLs), or directly with the Campbell theorem [89]. The latter is required also in the former case detailed below. Since the PGFLs are more general, they are given first. The PGFLs can be used to calculate the moments of a function, thus, also the mean interference power. The Laplace transform of the aggregate interference power is calculated first and is as given by [77, 78]

\[
\mathcal{L}_{I_{\text{agg}}}(s) = \mathbb{E} \left[ \exp \left( -s \sum_{i \in \zeta} I(r_i) \right) \right],
\]

(97)

where unit transmitting power is assumed. Since there is no random fading as discussed above, the above equation can be written as

\[
\mathcal{L}_{I_{\text{agg}}}(s) = \mathbb{E}_\zeta \left[ \prod_{i \in \zeta} \exp(-sl(r_i)) \right].
\]

(98)

The above expression is called the PGFL [77], which is denoted as

\[
\Omega(\nu) = \mathbb{E}_\zeta \left[ \prod_{x \in \zeta} \nu(x) \right],
\]

(99)

where

\[
\nu(x) = \exp(-sl(r_x)).
\]

(100)

The PGFL for the PPP is further given by [77]

\[
\Omega(\nu) = \exp \left( - \int_{\mathbb{R}^d} (1 - \nu(x)) \Lambda(x) dx \right),
\]

(101)

where \(\Lambda(x)\) is the intensity function of the PPP. In the case of a homogenous Poisson process

\[
\Lambda(x) = c_d d_s \rho_n x^{d_s - 1}
\]

(102)

with a fixed \(\rho_n\) over distances (homogenous network). In the sequel, the propagation space dimension for the signals and the node drop space dimensions are separated. The node drop space is the space in which the users are distributed. The propagation space has quantities \(c_d\) and \(d_s\) for the unit volume and the dimension of the space, respectively. Notation \(c_{d_s}\) is used for the unit volume of the node drop space in \(d_s\)-dimensional space.
This is done because a very likely scenario for the network would be a two-dimensional node drop space, but a three-dimensional path loss. Due to usage of the unit area term $c_d d_d$, the integral over $\mathbb{R}^d$ is just an integration over the distance $r$. The Laplace transform of the aggregate interference becomes

$$\mathcal{L}_{agg}(s) = \exp \left[ -c_d d_d pp_n \int_{\mathbb{R}^d} (1 - \exp(-sl(r))) r^{d-1} dr \right].$$

(103)

The above equation has closed form solutions if the path loss exponent is larger than the dimension of the space [77]. However, because the path loss exponent here follows the free space path loss, our path loss exponent is always smaller than the dimension of the space. Therefore, some solutions have to be numerically estimated. This is not a major problem, since it is still much faster to do the numerical computations than complete Monte Carlo system simulations. However, there is always a chance of reduced accuracy.

The $n$th moment of the interference power can be calculated as [80]

$$(-1)^n \frac{d^n}{ds^n} \mathcal{L}_{agg}(s) \bigg|_{s=0}.$$  (104)

The mean interference power becomes $-d/ds(\mathcal{L}_{agg}(s)|_{s=0})$. In the case of two-dimensional node drop space and three-dimensional path loss, the derivative has no closed form solution, but it reduces to the Campbell theorem as seen below. In the case $d_d = d_s$, (104) does have a closed form solution:

$$-\frac{d}{ds} \mathcal{L}_{agg}(s) \bigg|_{s=0} = \mathcal{T}(f) = \frac{p \rho_n}{\kappa_a(f)},$$

(105)

where $\mathcal{T}(f)$ is the mean interference power at frequency $f$ assuming unit transmitting power. This special case can also be shown directly through the Campbell theorem. In the case of unit transmitting power is not assumed, $s$ in all the above equations should be replaced with $s = P_{Tx}(f)s$ [77]. Summing all the effects, the estimates for the first two moments can be easily estimated by calculating the derivatives of (104). Marking the Laplace transform in (103) as

$$\mathcal{L}_{agg}(s) = \exp(-L(s)),$$

(106)
the inner function derivatives can be calculated rather straightforwardly:

\[ L'(s = 0) = P_{Tx}(f) p \rho_n e_{d_s} \int_0^\infty r^{d_s-1} \exp(-\kappa_s(f)r) dr, \quad (107) \]

and

\[ L''(s = 0) = P_{Tx}(f)^2 p^2 \rho_n \left( \frac{e_{d_s} e_{d_d}}{e_{d_s} e_{d_d}} \right)^2 \int_0^\infty r^{d_s-2d_s+1} \exp(-2\kappa_s(f)r) dr. \quad (108) \]

The mean interference power and its variance, respectively, can be calculated by plugging the above expressions into the derivatives of (106)

\[ \bar{I}(f) = L', \quad (109) \]

and

\[ \text{var}(\bar{I}(f)) = L'' + (L')^2. \quad (110) \]

Similar expressions can be derived for any higher moments. Also, directional antenna effects can be taken into account by multiplying the transmitting power with the expected antenna gains, which is detailed in Section 4.2.3.

If one is interested in the mean interference level only, the Campbell theorem offers more straightforward way to estimate it. It states that [89, p. 28]

\[ \mathbb{E} \left( \sum_{x \in \zeta} f(x) \right) = \int \frac{f(x)}{\Lambda(x)} dx. \quad (111) \]

Therefore, the mean interference power becomes

\[ \bar{I}(f) = c_{d_s} d_s p \rho_n \int \frac{\exp(-\kappa_s(f)r)}{c_{d_s} d_s r^{d_s-1}} dr. \quad (112) \]

Calculating the above expression, it can be seen that when \( d_s = d_d \), the equation can be solved as

\[ \bar{I}(f) = \frac{p \rho_n}{\kappa_s(f)}, \quad (113) \]

or

\[ \bar{I}(f) = P_{Tx}(f) \frac{p \rho_n}{\kappa_s(f)} \quad (114) \]

for non-unit transmitting powers. If \( d_s > d_d \), the above equation does not have a closed form solution as it was also seen above.
4.2.3 Directional antennas

Directional antennas are often envisioned being utilized in the higher frequencies because of the large path loss, but also because of the potential benefit of large multiple-input multiple-output (MIMO) transmission gains [21]. MIMO configurations would also be a natural way to realize directional antennas.

The above stochastic geometry analysis with isotropic antennas means that the path loss has unity value for the antenna gain factors $G_{Rx}(f, \Psi, \Phi)$ and $G_{Tx}(f, \Psi, \Phi)$ for the Rx and Tx, respectively, in all the directions $(\Psi, \Phi)$, where $\Psi$ is the azimuth angle and $\Phi$ is the elevation angle. There have been some works on directional antennas for the stochastic geometry, e.g., [87, 88].

When the directional antennas are taken into account, two cases can be considered: 1) the total transmitting power is the same as in the isotropic case, and 2) it is different to the total isotropic power. Both cases are rather straightforward extensions. The expected effect of the directional antennas can be modelled by the expected antenna gain. Assuming the same pattern for the Tx and Rx ends, and a unit integral over the total transmitting power over the antenna radiation pattern, the antenna gain averaged over all angles becomes

$$E_\nu[G_{Rx}(f, \nu)] = E_\nu[G_{Tx}(f, \nu)] = \frac{G_i}{c_d d_s} \int G_a(f, \nu) d\nu, \quad (115)$$

where $G_a(f, \nu)$ is the antenna pattern, $G_i$ is the gain compared to the total isotropic power, and $\nu$ is the azimuth/elevation angle dependent on the dimensions of the system. The factor $G_i$ is included just to isolate the two cases above, i.e., $G_i = 1$ for case one, and something else for the second case. If isotropic antennas are assumed, the above equation yields an expected gain of $1/c_d d_s$ in all directions, since the total transmitting power is obtained from

$$P_{Tx} \int_{\mathbb{R}^{d_t}} \frac{1}{c_d d_s} d^d R = P_{Tx}, \quad (116)$$

as $P_{Tx}$ sets the limit for the total transmitting power (assuming there is no additional gain).

Case one above yields the same values for the expected antenna gains as the isotropic case, because the total transmitting power and the average radiated/receiver power are the same due to the random angles to and from the interfering transmitters. Then results in Section 4.3 demonstrate that the expected interference is indeed the same if the
total transmitting power is the same. Regardless of the same interference, the antenna gain provides \( \max_u(G_{Tx}(f, v)) \times \max_u(G_{Rx}(f, v)) \) times better signal-to-interference ratio (SIR), because the desired link power is higher (recall Fig. 60 and the assumption regarding the desired link antenna directions).

If the additional expected antenna gain is less than, or larger than the effective isotropic gain, it naturally has an impact on the received average interference power. However, the SIR remains the same if the antenna pattern remains the same because the desired link power decreases or increases with the gain. In any case, the theoretical effect of the expected antenna gains can be included in the mean interference similarly to the case for free space path loss. The mean gain becomes

\[
\bar{I}(f) = p c_d d_p d_r E[G_{Rx}(f)] E[G_{Tx}(f)] \\
\times \int \frac{\exp(-\kappa(f)r)}{c_d d_p d_r r^{d_r-1}} d_r.
\]

(117)

The validity of this will be shown in the section on the numerical results, where the symmetrical antenna response is assumed in the elevation and azimuth directions. The utilized antenna patterns in two- and three–dimensional spaces are shown in Fig. 61. The antenna gain is always assumed to point towards the positive horizontal axis. Thus, the coordinate axes are rotated according to the system geometry. The figure does not take into account the actual gains in the direction of the antenna patterns. These are not the most realistic antenna patterns, but they allow easy validation due to their straightforward mathematical representation. If the antenna half beam width is \( \Psi/2 \), then the gains toward these angles become

\[
G_{2D}(f, -\Psi/2 : \Psi/2) = \frac{1}{\Psi},
\]

(118)

for the 2D path loss, and

\[
G_{3D}(f, -\Psi/2 : \Psi/2) = \frac{1}{2\pi(1 - \cos(\Psi/2))},
\]

(119)

for the 3D path loss. Please notice that the 3D antenna gain is defined only for the azimuth angle \( \Psi \). The property will be explained in more detail the next section, but it is related to the fact that simple (symmetric) antenna patterns can be represented in 2D space and by scaling as above and by scaling with the probability of antenna pointing in certain direction (as done below). The latter term comes from the expression for the area of a spherical cap, which also defines the steradian of the angle around the cone set by
the antenna half beam. Both gains reduce to isotropic gains if the antenna half beam is set to $\Psi/2 = \pi$.

The expected antenna gains for the cases of interest become

$$G_a(f) = G_{2D}(f, -\Psi/2 : \Psi/2) \left(\frac{\Psi}{2\pi}\right)^2,$$

(120)

for a 2D node drop and 2D path loss assuming identical gain pattern in the both Rx and Tx. The term $\Psi/2\pi$ is the probability that the Tx/Rx points at the Rx/Tx in a 2D path loss model. Similarly, the expected antenna gains for a 2D node drop and 3D path loss, and a 3D node drop and 3D path loss become, respectively,

$$G_a(f) = G_{3D}(f, -\Psi/2 : \Psi/2) \left(\frac{\Psi}{2\pi}\right)^2,$$

(121)

$$G_a(f) = G_{3D}(f, -\Psi/2 : \Psi/2) \left(\frac{1}{2}(1 - \cos(\Psi/2))\right)^2,$$

(122)

where $0.5(1 - \cos(\Psi/2))$ is the probability that the Tx/Rx points at the Rx/Tx in the 3D path loss model. The 2D node drop with 3D path loss is similar to the 2D/2D case because the interfering links and the desired link are assumed to be parallel to the node drop plane. If the desired link, or the interfering links, would point in random direction(s), the expected interference would be $G_{3D}(f, -\Psi/2 : \Psi/2) \left(\frac{\Psi}{2\pi}\right)^2 (1 - \cos(\Psi/2))$. However, it is unlikely the links would point in random elevation directions if all the nodes are on a plane. The fourth case of a 3D node drop and 2D path loss model is not very likely in any realistic systems, but the expected antenna gain can be calculated similarly as in the 2D/2D case, but with three–dimensional antenna gain (or the same as 2D/3D).
4.2.4 **Outage probability**

The success probability \( p_s(f) \) of the desired transmission can be calculated with (103) by evaluating the function at point \( s = \beta_l r \), where \( l(r) \) is the path loss of the desired signal and \( \beta_l \) is the threshold for the SIR [77], i.e.,

\[
p_s(f, \beta_l) = \mathbb{P}(S(f) > \beta_l l(r)) = \mathbb{E}\exp(-I_{agg}(f)\beta_l r^{-1}). \tag{123}
\]

However, this is only true if the received power is exponentially distributed, i.e., Rayleigh faded. In the other words, if the signal power is exponentially distributed, the SIR distribution will be precisely the Laplace transform. This is not the case in the problem here. Therefore, the moments are calculated from the derivatives of the Laplace function in (105). The difficulty is to find a suitable distribution for the interference. This is not an easy task, although the moments are known. Based on the simulations, one distribution has a relatively good fit with the simulated values: the log-logistic distribution. It has parameters that can be mapped from the theoretical moments with a fairly good fit. However, it fails in the cases of very dense or thin networks. Furthermore, the log-logistic distribution gives a good fit in the case of two-dimensional path loss, but
fails to estimate the interference distribution for three–dimensional path loss. This is demonstrated in the numerical results in Section 4.3.

The log-logistic distribution is characterized by the probability density function

\[ f(x) = \frac{(\beta / \alpha)(x/\alpha)^{\beta-1}}{(1 + (x/\alpha)^{\beta})^2}, \]  

where \( \alpha \) is the scale parameter, and \( \beta \) is the shape parameter. These can be solved numerically from the moments of the interference distribution and the moments of the log-logistic distribution

\[ E[X^k] = \alpha^k \frac{k\pi/\beta}{\sin(k\pi/\beta)}. \]  

(125)

Given the interference distribution, the outage probability is

\[ p_o(S(f), \beta) = 1 - p_s(f, \beta) = 1 - S(f)/\beta \int_0^{S(f)/\beta} f_I(x)dx, \]  

(126)

where \( f_I(x) \) is the probability density function of the interference power of the desired node.

4.3 Numerical results of the stochastic geometry

The utilized parameters are shown in Table 10. Frequency-dependent results are not given, but a fixed absorption coefficient is utilized, along with a path loss model without the antenna aperture term. Therefore, the results represent the interference intensity at the location of the desired receiver. The frequency dependencies are easy to add on in the form of the antenna aperture, as well the real and frequency dependent absorption coefficient. By leaving the frequency dependency out, the network specific behaviour of the interference can be shown better.

Before going into the numerical examples, the simulation model is introduced in the next section. This simulation model was used to validate all the presented stochastic models.

4.3.1 Simulation model for the THz networks

As mentioned in the previous section, the simulation model is based on two–dimensional antenna patterns. This can be done due to the utilization of a dot product to calculate the
Table 10. Parameters used for the stochastic models.

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Absorption coefficient</td>
<td>$\kappa_a$</td>
<td>0.1 cm$^{-1}$</td>
</tr>
<tr>
<td>Antenna beam width</td>
<td>$\Psi$</td>
<td>$\pi/4$</td>
</tr>
<tr>
<td>Probability to transmit</td>
<td>$p$</td>
<td>1</td>
</tr>
<tr>
<td>Transmitting power</td>
<td>$P_{Tx}(f)$</td>
<td>1 W/Hz</td>
</tr>
<tr>
<td>Transmitter–receiver distance</td>
<td>$r$</td>
<td>1 cm</td>
</tr>
</tbody>
</table>

The angle between the positive x-axis and the random point in space and because of the assumed perfectly symmetrical antenna patterns.

Regardless of the dimension of the space, the interfering nodes are distributed evenly around the origin (i.e., the desired receiver) from 0 to 100 cm distances. Thus, the 100–cm radius defines the node drop area or volume as a circle or a sphere. The 100–cm radius is large enough in the THz band to ensure the interference contribution to the origin is small beyond this distance (with reasonable transmitting powers). The nodes are distributed in spherical coordinates based on the disc/sphere point picking principle to achieve an even distribution for the nodes. In the case of isotropic antennas, it is enough to calculate the Euclidean distances from random points to the origin due to equal antenna gain in all directions. The distances and the antenna gains can then be input into the channel model

$$l(r) = \frac{\exp(-\kappa_a(f)r)}{c_d d_s r^d s_{d_s} r_d s_{d_r} - 1} G_{Rx}(f) G_{Tx}(f)$$

and the total interference is given by (96) with the average number of $N$ interfering nodes drawn from the Poisson distribution taking into account the transmitting probability $p$, effectively thinning the Poisson process.

In the case of directional antennas, the antenna directions (azimuth/elevation) are randomly chosen based on the disc/sphere point picking principle again to ensure even distribution for the direction. Next, the angles from the origin to the random points are calculated, as well as the angles from the random points with random coordinate axes to the origin. The position of the origin from the new random coordinate axes of the interfering nodes need to be calculated. This is a straightforward negation of the coordinates of the random nodes, i.e., if every interfering node is considered to lie in its own origin, then the original origin is at $(x', y', z') = (-x, -y, -z)$ from the nodes. This, however, means all the coordinate axes are oriented similarly. To take into account the rotations by the random transmitting angles, the rotation operation(s) is applied to the
coordinates to calculate the real location of the original origin in the new coordinate axes. The rotations can be done in many ways, but rotation matrices are used here to calculate the coordinates

\[
\begin{bmatrix}
  x' \\
  y'
\end{bmatrix} =
\begin{bmatrix}
  \cos(2\pi - \theta) & -\sin(2\pi - \theta) \\
  \sin(2\pi - \theta) & \cos(2\pi - \theta)
\end{bmatrix}
\begin{bmatrix}
  -x \\
  -y
\end{bmatrix}
\]

(128)

for the 2D node drop and

\[
\begin{bmatrix}
  x' \\
  y' \\
  z'
\end{bmatrix} =
\begin{bmatrix}
  \cos(2\pi - \phi) & 0 & \sin(2\pi - \phi) \\
  0 & 1 & 0 \\
  -\sin(2\pi - \phi) & 0 & \cos(2\pi - \phi)
\end{bmatrix}
\times
\begin{bmatrix}
  \cos(2\pi - \theta) & -\sin(2\pi - \theta) & 0 \\
  \sin(2\pi - \theta) & \cos(2\pi - \theta) & 0 \\
  0 & 0 & 1
\end{bmatrix}
\begin{bmatrix}
  -x \\
  -y \\
  -z
\end{bmatrix}
\]

(129)

for the 3D node drop, where the right-hand matrix rotates the coordinate axes about the z-axis, and the left-hand matrix rotates the coordinate axes about the y-axis. The subtraction of the random angles from 2\pi shifts the rotation matrix from rotating the point in the coordinate axis in the counter-clockwise direction to rotating the coordinate axis in the counter-clockwise direction. This is important with respect to the point mapping, since the trigonometric functions rotate in a counter-clockwise direction. Therefore, it is more convenient that the coordinate axes rotate in this direction in this intermediate phase. Based on the new coordinates, the angles from the origin to the nodes can be calculated with a dot product

\[
\vartheta_{Rx} = \cos^{-1}\left(\frac{x}{\sqrt{x^2 + y^2 + z^2}}\right),
\]

(130)

where \(\cos^{-1}(\cdot)\) is the inverse cosine function. Similarly, the angles from the nodes to the origin similarly, i.e.,

\[
\vartheta_{Tx} = \cos^{-1}\left(\frac{x'}{\sqrt{x'^2 + y'^2 + z'^2}}\right),
\]

(131)

assuming the transmitters and the receiver at the origin are pointed at the positive x-axis (a unit vector along the positive x-axis). Then the angles for the antenna patterns and the Euclidian distances can be input into the path loss model. Two-dimensional antenna patterns can be used even in three-dimensional space. The reason for this is the increased probability for the higher angles due to additional degree of freedom provided
by the elevation angle. Of course, this procedure depends on the antenna patterns, and this assumption can only be used with symmetric antenna patterns as in the case here.

### 4.3.2 Mean interference power

The mean interference power and the corresponding variance are calculated with (109) and (110) and compared against the raw moments of the simulation data given by $E[I_{\text{sim}}]$ with the results shown in Figs. 62 and 63, respectively. The simulation data was obtained with the simulation model presented in the previous section. Unlike in the case of directional antennas, the antenna gain was assumed to be one in all the directions. Figs. 62 and 63 show a very good agreement between the theory and the simulation for both the mean and the variance. The full 3D case suffers from a larger variation in variance caused by the relatively thinner network and in general a lower interference power because of the larger path loss. Still, the results give a perfect match, also in the case of SIR in Fig. 64. The SIR was calculated with the path loss model given in (91) and with the parameters given in Table 10. The short desired link distance leads to large SIR values, especially in the three-dimensional case. However, the lower dimensions show considerably lower SIRs.
### 4.3.3 Outage probability

The outage probabilities were calculated with (126) for the non-fading case and with (123) for the fading case. As it was mentioned earlier, the log-logistic distribution is

![Graph showing outage probability](image)

**Fig. 63.** The variance of the mean interference power versus the number of nodes in the network.

![Graph showing SIR](image)

**Fig. 64.** SIR versus the number of nodes in the network.

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not a perfect solution for describing the general interference distribution. A perfect distribution would be a three–parameter generalized extreme value (GEV) distribution [143], but it requires simulation data for parameter fitting.

The results in Figs. 65–68 comprise the simulation results of the probability of an outage for the non-fading and the fading cases (unit mean Rayleigh fading), as well as the theoretical results given by the log-logistic distribution, Eq. (123), and the GEV (with simulation aided parameters). Interestingly, as predicted by the previous works, the Laplace transform of the aggregate interference power gives the outage distribution in the presence of fading and this is true also in the case of the THz band.

As it was discussed above, the GEV gives a perfect distribution for the interference distribution without fading. A fairly good option is given by the log-logistic distribution, which can be mapped directly from the raw moments from the theory. It provides a good fit in the case of 2D path loss (see Figs. 65 and 66), with limitations in the accuracy in the case of very thin networks. When using the 3D path loss model, it mostly fails to predict the interference distribution, particularly if the node density is too low or too high. Therefore, the interference distributions in the THz band should be subject of further investigation in the future. This is a very important issue, since the proper distributions combined with the theories would give all the required information on the network behaviour.

Fig. 65. Outage probabilities and the simulation data for 2D node drop and 2D path loss versus the desired link’s Rx–Tx distance [© 2017 IEEE].
Fig. 66. Outage probabilities and the simulation data for 3D node drop and 2D path loss versus the desired link's Rx–Tx distance © 2017 IEEE.

Fig. 67. Outage probabilities and the simulation data for 2D node drop and 3D path loss versus the desired link’s Rx–Tx distance © 2017 IEEE.
4.3.4 Directional antennas

The mean interference powers and SIRs for the directional antenna scenarios are given in Figs. 69–74 based on the models in Section 4.2.3. There are two observations to be made from the results. Firstly, if the effective isotropic radiated power is the same, the directional antennas do not decrease or increase the interference power. This is because the expected antenna gains are equivalent in both the isotropic and the directional antennas. An exception to this is caused by the 2D node drop and 3D path loss. In this case, and with the assumptions here, the antenna patterns are parallel to the plane. Therefore, the probability of receiving interference from the plane is the same as with the two dimensional path loss, however, with three dimensional path gain on the antenna. This is clearly visible in Fig. 70, which shows approximately 10 dB larger interference power for the directional antenna case in comparison to the isotropic case. This is a direct consequence of the probabilities of receiving interference, which are (similarly as in (121) and (122)) $\Psi/2\pi$ for the 2D case and $0.5(1 - \cos(\Psi/2))$ for the 3D case. In this case the expected increase of the interference when moving to the directional antennas is $(\Psi/2\pi)^2/[0.5(1 - \cos(\Psi/2))]^2 = 10.79 \equiv 10.33dB$ with $\Psi = \pi/4$.

In the general case, if the total power is increased or decreased with respect to the isotropic power, it naturally has an impact on the total interference power, as it can be
seen in Figs. 69–71 (four-fold increase and decrease in power for high and low power cases, respectively). This leads to the second interesting finding: effective isotropic power has no impact on SIR, because both the desired and the interfering links are scaled simultaneously. This follows naturally from the assumption of an interference limited network. Although the directional antennas have no impact on the total interference power in the network (given equal total transmitting power in the isotropic case), they do give a large gain in the SIR performance because of the increased desired link power, which is assumed to be perfectly aligned in these examples. As discussed earlier, this is a very likely scenario in the THz band because of the large path loss. Directional links will make it significantly easier to mitigate the large losses in the channel, as it can also be seen in the results. Of course, they reflect the effects of very simple system geometry with a simple antenna pattern. Regardless of this, the results here show that stochastic geometry can be very efficiently used in the THz band network interference modelling.

Fig. 69. Mean interference power for 2D node drop and 2D path loss versus the number of nodes in the network for various antenna configurations.
Fig. 70. Mean interference power for 2D node drop and 3D path loss versus the number of nodes in the network for various antenna configurations.

Fig. 71. Mean interference power for 3D node drop and 3D path loss versus the number of nodes in the network for various antenna configurations.
Fig. 72. SIR for 2D node drop and 2D path loss versus the number of nodes in the network for various antenna configurations.

Fig. 73. SIR for 2D node drop and 3D path loss versus the number of nodes in the network for various antenna configurations.
4.4 Discussion

It has been shown that stochastic geometry can be utilized in the network analysis for dense THz networks, such as nanonetworks. The moments of interference can be reliably estimated based on the stochastic models presented here. Similarly, the outage probabilities can also be estimated. The full evaluation of the outage will require considerable future work on the interference distributions. For instance, the GEV distribution gives a perfect fit for the simulations, but requires simulation data for parameter fitting. This is not a desired feature when the core idea of stochastic geometry is to avoid heavy simulations. Instead, the interference distribution should be parameterized from the theoretical moments, or directly from the Laplace transform.

Directional antennas can also be included in the models. This is very important in the case of higher frequencies, as the high loss links will almost surely require highly directional antennas to offer reliable communications. Very simple and illustrative antenna patterns were assumed here, leaving room for future work on more realistic antenna patterns. However, the antenna pattern has no impact on the overall validation of the mean interference models since the results can be scaled with the mean antenna gain.
Thus, more complex antenna patterns can be utilized in the place of the mathematically and geometrically simple ones. The analysis herein shows that stochastic geometry can very efficiently be used even at very high frequencies. This makes the evaluation of future THz band networks easy with the help of the tool provided by the stochastic geometry.
5 Conclusions and future work

The THz band has become a very popular research topic during the past few years. Suitable communication solutions for the THz band have been proposed in a number of papers. Yet, the channel behaviour is not completely understood. The LOS path is very well known and can be estimated based on spectroscopic catalogues. However, the NLOS paths require investigations and more work in the future. This thesis focused on the channel characterization in various NLOS and LOS phenomena, such as small particle scattering, penetration loss, reflections and scattering on surfaces, and diffraction properties.

Molecular absorption is familiar from earlier works and was briefly discussed. Molecular absorption causes frequency selective, yet deterministic loss to the signals. This has been assumed to cause a new type of noise known as transmission induced noise, which is caused by the re-radiation of absorbed energy. This was analysed from multiple physical viewpoints. The conclusion was that this noise is most likely very weak. However, the proportions of different phenomena (e.g., the amount of energy that transforms into heat, or directly radiates in the medium, etc.) are unknown and require further investigations in the future. These proportions define the strength of the transmission induced noise. Whereas as a single link will most likely not suffer from the noise, a dense nanonetwork may have enough energy to cause an increase in the noise level. This area is subject to future work.

It was also shown that under particular conditions, i.e., if the scattering particles are large in number and size, small particle scattering causes frequency dependent loss to the signals, viz., increasing loss towards the larger frequencies. This may also lead to time delayed signal components as a consequence of multiple sequential scattering events. These time delayed signal components may also have an effect on the energy detection performance due to the limited detection window length. On the other hand, delayed signal components may increase the received power because the power can be gathered also outside the LOS path. The risk of ISI still makes multiple scattering an undesired phenomenon. The future work in this case is to extend the small particle loss model with the Mie scattering theory in order to take into account arbitrary sized particles in the atmosphere.
The reported measurement results included penetration loss through different materials, reflection and scattering from surfaces, and diffraction effects. The common feature of all these is that the higher frequencies are more attenuated. This suggests a higher probability for establishing links at the lower THz frequency bands if the LOS path is obstructed. This decreases the utilization potentiality of the THz band in an arbitrary environment. However, the measurements did show that the NLOS path could be utilized to establish connections, thus, making the THz band more versatile than some have expected. In fact, the THz band was shown to have quite a rich multipath environment.

The future work on the measurements should include more measurements on different materials. This is important for channel estimation, but also from the viewpoint of simulation models, such as ray tracing models. These require detailed information on the propagation environment, but especially the reflection and scattering properties of various materials of various roughnesses. One important subject for the measurements are the reflection coefficient measurements. These require a specific measurement setup, but would provide a lot of valuable information on the general behaviour of different materials. The refractive indices could be used in estimating reflection and penetration efficiencies, as well as the delay properties of the NLOS paths.

As a final part of this thesis, stochastic geometry was utilized to derive models for estimating the mean interference level and outage probability in THz band networks. The main difference to the previous models was the channel model which included the FSPL, but importantly also molecular absorption loss. The mean interference model was proven to work very well and match the simulation results perfectly. The outage probability can be estimated with reasonable accuracy, but it requires further investigations in the future. The main issue is in the interference distributions. In the presence of Rayleigh fading, the existing stochastic geometry models work very well (as it was shown). However, without it, the interference distributions become difficult to estimate and suitable distributions require further work in the future. Still, the stochastic geometry is very useful with respect to THz band network level research.

The overall picture of the THz band shows great potential for future short range communication systems. It can support NLOS paths, which is an excellent feature as the LOS condition cannot always be guaranteed. Based on the work here, and the works done by research groups around the globe, the THz band will offer great opportunities for many applications. The main obstacle today is the hardware, a minor bump on the road towards the super high data rates offered by the higher frequencies.
References


Appendix 1 Derivation of the Emission Rate for Molecular Noise

The famous Einstein coefficients and their relationships are [144]

\[ A_{21} = \frac{8\pi hf^3}{c^3} B_{21}, \]  
(132)

\[ g_1 B_{12} = g_2 B_{21}, \]  
(133)

where \( A_{21} \) is the Einstein coefficient for spontaneous emission, \( B_{21} \) is the Einstein coefficient for stimulated emission, \( B_{12} \) is the Einstein coefficient for absorption and constants \( g_1 \) and \( g_2 \) are statistical weights (degeneracies) of the states 1 and 2, respectively. These coefficients correspond to the two-level approximation of the system and can be obtained from the HITRAN catalogue [13], except for the \( B \) coefficients, but which depend directly on the \( A \) coefficient and the degeneracies of the states as shown above. The unit of the Einstein \( A \)-coefficient is \( \text{s}^{-1} \), i.e., the decay as a function of time, or the probability that the upper state will decay to the lower state. Thus, the inverse of Einstein \( A \)-coefficient is telling the lifetime of the state. The unit of the Einstein \( B \)-coefficients is \( \text{s}^{-1}/(\text{J cm}^{-3}\text{Hz}^{-1}) \). As a consequence, the \( B \)-coefficients must be multiplied with energy density per unit frequency interval at frequency of the line, or in \( \text{J cm}^{-3}\text{Hz}^{-1} \).

The energy density at frequency \( f \) in LTE can be obtained from the Planck’s law [113, Chapter 3.6]

\[ B_f = \frac{A_{21}}{B_{21}} \left( \frac{\bar{n}_1 g_2}{\bar{n}_2 g_1} - 1 \right)^{-1} = \frac{8\pi hf^3}{c^3} \left( \exp \left( \frac{hf}{k_B T} \right) - 1 \right)^{-1}, \]  
(134)

where \( \bar{n}_1 \) and \( \bar{n}_2 \) are populations of the upper and lower energy levels under LTE condition, respectively. It should be noticed that \( B_f, \bar{n}_1, \bar{n}_2, g_1, g_2, A_{21} \) and \( B_{21} \) are all functions of frequency. Furthermore, this Planck’s law is different to (10) as it gave the energy flux and the above Planck’s law is for energy density, which is obtained by relations [113, Chapter 3.6]

\[ \frac{A_{21}}{B_{21}} = \frac{8\pi hf^3}{c^3} \]  
(135)

and

\[ \frac{\bar{n}_1}{\bar{n}_2} = \frac{g_1}{g_2} \exp \left( \frac{hf}{k_B T} \right). \]  
(136)
We denote the populations in the non-LTE with $n_1$ and $n_2$, which should not be confused with the LTE populations ($\bar{n}_1$ and $\bar{n}_2$). Under the LTE assumption, a detailed balance between the decay rates is required [113, Chapter 3.6]

$$n_1 B_{12} B_f = n_2 A_{21} + n_2 B_{21} B_f,$$

(137)

The population of states can be obtained as [144]

$$\bar{n}_1 = \frac{g_1 N}{Q_{tot}(T)} \exp \left( - \frac{h c E_1}{k_B T} \right),$$

(138)

$$\bar{n}_2 = \frac{g_2 N}{Q_{tot}(T)} \exp \left( - \frac{h c E_2}{k_B T} \right),$$

(139)

where $N$ is the total number of molecules ($N = pN_A/(RT)$, where $p$ is pressure, $N_A$ is Avogadro’s number and $R$ is gas constant) and $Q_{tot}(T) = \sum \eta g_\eta \exp(-h c E_\eta/(k_B T))$ (140) is the total partition function, where $g_\eta$ is the degeneracy of state $\eta (\{1,2\})$ and $E_\eta$ is the corresponding energy of the state (given by the HITRAN in units 1/cm, thus the $c$ as a multiplier). These parameters are unique for each molecule and the absorption line of the molecule. It is easy to show that $N(f) = \Sigma f n_{1,f} + \Sigma f n_{2,f}$. Therefore, (138) and (139) hold.

The full detailed balance can be given as [113, Chapter 3.6]

$$n_1 B_{12} B_f - n_2 A_{21} - n_2 B_{21} B_f + n_1 C_{12} - n_2 C_{21} = 0,$$

(141)

where $C_{12}$ and $C_{21}$ are the collision rates of the states 1 and 2 respectively. The collisional effects cancel in the LTE (as do the radiative effects). In the further calculations, the environment is assumed to be in LTE, while the transmission breaks the LTE locally, i.e., the collisional effects are neglected and focus is on the radiative transfer of energy. Furthermore, the absorbed power is assumed to be small enough to prevent the local warming of the environment. This assumption is directly related to the collisional effects being assumed to remain in the LTE. This is also shown to be rather good approximation based on calculations here on the temperature increase due to the absorbed energy density in Section 2.2.6.

We begin the calculation of the molecular absorption noise from the rate between the populations in the non-LTE. It can be derived from the non-LTE Planck’s function [113, Chapter 8.2.1]

$$J_f = B_f \frac{n_2 \bar{k}_f}{n_2 \bar{k}_f},$$

(142)
where $k_f$ is absorption coefficient in the LTE and $k_f$ is absorption coefficient in the non-LTE. Ratio of the non-LTE and LTE absorption coefficients is defined as

$$\frac{k_f}{k_f} = \frac{n_1}{\bar{n}_1} \frac{1 - \frac{\bar{n}_1}{\bar{n}_2}}{1 - \exp\left(-\frac{hf}{k_BT}\right)}.$$  

From these two equations the population ratios can be derived in the non-LTE at point $r$ at the zeroth time instant with respect to the re-emission of the energy as

$$\frac{n_1(t=0)}{n_2(t=0)} \equiv C(r) = \frac{g_1}{g_2} + \frac{B_f}{J(r)} \frac{\bar{n}_1}{\bar{n}_2} - \frac{B_f}{J(r)} \frac{\bar{n}_1}{\bar{n}_2} \exp\left(-\frac{hf}{k_BT}\right),$$

where $C(r)$ is the ratio between $n_1(t)$ and $n_2(t)$, $J(r)$ is now the total energy in the system. In the non-LTE, the populations can be solved, given the total energy density $J(r)$ is known and the collision rates are known. Regarding the collision rates, the assumptions above are in effect. The total energy density at distance $r$ is estimated by a summation of the Planck’s law and the absorbed energy as

$$J(r) = B_f + P_{T_{x}} \frac{\kappa_{a} e^{-\kappa_{a} r}}{4\pi r^2}. $$

It can be seen that if $\kappa_{a} = 0$, $J(r) = B_f$. Then (142) reduces to the Planck’s function in the LTE. From (144), $n_1(0)$ and $n_2(0)$ can be solved from the relationships: $n_1(0)/n_2(0) = C(r)$, $n_1(0) = NX$, and $n_2(0) = N - n_1(0)$, where $X$ is an arbitrary constant for which $n_1(0) = NX$, where $N$ is the total number of molecules. This yields $n_1(0) = NC(r)/(C(r) + 1)$ and $n_2(0) = N/(C(r) + 1)$. By knowing the initial populations and the LTE values for them (Eqs. (138) and (139)), the populations of the upper and lower states can be derived.

The derivation of the time dependent populations here begins with assumption that the upward and downward rates are equal in LTE. Furthermore, it is assumed that in non-LTE the upper state population has increased due to absorption and the population rate change is

$$\frac{dn_2}{dt} = -(n_2(t) - \bar{n}_2)R_{21},$$

where $n_2(t) - \bar{n}_2$ describes the deviation of the upper state population from the LTE value and $R_{21}$ is the rate at which the overloaded upper state decays towards the LTE value. The lower state is assumed to have small effect on the upper state since it is underpopulated compared to the LTE. The minus sign is due to the derivative is negative as a consequence of decaying upper population. In LTE, $dn_2/dt = 0$. The above equation meets this demand as in LTE $n_2(t) = \bar{n}_2$ and due to $dn_2/dt = 0$. 

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We can derive the time domain population of the upper state by rearranging (146) and multiplying with \( \exp(R_{21}t) \):

\[
\frac{dn_2}{dt} e^{R_{21}t} + n_2(t)R_{21}\exp(R_{21}t) = \bar{n}_2 R_{21} e^{R_{21}t}.
\]  

(147)

The term \( \frac{dn_2}{dt} \exp(R_{21}t) + n_2(t)R_{21} \exp(R_{21}t) = d/dt(n_2(t) \exp(R_{21}t)) \) by partial derivative rule. Integrating both sides

\[
\int \frac{d}{dt}(n_2(t)e^{R_{21}t})dt = \int \bar{n}_2 R_{21} e^{R_{21}t} dt
\]  

(148)

yields

\[
n_2(t)e^{R_{21}t} = \bar{n}_2 e^{R_{21}t} + D,
\]  

(149)

where \( D \) is integration constant. From the initial value for \( n_2(t = 0) \), \( D = n_2(0) - \bar{n}_2 \). Thus,

\[
n_2(t) = \bar{n}_2 + (n_2(0) - \bar{n}_2)e^{-R_{21}t}.
\]  

(150)

We can check the correctness of this equation by inspecting the value of \( n_2(t) \) at the boundaries. At \( t = 0 \), (150) gives \( n_2(0) \) and at \( t = \infty \), (150) gives \( \bar{n}_2 \) as it could be expected as the overloaded upper state should decay towards the LTE value after some time from the excitation event. Knowing the \( n_2(t) \), \( n_1(t) \) becomes \( N - n_2(t) \). The number of possible transitions remains at constant value \( N = \bar{n}_1 + \bar{n}_2 \equiv n_1(t) + n_2(t) \).

Therefore, the lower state population can be written as

\[
n_1(t) = \bar{n}_1 - (n_2(0) - \bar{n}_2)e^{-R_{21}t}.
\]  

(151)

It should be noticed that this derivation is an approximation of the reality. We do not take into account the fact, that even in the LTE, there are upward and downward fluxes of energy due to the Einstein coefficients. However, the fluxes are assumed to cancel each other in LTE. Furthermore, the upward flux is assumed to be negligible in non-LTE.

As a consequence, it is assumed that only the overloaded part of the population decays and the equations above holds. The full non-LTE problem is not easy to solve. This can be seen by starting from the full problem:

\[
\frac{dn_2}{dt} = n_1(t)R_{12} - n_2(t)R_{21},
\]  

(152)

where \( R_{12} = B_{12}B_f, n_1(t)R_{12} \) is a source term for \( n_2(t) \) and \( n_2(t)R_{21} \) is a sink term. Similarly, for \( n_1(t) \):

\[
\frac{dn_1}{dt} = n_2(t)R_{21} - n_1(t)R_{12}.
\]  

(153)
With similar approach to above, it is easy to derive the populations. However, assuming constant rates $R_{12}$ and $R_{12}$, the end result does not hold. The full non-LTE problem would therefore require time dependent rates. That will increase the complexity of the calculations significantly. Therefore, the approximation for the populations given above is utilized.

With the known time dependent populations $n_1(t)$ and $n_2(t)$, the additional energy density at point $r$ due to absorption can be estimated by subtracting from (142) the LTE energy density, i.e.,

$$J_n(T, r, t, f) = B_f \frac{n_2(t)}{n_1(t)} \frac{n_1(n_2(n_1)^{-1} \exp \left( -\frac{hf}{kBT} \right) - B_f.}$$ (154)

The validity of this excess energy can be shown at the boundaries. When time approaches infinity, it is easy to see that (154) is zero by looking at (150) and (151). When, on the other hand, time is zero, it can be shown that the energy in the system is the same as the absorbed energy at point $r$. This can straightforwardly be shown by calculating the equality

$$B_f \frac{n_2(0)}{n_1(0)} \frac{n_1(n_2(n_1)^{-1} \exp \left( -\frac{hf}{kBT} \right) - B_f = \frac{d}{dr} \left( 1 - e^{-\kappa_0(f)r} \right) \right)}{4\pi r^2}. (155)$$

This is easy to confirm by simple manipulations and with the relations given by (144) and (145). This proves that the approach is correct at the boundaries. However, this is only a partial validation. The energy decay is hard to validate, mainly because of the absence of the measurement data on the noise.
Appendix 2 Proof of the Equality of the Two Multiple Scattering Models

Proof that the pulse theory in [57] can be used to extend the two dimensional impulse model in [58] into channel model for pulse based communications. This is done by modifying the impulse theory in [58] with pulse theory in [57] and then by showing that the output of the both models is identical, thus the models are identical and the extension can be done for the model in [58]. The pulse theory in [57] is

\[ I_F(r,t) = \frac{e^{-r\kappa}}{2\pi r} + \frac{\sigma\kappa}{2\pi} \int \frac{\exp \left( -\kappa(R - \sigma\sqrt{R^2 - r^2}) \right)}{\sqrt{R^2 - r^2}} \Theta(R - r) dR. \]  

(156)

The modified model for pulse theory from [58] is achieved by integrating the 2D equation over the total distance traveled by a photon \( R \), i.e., over \( ct \). The integration limits are considered in the text. For proofing purposes, a general integral is enough. The pulse version of the theory in [58] reads

\[ I_P(r,t) = \frac{e^{-r\kappa}}{2\pi r} + \int \frac{e^{-\frac{R}{2\sigma}}}{2\pi\sigma R} \left( 1 - \frac{R^2}{2} \right)^{-\frac{1}{2}} \]  

\[ \times \exp \left( \frac{1}{\sigma^2} \left( \sqrt{R^2 - r^2} - R \right) \right) \Theta(R - r) dR, \]  

(157)

If \( I_F(r,t) = I_P(r,t) \), the proof is evident for \( R < r \), since in this region \( I_F(r,t) = I_P(r,t) = 0 \). The proof is also evident in the case ballistic term (\( R = r \)), since the ballistic terms are identical, i.e.,

\[ I_F(r,t = \frac{r}{c}) = I_P(r,t = \frac{r}{c}) = \frac{e^{-r\kappa}}{2\pi r}. \]  

(158)

thus

\[ I_F(r,t) = I_P(r,t), \quad \forall \ t \leq \frac{r}{c}. \]  

(159)

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In the case $R > r$ ($I^F(r,t) = I^P(r,t)$),

\[
\frac{\sigma \kappa_e}{2\pi} \int \frac{\exp\left(-\kappa_e(R - \sigma \sqrt{R^2 - r^2})\right)}{\sqrt{R^2 - r^2}} \Theta(R - r) dR
\]

\[
= \int \frac{e^{-\frac{R}{2}}}{2\pi l_s R} \left(1 - \frac{R^2}{R^2}ight)^{-\frac{1}{2}} \exp\left(\frac{1}{l_s} \left(\sqrt{R^2 - r^2} - R\right)\right) \times \Theta(R - r) dR.
\]

(160)

Since the integrals are expected to be equal, the integration is removed from the both sides. Also, $(1 - r^2/R^2)^{-1/2} = R/\sqrt{R^2 - r^2}$. Thus, $\Theta(R - r)/(2\pi \sqrt{R^2 - r^2})$ can be subtracted from the both sides. Then

\[
\frac{\sigma \kappa_e \exp\left(-\kappa_e(R - \sigma \sqrt{R^2 - r^2})\right)}{\sqrt{R^2 - r^2}}
\]

\[
= \frac{e^{-\frac{R}{2}}}{l_s} \exp\left(\frac{1}{l_s} \left(\sqrt{R^2 - r^2} - R\right)\right).
\]

(161)

Now the both sides can be multiplied with $l_s$. Since $\sigma = \frac{1}{l_s}/(\frac{1}{l_s} + \frac{1}{l_a})$ and $\kappa_e = \frac{1}{l_s} + \frac{1}{l_a}$, the multiplier $\sigma \kappa_e$ equals unity (after the multiplication with $l_s$). By slightly rearranging the equations:

\[
\exp\left(-\kappa_e R - \kappa_e \sigma \sqrt{R^2 - r^2}\right)
\]

\[
= \exp\left(-\frac{R}{l_a} + \frac{1}{l_a} \left(\sqrt{R^2 - r^2} - R\right)\right).
\]

(162)

It is easy to show that the contents of the exponent functions are equal, thus the proof is fulfilled as

\[I^F(r,t) = I^P(r,t), \quad \forall t = \frac{R}{c} \quad \square.\]

(163)
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