Analysis, Measurement, and Modeling of Millimeter Wave Channels for Aviation Applications

Zeenat Afroz

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ANALYSIS, MEASUREMENT, AND MODELING OF MILLIMETER WAVE CHANNELS FOR AVIATION APPLICATIONS

by

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DEDICATION

I dedicate my dissertation to my mother Rabeya Akter, my husband Talukder Golam Rabby and my son Talukder Farhan Rayeed. Their unconditional love, supports and advice encouraged me to earn my Ph.D.
ACKNOWLEDGEMENTS

First and foremost, I would like to express my gratitude to my advisor Dr. David Matolak for his guidance and support throughout my Ph.D. journey. I would also like to thank my Ph.D. committee members, Dr. Mohammad Ali, Dr. Alphan Sahin, and Dr. Sanjib Sur, for their valuable feedback. Many thanks to Dr. M. Cenk Erturk, Dr. Hosseinali Jamal, Dr. Nozhan Hosseini, Dr. Jinwen Liu, Dr. Mohanad Mohsen, Patrick Murphy, Hudson Dye, Eddie King, Nathan Stofik, Isaias Ortiz Bautista, Safi Shams Muhtasimul Hoque, and Mohammad Hassan Adeli for their assistance.
ABSTRACT

Millimeter wave (mmWave) communication systems can employ a large amount of spectrum, and can consequently offer large data rates, e.g., multi-Gigabits-per-second. This technology can be used in many sectors: aviation, vehicles, public transportation, robotics, autonomous factories, etc. Yet mmWave communication systems suffer from some propagation challenges, including large free space path loss (PL), large penetration loss, and large diffraction loss. Hence, it is vital to quantify these and other channel effects to ensure link reliability. Most mmWave systems will employ directional antennas to enable acceptable link distances. In many settings this will require directional receiver antennas to rotate in azimuth to capture the strongest received signal in non-line-of-sight (NLOS) regions. Although terrestrial mmWave communication systems are advancing steadily, mmWave communication system applications in aviation and airport settings are still in their infancy. As commercial aviation is experiencing rapid growth, mmWave bands can be used for short range applications within airport areas, which can support a large robust data transfer. For this reason, research on mmWave wireless channel characterization for the aviation environments is a nascent area of study. In this dissertation, we present our measurement results for the 90 GHz band for two different types of settings: non-aviation public areas (indoor hallways and outdoor streets), and aviation settings (an airport maintenance hangar and airport baggage claim areas). We address line-of-sight (LOS), NLOS, mixed LOS/NLOS, and LOS-to-NLOS transitions
settings, and provide channel statistics and models based upon extensive measurements using a 500 MHz bandwidth chirp signal. For all measurements, the transmitter (Tx) was fixed and the receiver (Rx) was moved. Emulating what an operational mmwave system will do, the Rx antenna was rotated to obtain maximum received power when the Rx was partially or fully obstructed. The multi path components (MPCs) measured allow us to compute the most common measure of dispersion, root mean-square delay spread (RMS-DS), as well as the spatial PDP correlation coefficient, used to assess statistical stationarity. We corroborated our results with geometric analysis and ray-tracing simulations.

For non-aviation environments, we estimated parameters for the close-in free space reference distance PL model using both simulated (ray-tracing software) and measured data. PL exponents are 1.6 for outdoor and 1.8 for indoors, smaller than for free space because of waveguiding. We observed rapid PL changes in the LOS to NLOS transition regions, 15 dB/20 cm for indoor and 15 dB/10 cm for outdoor. Abrupt changes of the strongest-component angle of arrival (AoA), up to 60 degrees over a few cm were also observed. Very small stationarity distances, as small as six wavelengths, were found in NLOS settings. For our second environment type, aviation settings, we measured 90 GHz channel characteristics in the Jim Hamilton L.B. Owens Airport (CUB), Columbia, SC, USA, and Columbia Metropolitan Airport (CAE), West Columbia, SC, USA. At CUB, we measured in an atypical environment, the airport maintenance hangar. This crowded environment contained multiple aircraft, metallic objects, and other obstacles, whose positions moved throughout the day; thus we characterized this setting as a mixed LOS/NLOS environment. For this mixed setting the pathloss exponents exceed that for free space, with a maximum of 3.14. As the Rx is surrounded by several rich reflectors, the range
of AoA at CUB is larger than 104°, larger than in typical office environments. We also quantified the RMS DS, finding a maximum value of 24 ns for the NLOS case. The minimum stationarity distance for both LOS and NLOS settings at CUB maintenance hangar is 0.5 m. Aviation-setting measurements were also conducted at CAE baggage area for LOS, mixed, and LOS to NLOS transition settings. The CI model PL exponent is close to the free space PL exponent, as expected in this open setting. A relatively large RMS DS was observed in both LOS and NLOS settings: 20.5 ns for LOS and 23 ns for NLOS. The maximum LOS RMS DS in aviation environments is approximately double that of our non-aviation scenarios. These results will be of use to mmWave network designers in this unique aviation application.
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<tr>
<td>5G</td>
<td>Fifth-generation</td>
</tr>
<tr>
<td>6G</td>
<td>Sixth-generation</td>
</tr>
<tr>
<td>ABG</td>
<td>alpha-beta-gamma</td>
</tr>
<tr>
<td>AoA</td>
<td>Angle of arrival</td>
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<tr>
<td>AoD</td>
<td>Angle of departure</td>
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<tr>
<td>AS</td>
<td>Angle spread</td>
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<tr>
<td>BSs</td>
<td>Base stations</td>
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<td>BW</td>
<td>Bandwidth</td>
</tr>
<tr>
<td>BPF</td>
<td>Band pass filter</td>
</tr>
<tr>
<td>CI</td>
<td>Close-in</td>
</tr>
<tr>
<td>CIR</td>
<td>Channel impulse response</td>
</tr>
<tr>
<td>DS</td>
<td>Delay spread</td>
</tr>
<tr>
<td>FAA</td>
<td>Federal Aviation Administration</td>
</tr>
<tr>
<td>FI</td>
<td>Floating-intercept</td>
</tr>
<tr>
<td>FSPL</td>
<td>Free space path loss</td>
</tr>
<tr>
<td>HPBW</td>
<td>Half power beam width</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate frequency</td>
</tr>
<tr>
<td>KED</td>
<td>Knife-edge diffraction</td>
</tr>
<tr>
<td>LOS</td>
<td>Line-of-sight</td>
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</table>
LO ................................................................. Local oscillator
MIMO ............................................................. Multiple-input multiple-output
mmWave ............................................................ Millimeter Wave
MPCs ................................................................. Multipath components
NASA .............................................................. National Aeronautics and Space Administration
NLOS ............................................................... Non-light-of-sight
OEW ................................................................. open ended waveguide
PDPs ................................................................. Power delay profiles
PL ................................................................. Path loss
QSR ................................................................. Quasi-stationarity regions
RIS ................................................................. Reconfigurable Intelligent Surface
RMS-DS ............................................................. Root-mean-square delay spread
Rx ................................................................. Receiver
RT ................................................................. Ray tracing
RF ................................................................. Radio frequency
SPCC ............................................................... Spatial power delay profile correlation coefficients
SD ................................................................. Standard deviation
SSA ................................................................. Signal and spectrum analyzer
SNR ................................................................. Signal-to-noise ratio
Tx ................................................................. Transmitter
UTD ................................................................. Uniform theory of diffraction
UE ................................................................. User equipment
VDA ............................................................... V-dipole antennas
VSG.......................................................... Vector signal generator
WI................................................................. Wireless Insite
XPD.............................................................. Cross polarization discrimination
CHAPTER 1
INTRODUCTION

1.1 BACKGROUND: MILLIMETER WAVE CHANNEL CHARACTERIZATION AND MODELING

The requirement of cellular data traffic is growing at a rate of 40–70% per year [1]. The capacity of the new fifth generation (5G) wireless communication systems should be increased within the next decade by a factor of 1000 times relative to current levels [2]. The 5G capacity gains are expected to be provided in different ways, e.g., using multiple-input multiple-output (MIMO) antennas [3], making small cell coverage zones for greater network densification [5], using new modulation waveforms, and using wideband (> 100 MHz) channels, multi-user, three-dimensional (3D) MIMO and large channel bandwidths [4]. “Massive” MIMO antenna arrays and smaller arrays can be used at base stations (BSs) and the mobile user equipment (UE), respectively. Any new waveforms will offer the smallest capacity increase, whereas the wider channel bandwidths, with links using directional antennas, will provide the largest capacity increase. As the existing microwave bands are already congested and allocated for other purposes, additional spectrum cannot come from this. Hence, new millimeter-wave (mmWave) spectral bands within the 30–300 GHz frequency range are being considered for 5G and beyond, because of the ample amount of unused spectrum relative to the microwave bands [6]. The World Radio
Conference in 2015 (WRC-15) approved several candidate bands for 5G: 24.25–27.5 GHz, 31.8–33.4 GHz, 37–43.5 GHz, 45.5–50.2 GHz, 50.4–52.6 GHz, 66–76 GHz, and 81–86 GHz. In addition to these bands, the 60 GHz band, 90 GHz band, and 118 GHz bands may also be used.

Ultra-high-definition streaming, wireless cognition, and centimeter-level position location applications will increase indoor data traffic. Indoor mmWave systems can be effectively isolated from outdoor co-channel cellular systems due to large outdoor-to-indoor (O2I) penetration loss of up to 60 dB at mmWave frequencies. As mmWaves have short wavelength, they do not diffract well and are sensitive to the dynamic blockage of the human body as well [7], [8]. To overcome the additional path loss (PL), directional and steerable high gain antennas with beamforming techniques are essential [9]. Hence, time-variant directional channel models are vital for proper system design.

Statistical channel models up to 100 GHz are presented by IEEE 802.11 ad/ay and 3GPP TR 38.901 for indoor scenarios, e.g., residence, office, shopping mall, and factory [10] - [12]. In addition, IEEE 802.11 ad/ay channel models adopted a double-directional Channel Impulse Response (CIR) model based on field measurements and complementary ray-tracing simulations for 60 GHz with dual polarizations. These channel models provide temporal and angular channel statistics for conference rooms, cubical environments, and living rooms [10], [11]. Moreover, a unified geometry-based statistical channel model for indoor and outdoor scenarios has been proposed by the 3GPP TR 38.901 for frequencies from 0.5 to 100 GHz. There are different values of large-scale parameters, i.e., delay spread (DS), angle spread (AS), Rician K factor, and shadow fading (SF) standard deviation for different scenarios, each of which is required in the channel generation procedure [12]. The
estimation of channel characteristic variation over wide frequency ranges is useful for futuristic applications such as intelligent reflecting surfaces [13] and precise position location [14]. In non-line-of-sight (NLOS) regions, the transmitter (Tx) and the receiver (Rx) antennas are pointing in the best direction (possibly to reflectors) to enable the Rx to capture the maximum received power [15], [16], [17]. Despite the fairly large amount of existing work on mmWave channels, because of the large variety of mmWave applications, and the continuing demand for more reliable links, the study of mmWave channel characteristics is still an active area of research.

Because of the complexity, expense, and challenges of channel measurement campaigns, in addition to analytical models, much effort has been expended to develop accurate computer simulation-based channel models. Ray-tracing (RT) models have become a favored type of such models, and we also explore their use in this dissertation. The RT model is essentially a functional database, where each database element is represented by its coordinates and electrical properties. RT tools are based on the geometry of the deployment scenario, and these high frequency approximations contrast with analytical or stochastic models, by incorporation of much more site-specific information. An RT model can provide greater accuracy than analytical models by characterizing each MPC’s propagation properties: time delay, Doppler shift, polarization, angle of departure (AoD) at the Tx, and angle of arrival (AoA) at the Rx [18]. In comparison to microwave bands, Doppler frequency is larger and consequently the coherence time is small because of the large frequency of mmWaves. To compensate, directional antennas reduce fading characteristics, but such reductions are highly dependent on AOA/AOD. Simulators based on ray tracing, e.g., WinProp(R) and Wireless InSite(R) (WI) software, can model the
temporal and spatial evolution of the channel, a necessary feature for future wireless systems. Still, accurate and wide area RT models can be computationally expensive, hence limiting the scalability of simulations. Many RT techniques have been developed, e.g., in indoor channels for specific propagation areas [19], [20]. Finally on this, RT models must always be validated with empirical data.

1.2 MOTIVATION OF AVIATION 90 GHZ BAND CHANNEL MODELING

Wideband mmWave systems offer many opportunities for a dynamically adaptive secure network. Millimeter waves have been a research focus over the recent few years for use in fifth generation (5G) wireless communication systems. Hence, many research groups and industry entities have been developing channel models at mmWave frequencies. Over the past decade, researchers have conducted numerous mmWave channel measurement campaigns to quantify characteristics at several mmWave frequencies: 28 GHz, 39 GHz, 60 GHz, 73 GHz, and several others. Although some measurements have been done at higher mmWave frequencies, not as much has appeared for the 90 GHz band, a portion of which has been allocated for indoor applications [21]. This research is focused on the 90 GHz band for both indoor and outdoor scenarios. We quantify channel characteristics for both indoor and outdoor settings in the 90 GHz band, focusing on line-of-sight (LOS)-to-NLOS transitions, and airport applications. These channel transitions can present some of the most challenging conditions to link reliability, which are increased at mmWave bands relative to channels at lower frequencies. Example features within the topic of transitions that we quantify include the range and rate of change of angle of arrival of the strongest MPCs, root mean-square delay spread (RMS DS), and statistical stationarity distance. In indoor and outdoor environments, propagation measurements at mmWave frequency bands
are necessary for creating statistical channel models, which play a vital role for helping develop new standards and technologies for wireless communications systems. In addition, channel models at mmWave frequencies that predict signal strength and multipath time delays and angles will be required for effective modem and system design.

Unprecedented global growth of the aviation industry demands new communication techniques to support and manage a larger number of aircraft, both in the air, and at airports. Research outcomes on mmWave channels can be used in designing airport network systems that will increase communication capacity & security for a wide range of applications. Although the terrestrial application of mmWave systems is advancing at a rapid pace, the use of mmWave communication systems in aviation systems or airports is still in its infancy. In this regard, characterization of mmWave links and accurate, quantitative mmWave channel models are needed to enable various links in the airports. Therefore, we conducted channel measurements both in a small airport, the Jim Hamilton–L.B. Owens Airport, in Columbia, SC and at the Columbia Metropolitan Airport, West Columbia, SC. Subsequently, we developed models and quantified channel parameters that will benefit consumers (the flying public) as well as airlines, regulatory bodies, and airport operation systems, and in turn the growth of the aviation industry and the US economy as a whole.

1.3 DISSERTATION SCOPE

In this dissertation, measurement results for the 90 GHz band using a 500 MHz bandwidth chirp signal have been provided. The measurement settings are both LOS and NLOS: outdoor streets, indoor hallways, an airport maintenance hangar and an airport baggage area. Path Loss (PL) models are developed, and characteristics of LOS-to-NLOS
transitions are quantified. Broadly speaking, the scope of the dissertation is the quantification of key channel characteristics for 90 GHz band mmWave wireless systems in some indoor and outdoor settings, leading to complete channel characterization for several office and airport environments. This characterization includes empirical (and RT simulation) models for propagation PL, root mean-square delay spread (RMS-DS), and statistical stationarity distance, with a focus on channel transitions.

The dissertation is organized as follows:

1. [Chapter 2] PL basics and review of the literature relevant to mmWave channel characterization are described in chapter 2. The ray tracing concept is also explained.

2. [Chapter 3] In chapter 3, 90 GHz channel measurement equipment, channel sounding processing, calibration, and ray tracing simulation basics, with an example result, are provided.

3. [Chapter 4] Chapter 4 contains a detailed discussion of the measurement environments and measurement procedures.

4. [Chapter 5] In chapter 5, for LOS settings, results from measured and RT software are described and compared for indoor hallways and outdoor street environments. Channel parameters for LOS to NLOS transition settings, e.g., RMS DS, AOA, stationarity distance, etc., are explained in detail.

5. [Chapter 6] Results of unusual indoor environments, specifically an airport maintenance hangar and an airport baggage area, are analyzed in detail in chapter 6. Channel parameters are computed for LOS, NLOS, LOS to NLOS transition and mixed (combined LOS and NLOS) environments.
6. [Chapter 7] Conclusions and suggestions for future research are contained in chapter 7.

1.4 DISSERTATION OBJECTIVES

This dissertation focuses on channel characterization for the 90 GHz band. The objectives of the dissertation are as follows:

1. Develop analytical, empirical and RT simulation channel models for wide band channels in multiple environments (indoor hallways and outdoor streets), both LOS and NLOS. Models include PL, as well as power delay profiles (PDPs); we compare our results to existing and related work, highlighting differences.

2. Quantify LOS to NLOS channel transition statistics for both indoor and outdoor settings (indoor hallways and outdoor streets), e.g., DS, rate of change of AoA, and spatial PDP correlation coefficients (SPCC).

3. Construct empirical models for wide band 90 GHz channels in example airport environments, e.g., an airport maintenance hangar, and an airport baggage area, for LOS, NLOS, LOS to NLOS transition, and mixed LOS-NLOS environments.

1.5 DISSERTATION CONTRIBUTION

NASA sponsored the project, “Hyper-Spectral Communications and Networking for Air Traffic Management (ATM),” (HSCNA), which supported this dissertation work. The project duration was 2017 to 2022. Three IEEE conference papers have been published to date. One journal paper is under review and one journal paper is in preparation. Publications from this dissertation work are as follows.

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1 The notation C denotes conference paper, J denotes journal paper and P denotes poster.


Two posters were also presented:


CHAPTER 2
MILLIMETER WAVE CHANNEL MEASUREMENT AND MODELING LITERATURE REVIEW

2.1 PATH LOSS

Path loss or attenuation is the gradual reduction in power density of a radio wave as it propagates through a medium. Path loss estimation is important to ensure an acceptable quality of service and provide coverage over the desired geographic area by placing transmitters at appropriate locations with appropriate transmit power. The free space path loss (FSPL) equation is derived under the far field assumption: propagation distance between Tx and Rx \( d \gg \) largest physical dimension of the antenna \( D \), \( d \gg \) Friis distance \( d_f \), \( d_f > \left( \frac{2D^2}{\lambda} \right) \), \( d_f \gg \lambda \), and \( d_f \gg D \). Under these assumptions, the FSPL is given by (2.1):

\[
L_f = 20 \log_{10} f + 20 \log_{10} d - 147.6 \quad (2.1).
\]

Here, \( f \) = frequency, and \( \lambda \)=Wavelength.

Multipath components, which are delayed from each other in time, and generally incur different attenuation, are received at the Rx due to reflection, diffraction, and refraction. Figure 2.1 shows a two ray channel model: (i) a LOS ray and, (ii) a reflected ray from the ground.
For this 2-ray model, the total received electric fields (initially assuming isotropic antennas) are represented by (2.2).

\[ E = E_0 e^{-j\beta R_1} + E_1 e^{-j\beta R_2} = E_0 e^{-j\beta R_1} (1 + \rho e^{-j\Delta}) \] (2.2)

Here, \( E_0 \) = electric field for direct free space wave,

\( E_1 \) = electric field for reflected wave,

\( \rho \) = reflection coefficient of the earth,

\( \Delta \) = phase difference between the direct path and the ground-reflected path, and

\( \beta = \frac{2\pi}{\lambda} \) = Wave number.

\( R_1 \) and \( R_2 \) can be found by trigonometry, using (2.3) and (2.4), respectively.
\[ R_1 = \sqrt{d^2 + (h_T - h_R)^2} \] (2.3)

\[ R_2 = \sqrt{d^2 + (h_T + h_R)^2} \] (2.4)

Here, \(d\) = distance between Tx and Rx,

\(h_T\) = Tx antenna height, and

\(h_R\) = Rx antenna height.

Thus, the magnitude of the electric field and the received power \(P_R\) are represented by (2.5) and (2.6), respectively.

\[ |E| = |E_0| \left| \left( 1 + \rho e^{-j\Delta} \right) \right| \] (2.5)

\[ P_R = |E|^2 \frac{\lambda^2 G_R}{4\pi\eta} \]

\[ = \left( \frac{\lambda}{4\pi d} \right)^2 \left| \left( 1 + \rho e^{-j\Delta} \right) \right|^2 P_T G_T G_R \] (2.6)

Here, \(P_T\) = transmitted signal power,

\(G_T\) = Tx antenna gain,

\(G_R\) = Rx antenna gain,

\[ |E_0|^2 = \frac{30P_T G_T}{d^2}, \] and

\(\eta\) = characteristic wave impedance of free space.

If directional antennas are used, then \(G_T, G_R\) and \(\rho\) depend on incident angle of the rays, and will be different for the LOS and reflected rays. Reflection coefficients of
perpendicular and parallel polarization can be determined using (2.7) and (2.8), respectively [22].

\[
\rho_\perp = \frac{\cos \theta_i - \frac{\varepsilon_2}{\varepsilon_1} \sin^2(\theta_i)}{\cos \theta_i + \frac{\varepsilon_2}{\varepsilon_1} \sin^2(\theta_i)} \tag{2.7}
\]

\[
\rho_\parallel = \frac{-\frac{\varepsilon_2}{\varepsilon_1} \cos \theta_i + \sqrt{\varepsilon_2 \varepsilon_1 - \sin^2(\theta_i)}}{\frac{\varepsilon_2}{\varepsilon_1} \cos \theta_i + \sqrt{\varepsilon_2 \varepsilon_1 - \sin^2(\theta_i)}} \tag{2.8}
\]

Here, \(\theta_i\) = incident angle,

\(\varepsilon\) = permittivity of a lossy medium is denoted by \(\varepsilon = \varepsilon_0 \varepsilon_r + j\frac{\sigma}{\omega}; \varepsilon_0\) = vacuum permittivity, \(\varepsilon_r\) = relative permittivity, and \(\sigma\) = conductivity.

For grazing incidence, i.e., \(\theta_i = 90^\circ\), \(\rho_\perp = -1\) and \(\rho_\parallel = 1\). If \(d >> h_T, h_R\) and for grazing incidence, the 2-ray PL model is represented by (2.9).

\[
L_p = -10 \log_{10} G_T - 10 \log_{10} G_R - 20 \log_{10} h_T - 20 \log_{10} h_R + 40 \log_{10} d \tag{2.9}
\]

This two-ray model is applicable to determine PL in environments consisting of no obstructions between the Tx and the Rx. The two ray model can apply between a ground station and an airborne platform where both LOS and ground reflected components suffer from impairment, e.g., flat fading and frequency selective fading.

In a general environment, the Rx receives more than two components that result in a more complex interference pattern among the components; the number of multipath components depends on several factors: frequency band, medium of transmission and of course the environment. In terrestrial communication, a receiver on the street, in a factory space, in a large room or in a shopping mall will typically receive many multipath
components due to the presence of the fixed (or moving) man-made structures, e.g., buildings, bridges, tunnels, hills, mountains, trees, foliage, vehicles, trains, trams or people. All MPCs occur due to reflection, scattering, diffraction from these structures and refraction through the atmosphere.

2.2 MILLIMETER WAVE SMALL SCALE CHANNEL CHARACTERISTICS

Millimeter wave channels are sparse in the angle and delay domains [23]. Millimeter wave signals suffer a large FSPL and diffraction loss. The root mean square delay spread is one of the most popular methods of quantifying channel time dispersion. Inter symbol interference can be caused by a large delay spread relative to the transmitted symbol duration. As we use directional antennas for mmWave communications, fewer MPCs are captured by the Rx compared to the case with omnidirectional antennas. For this reason, DS can be reduced. The spatial consistency of a channel defines the smooth spatial characteristics of the channel when the Tx or the Rx moves through a space. It is expected that channel characteristics for closely located users are highly correlated. Spatial consistency covers different large-scale parameters, small-scale parameters of delays, AoAs and AoD. If the correlation coefficient of the targeted channel properties is above a certain threshold, the spatial extent of this is defined as a stationarity region. The channel stationarity plays a vital role in channel modeling and estimation, and is useful to design experiments/simulations to obtain accurate estimates and reproduce channel parameters in modeling.

The authors in [24] quantified spatial stationarity regions in three bands in a basement environment (NLOS scenario). It was observed that the spatial stationarity regions of mmWave bands were much smaller than those at UHF/microwave frequencies;
stationarity distances were found to be 0.05 m, 0.03 m, and 0.48 m, for 14-16 GHz, 28-30 GHz, and 2-4 GHz, respectively. The directional antennas’ orientation with respect to the surrounding environment can also affect the stationarity and correlation distances. Hence, delay spread and stationarity regions should be carefully characterized in 5G channel modeling.

2.3 RAY TRACING CONCEPT AND SIMULATION

Ray tracing is a technique that uses a “high frequency” approximation, in which all obstacles in the environment are assumed large with respect to a wavelength. Although the assumption is not always valid, the tracing of “rays” (along the propagation vectors) can account for most channel effects in settings with several large obstacles. This is the most important propagation modeling approach to provide first-order designs for implementation of future mmWave wireless network infrastructure implementations, as we often observe highly reflective channels in both indoor and outdoor scenarios. We can use simulation based on ray tracing techniques for any environment to model and discretize the energy radiated in space. A radiating source is modeled as a discrete source having variable radiation amplitudes over a sphere; see Figure 2.2. A fractional spatial portion of radiated energy from (or to) an antenna is represented by ray tubes. This is sometimes known as brute force ray tracing [25].

The physical environment’s database is a vital component of ray tracing engines, because accurate physical dimensions and environmental electromagnetic properties play a crucial role in their accuracy. The method of images can be used instead of “brute force” ray tracing for simple physical environments, i.e., identifiable specific reflection points in the physical model. The ray tracing-based model has advantages: after the model is
constructed, we can easily change carrier frequency, transmit and receive antennas, and physical environmental models.

![Ray tracing techniques](image)

Figure 2.2. Ray tracing techniques, from left to right: radiation discretization from the transmitter and ray tubes representing each ray [25].

Here we will briefly discuss received power considering multiple rays.

The received LOS signal can be represented as (2.10) [26],

$$R_0(n) = \frac{\lambda}{4\pi d_0} \sqrt{G_T(\theta^{Tx}, \phi^{Tx}) G_R(\theta^{Rx}, \phi^{Rx}) S(n - \tau_0) e^{-\frac{j2\pi d_0}{\lambda}} |\rho_0^{Tx} \cdot \rho_0^{Rx}|}.$$  (2.10)

Here, $S$ is the transmitted signal,

$G_T(\theta^{Tx}, \phi^{Tx}) =$ gain of Tx antenna at elevation and azimuth angle of $\theta^{Tx}$ and $\phi^{Tx}$, respectively,

$G_R(\theta^{Rx}, \phi^{Rx}) =$ gain of Rx antenna at elevation and azimuth angle of $\theta^{Rx}$ and $\phi^{Rx}$, respectively,

$$\tau_0 = \frac{d}{c} = \text{LOS component delay}; \ c = \text{speed of light}, \ d = \text{distance of the LOS component},$$
\[ e^{-\frac{j \pi d_0}{\lambda}} \] is the phase term of the LOS component, \( \rho^T_x, \rho^R_x \) = dot product between the polarization unit vectors of the Tx and the Rx electric field, respectively.

The gain term for the antennas is given by (2.11),

\[
\sqrt{G_T(\theta^T_x, \phi^T_x)G_R(\theta^R_x, \phi^R_x)} = g^{(T, \theta)}(\theta^T_x, \phi^T_x)g^{(R, \theta)}(\theta^R_x, \phi^R_x) +
\]

\[
g^{(T, \phi)}(\theta^T_x, \phi^T_x)g^{(R, \phi)}(\theta^R_x, \phi^R_x). \quad (2.11)
\]

Here, \( g^{(T, \theta)}(\theta^T_x, \phi^T_x) \) and \( g^{(T, \phi)}(\theta^T_x, \phi^T_x) \) represent AoD in the elevation and azimuth plane, respectively, and \( g^{(R, \theta)}(\theta^R_x, \phi^R_x) \) and \( g^{(R, \phi)}(\theta^R_x, \phi^R_x) \) represent AoA in the elevation and azimuth plane, respectively.

\( g^\theta(\theta, \phi) \) can be expressed as (2.12)

\[
g^\theta(\theta, \phi) = \sqrt{|G_\theta(\theta, \phi)|}e^{j\phi_\theta} \quad (2.12)
\]

Here, \( G_\theta \) = antenna gain, and

\( \phi_\theta \) = relative phase of the \( \theta \) component of a ray.

Total gain mentioned in (2.12) should be maximized if both Tx and Rx are aligned to their boresight.

Like the LOS component, the multiple dominant received rays reflected from the environment, with ray index \( i = 1, 2, 3, \ldots M \), can be expressed as (2.13).

\[
R_i(n) = \frac{X_i(\psi_i)}{4\pi d_i} \sqrt{G_T(\psi_i^{Az}, \psi_i^{El})G_R(\psi_i^{Az}, \psi_i^{El})S(n - \tau_i)e^{-\frac{j2\pi d_i}{\lambda}}|\rho^T_x, \rho^R_x|} \quad (2.13)
\]
Here, $\Gamma_i(\psi_i) = \frac{\sin(\psi_i) - Y}{\sin(\psi_i) + Y}$ = Fresnel reflection coefficient for material $\varepsilon_r$; for vertical polarization $Y_v = \frac{\sqrt{\varepsilon_r - \cos^2(\psi_i)}}{\varepsilon_r}$, and for horizontal polarization $Y_h = \sqrt{\varepsilon_r - \cos^2(\psi_i)}$. The total received power considering non-coherent addition of LOS and reflected rays are represented by (2.14).

$$P_R(d_i) = E[|R_0(n) + \sum_{i=1}^{M} R_i(n)|^2] \quad (2.14)$$

$$P_R(d_i) = E[|R_0(n)|^2] + \sum_{i=1}^{M} E[|R_i(n)|^2]$$

If, $S(n) \approx S(n-\tau_0) \approx S(n-\tau_i)$, $|\rho_i^{Tx}.\rho_i^{Rx}|=1$, and $P_T = E[|S(n)|^2]$=transmitted power, the total received power can be expressed as (2.15).

$$P_R(d_i) = P_T \left(\frac{\lambda}{4\pi}\right)^2 \left|\frac{G_T(\theta^{Tx},\phi^{Tx})G_R(\theta^{Rx},\phi^{Rx})}{d_0}\right| + \sum_{i=1}^{M} \frac{\Gamma_i(\psi_i)\sqrt{G_T(\psi_i^{Az},\psi_i^{El})G_R(\psi_i^{Az},\psi_i^{El})e^{-j\Delta\Omega_i}}}{d_i} \quad (2.15)$$

Here,

$$\Delta\Omega_i = \frac{2\pi(d_i-d_0)}{\lambda} \text{ for } i=1, 2, 3, \ldots M.$$ 

Five dominant rays i.e., LOS, and rays reflected from ground, ceiling, and two walls can often be satisfactory for indoor settings, whereas in open outdoor settings, two dominant rays can be considered [26]; see Figure 2.3.
Wireless InSite® software, an electromagnetic simulation tool, is used to predict the effects of buildings and terrain on the propagation of electromagnetic waves. This software performs the electromagnetic calculations, and then evaluates the signal propagation characteristics. We can either construct the environment by using WI’s editing tools or import the virtual building and terrain environment using several file formats, e.g., DXF, shapefile, DTED and USGS. In addition, the study area can be defined to specify separate calculations for portions of the overall area. The calculations are made by launching rays from the Tx, interacting rays with geometrical features that include reflections from feature faces, diffractions around feature edges, transmissions through feature faces, and propagating them through the defined geometry to receiver locations.

Wireless InSite uses the Uniform Theory of Diffraction (UTD) to evaluate a ray path’s electric field, and provides accurate results for typical applications from approximately 100 MHz to approximately 100 GHz when the environment geometry is large compared to the propagating wave wavelength [27]. We use the X3D propagation...
model that includes atmospheric absorption, and this model is viable for its wave propagation calculations up to millimeter wave frequencies. Arriving ray paths are combined at each receiver location and some quantities can be predicted, e.g., electric and magnetic field strength, received power, interference measures, PL, DS, AoD, impulse response, electric field vs. time, electric field vs. frequency, and PDP. Wireless InSite shows visual representations of some results, placing visually within the modeled environment: transmitter coverage areas and power distributions. It also allows overlays of data and provides quick comparison to imported measurements, or even previous WI calculations.

2.4 MILLIMETER WAVE CHANNEL MEASUREMENT MODELING AND SIMULATION IN DIFFERENT ENVIRONMENTS

In this subsection, we cite some representative examples from the literature. Our results focus on the mmWave bands, but we also include some results in the microwave bands taken in airport environments, for comparison. In [17], results of typical indoor office environment measurements were reported for the 28 and 73 GHz frequency bands. The measurement signal bandwidth (BW) was 400 MHz, and rotatable directional horn antennas for both co-polarized and cross-polarized antenna configurations were used. The authors observed that the minimum values of RMS delay spreads considering the best beam are in general within 4 ns regardless of frequency, environment, and antenna polarization. This indicates that the “best” beam can simultaneously minimize PL and RMS delay spread.

The authors of [28] estimated channel DS and AS for 73 GHz, and created empirical channel models. They found that the indoor (office-type environment) model produced smaller RMS-DS and slightly larger azimuth spreads than the outdoor model. In [29], the
authors proposed a statistical PL model for 15 GHz in an indoor corridor environment incorporating Tx height. The authors observed that the average PL exponent was 1.93 and the standard deviation (SD) is closely related to Tx height. Values of Ricean $K$-factor for different locations were 5 to 10 dB.

In non-line of sight environments, penetration loss of obstructions plays a vital role. The authors of [30] and [31] performed measurements at 28 GHz and 40 GHz to study penetration losses for common obstructions such as wood, water, human hands, and leaves. Attenuations from metal and water can reach 30-40 dB. Typical building material attenuations in mmWave frequency ranges are of interest for link budget calculations.

In [32], the authors report on a collaborative measurement campaign to estimate attenuation of several typical building materials — clear glass, drywall (plasterboard), plywood, acoustic ceiling tile, and cinder blocks—at three mmWave frequencies (28, 73, 91 GHz) using directional antennas. The authors found that the specific attenuations range from approximately 0.5 dB/cm for ceiling tile at 28 GHz to approximately 19 dB/cm for clear glass at 91 GHz.

In [33], the authors provided penetration and reflection characteristics of common building materials (e.g., dry wall, plywood, etc.) at 60, 71, and 81 GHz. The reflection coefficients at these high frequencies were measured by varying incident angles to the surface of a material and by using both directional horn antennas and a narrow-band signal. Antenna polarization effects were also examined: higher path gain (45.4 dB) in co-polarization compared to cross-polarization in LOS scenarios, approximately 0.86 dB/cm larger average penetration loss of plywood for cross-polarization than for co-polarization. It was also noticed that the reflection coefficients are large at higher frequencies, from 0.38
to 0.83, and 0.33 to 0.76 for different incident angles (20’ to 60’) in 81 GHz and 60 GHz channels, respectively.

The authors of [34] took measurements at 28, 38, 60 and 73 GHz, and developed multipath models: as expected, more resolvable MPCs were obtained at 28 GHz than at 73 GHz. The authors of [35] investigated PL for urban micro-cellular and indoor office scenarios at 28 GHz and 73 GHz. They observed that the close in (CI) reference distance model and multi-frequency CI models with a frequency-weighted PL exponent (CIF) exhibit lower computational complexity than the single-frequency floating-intercept (FI) model and the multi-frequency alpha-beta-gamma (ABG) model. The CIF model is more appropriate for indoor modeling whereas the CI model was found to be suitable for outdoor environments over multiple frequencies. Moreover, PL increases with frequency in indoor environments, as expected. In contrast, PL shows little dependence on frequency beyond the first meter of free space propagation in outdoor scenarios.

The authors analyzed large scale fading characteristics, i.e., PL exponent, cross polarization discrimination ratio (XPR), standard deviation of shadow fading, etc., for the 45 GHz band in three different indoor scenarios in [36]: a conference room, a living room, and a cubicle office. The measurements were conducted with horn, an open ended waveguide (OEW), and V-dipole antennas (VDA), and they observed that average XPRs for horn, OEW, and VDA are 28.5, 18.5, and 3 dB, respectively, in LOS settings. The authors also noticed that a large beamwidth yields a large received power fluctuation, as intuition would predict.

In [37], the authors addressed frequencies of 5, 31, and 90 GHz and measured, modeled, and simulated channel characteristics for an indoor corridor environment.
Simulations were conducted via the X3D ray-tracing method in WI software. Their results indicate that measured and simulated CI PL model slopes were close, differing by less than 0.3, and model standard deviations differed by less than 2 dB for all frequencies for the LOS case. The differences are less than 0.6 for slope and 5 dB for standard deviations for NLOS channels.

The authors of [26] considered mmWave end-to-end propagation modeled by individual ray sources and showed that the two strongest (“dominant”) rays are sufficient to model the channel for outdoor open areas, whereas the indoor corridor requires more rays, e.g., 5 dominant rays, to yield a good fit between measurement and simulation results. It was also found that PL slope was very close to that of free space for an outdoor open area. In addition, indoor PL was small for their low gain antennas due to reception of multi-path components.

In [38], the authors present the measurement results and analysis obtained from a campaign conducted in a tunnel-like corridor in UESTC. They extended the PL model to a more general form based on the validated RT simulations, where the characteristics of the corridor structure, including the angle of the corridor corner and the length of the LOS section are considered as modeling parameters. In addition, a stochastic approach is used to model the effect of different Tx/Rx antenna locations on the cross-section of the corridor. The analysis shows that the power loss can be up to 30 dB if the corner angle between LOS and NLOS is 90° and the corner loss diminishes if the corner angle is 180°.

In [8] the authors showed knife-edge diffraction (KED) and a creeping wave linear model can predict diffraction loss around typical building objects from 10 to 26 GHz, and human blockage measurements at 73 GHz were used to fit a double KED model, which
incorporates antenna gains. The authors also showed how the received signal changes as the mobile moves and transitions from LOS to NLOS locations, with reasonably stationary signal levels within clusters in local area measurements. Wideband mmWave received signal levels were observed to fade from 0.4 dB/ms to 40 dB/s, depending on travel speed and surroundings.

The author of [39] presents an improved channel simulator NYUSIM 2.0 that can simulate spatially consistent channel realizations based on the existing drop-based channel simulator NYUSIM 1.6.1. To generate spatially correlated and time-variant channel coefficients, a geometry-based approach using multiple reflection surfaces is proposed. In addition, a four-state Markov model was implemented in NYUSIM using results from 73 GHz pedestrian measurements for human blockage to simulate dynamic human blockage shadowing loss. A parabolic model was adopted from the 5G Channel Modeling special interest group for outdoor-to-indoor penetration loss and implemented in NYUSIM 2.0 to reproduce channel data in Monte Carlo fashion.

The channel quasi-stationarity regions (QSR) for 28 GHz were quantified in a V2X generic high-dense urban scenarios in [40]. The authors validated their three-dimensional ray-launching (3D-RL) algorithm’s results with measurements. A decrease in the QSR span was observed in this mmWave frequency compared to sub-6 GHz schemes: the QSR spans were 0.4 m, 0.6 m, and 1 m for the mean correlation threshold of 0.9, 0.8, and 0.7, respectively.

Authors in [41] presented omnidirectional small-scale fading measurements and analysis at 73 and 81 GHz in an airport environment. They measured more than 825 PDPs at each mmWave band in an outdoor airport and an indoor campus environment in Boise.
airport. Their measured data shows that small-scale fading of received signal amplitudes follows both a Ricean and a log-normal distribution in the outdoor airport, both LOS and NLOS scenarios, at 73 and 81 GHz bands. In addition, for the 73 and 81 GHz outdoor channel, the cumulative distribution function shows Ricean behavior with K-factor ranging from 3–13 dB for LOS, and 21–27 dB for NLOS. For the 73 GHz LOS channel, a log-normal distribution is bounded by standard deviation $\sigma = 0.15$ dB and 0.37 dB.

The authors of [42] conducted measurement at Boise Airport and Boise State University and investigated the CI reference PL and FI PL models for both LOS and NLOS scenarios using high gain directional antennas. The authors found that the PL exponents are larger in the outdoor settings than in the indoor.

The authors of [43] provide measurement and models for propagation PL in two indoor airport environments—a typical small airport terminal building and a more unusual aircraft maintenance hangar—in two frequency bands (30 GHz and 5 GHz bands). They showed that LOS PL exponents of the airport terminal building are near the free-space value of two, and standard deviations are a few dB. They also observed that NLOS PL exponents are larger (~3-4), with larger standard deviations as well. In the hangar, for mixed environments and for both frequency bands, PL exponents are near three, and standard deviations are only 2.8 dB for 5 GHz, and 6.3 dB for 31 GHz.

We also describe results from a few measurements in aviation environments, and note that C band measurements are more prevalent than in mmWave bands. We summarize some C band measurement results here and illustrate the variation of different channel parameters as a function of different factors in airport environments: environment settings (LOS or NLOS), bandwidth, antenna height, etc. In [44] the authors describe results for the
airport surface environment in the 5-GHz band using a 50-MHz bandwidth test signal. Thousands of PDPs were obtained and processed to develop empirical tapped-delay line statistical channel models for large airports. The authors also developed a log-distance PL model. The median RMS DS ranged from 500 to 1000 ns for these airports, with the 90th percentile RMS delay spreads being approximately 1.7 µs. The most interesting findings are MPCs non-stationarity, with changing of component statistics over the propagation region and a substantial amount of severe (worse than Rayleigh) fading.

In [45], the authors provide results of the channel measurement campaign at Munich airport, in the C band using a test signal of 120 MHz BW. The highest value of 99th percentile delay spread of 1.5 µs estimated for taxiways, where 90th percentile values are very close to 1.7 µs for all three large US airports. They also observe that Doppler spread remains below 60 Hz in 99% of all cases for apron tracks. In shadowed areas, higher Doppler spread values have been measured but with limited confidence due to insufficient received power.

In [46], the authors described a channel measurement campaign at several smaller general aviation airports (Ohio University airport, Burke Lakefront airport in Cleveland, OH, and Tamiami (TA) airport, Kendall, FL.) in the 5 GHz band. They performed measurements using a 50-MHz bandwidth signal on the airport surface, and the Rx was moved in a mobile van. Measurements were computed with omni directional monopoles antennas in LOS, NLOS and NLOS-Specular settings. Empirical stochastic channel models for 5 and 10 MHz bandwidths were developed. The Weibull distribution parameter ($\beta \approx 1.5$) and Nakagami-$m$ factor ($m \approx 0.7$) were quantified for the modeling the observed severe fading on the airport surface area.
Measurement and modeling results in the 5 GHz band for a 50 MHz channel bandwidth were provided in [47]. The authors computed results from measurements at Miami International airport using omni directional antennas for mobile and directional horn antennas for non-mobile setups. Delay and fading amplitude characteristics were estimated around the airport surface area. The authors noticed the rich but different nature of scattering (non-isotropic distribution of scatterers) at the airport compared to the typical terrestrial cellular (urban and suburban) channel. Transmissions from the air traffic control tower showed greater channel dispersion than transmission from lower-antenna-height airport field sites.

Although some measurements have been done at higher mmWave frequencies, not as much has appeared for the 90 GHz band, a portion of which has been allocated for indoor applications. There is ample research on LOS and NLOS scenarios, but little (if any) research on LOS to NLOS transition channel characterization, where channel stationarity distance, AoA, DS, and AoD are crucial parameters for future mmWave wireless channel characterization. Hence in this dissertation, we provide measurement results for this 90 GHz band, using a 500 MHz bandwidth signal. Our environments are both indoor and outdoor settings (including aviation environments), both LOS and NLOS. In particular, we study characteristics of LOS-to-NLOS transitions, where channel characteristics can change abruptly, making link reliability challenging. We provide a brief summary of several key articles in Table 2.1.
Table 2.1 A brief summary of related channel results from the literature for various frequency bands, environments, types of analysis, and antennas.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Frequency Band [GHz]</th>
<th>Environment</th>
<th>Analysis Type</th>
<th>Antenna Type</th>
<th>Presented Results</th>
</tr>
</thead>
<tbody>
<tr>
<td>[17]</td>
<td>28, 73</td>
<td>Indoor office, LOS and NLOS</td>
<td>M</td>
<td>Horn</td>
<td>PL, RMS DS</td>
</tr>
<tr>
<td>[34]</td>
<td>28, 38, 60, 73</td>
<td>Outdoor, LOS and NLOS</td>
<td>M</td>
<td>Horn</td>
<td>PL, # MPCs, RMS DS, outage probabilities, statistical spatial channel model statistics</td>
</tr>
<tr>
<td>[48]</td>
<td>73</td>
<td>-</td>
<td>M</td>
<td>Horn</td>
<td>XPD, penetration loss</td>
</tr>
<tr>
<td>[35]</td>
<td>28, 73</td>
<td>Urban micro cellular Indoor, LOS, NLOS</td>
<td>M</td>
<td>Horn</td>
<td>PL</td>
</tr>
<tr>
<td>[30]</td>
<td>28, 40</td>
<td>Outdoor, LOS, NLOS</td>
<td>M</td>
<td>Horn</td>
<td>Penetration loss, reflectivity, AoA, PL</td>
</tr>
<tr>
<td>[31]</td>
<td>28</td>
<td>Indoor and outdoor</td>
<td>M</td>
<td>Horn</td>
<td>Reflection and penetration</td>
</tr>
<tr>
<td>[61]</td>
<td>28, 60, 73</td>
<td>Courtyard, in-vehicle, LOS, NLOS</td>
<td>M</td>
<td>Horn</td>
<td>PL models using correction factors, beam combining PL models</td>
</tr>
<tr>
<td>Reference</td>
<td>Frequency (GHz)</td>
<td>Environment</td>
<td>Configuration</td>
<td>Antenna</td>
<td>Measurement Parameters</td>
</tr>
<tr>
<td>-----------</td>
<td>----------------</td>
<td>-------------</td>
<td>---------------</td>
<td>---------</td>
<td>------------------------</td>
</tr>
<tr>
<td>[32]</td>
<td>28, 73, 91</td>
<td>Indoor (Lab)</td>
<td>M Horn</td>
<td></td>
<td>Attenuation</td>
</tr>
<tr>
<td>[29]</td>
<td>15</td>
<td>Indoor Corridor</td>
<td>S &amp; M Horn</td>
<td></td>
<td>Received power, PL, RMS DS</td>
</tr>
<tr>
<td>[37]</td>
<td>5, 31, 91</td>
<td>Indoor, LOS and NLOS</td>
<td>S &amp; M Omni (S &amp; M Horn) (5 &amp; 91 GHz)</td>
<td></td>
<td>PL, simulation using WI</td>
</tr>
<tr>
<td>[26]</td>
<td>28</td>
<td>Indoor, outdoor, LOS</td>
<td>S &amp; M Horn</td>
<td></td>
<td>PL, Ray based model, polarization effect, RT using WI, Ricean K-factor</td>
</tr>
<tr>
<td>[38]</td>
<td>41</td>
<td>Indoor corridor, NLOS</td>
<td>S &amp; M Horn</td>
<td></td>
<td>PL, RMS DS, MPCs, PDPS</td>
</tr>
<tr>
<td>[28]</td>
<td>73</td>
<td>Indoor office, LOS and NLOS</td>
<td>S &amp; M Horn</td>
<td></td>
<td>PL, DS, AS</td>
</tr>
<tr>
<td>[45]</td>
<td>5.2</td>
<td>Munich airport surface</td>
<td>M Omni</td>
<td></td>
<td>MPCs, DS, Doppler spread, Coherence BW &amp; time</td>
</tr>
<tr>
<td>[49]</td>
<td>28</td>
<td>Airport, terminal hall, indoor, outdoor, LOS, NLOS</td>
<td>M Horn</td>
<td></td>
<td>PL, reflection coefficient, penetration loss, MPCs, AoA, AoD</td>
</tr>
<tr>
<td>Ref</td>
<td>52</td>
<td>60</td>
<td>Airport, baggage &amp; gate area, University (hallways, outside building), LOS, NLOS</td>
<td>M</td>
<td>Horn</td>
</tr>
<tr>
<td>------</td>
<td>------</td>
<td>----</td>
<td>----------------------------------------------------------------------------------</td>
<td>---</td>
<td>------</td>
</tr>
<tr>
<td>Ref</td>
<td>50</td>
<td>5.1-5.15</td>
<td>Barajas airport, Madrid, Spain, Gate &amp; taxing area, LOS, NLOS, Mixed</td>
<td>M</td>
<td>Sector (BS)</td>
</tr>
<tr>
<td></td>
<td>51</td>
<td>5.2</td>
<td>Munich airport, surface</td>
<td>M &amp; S</td>
<td>-</td>
</tr>
</tbody>
</table>

- S Simulation
- M Measurement
CHAPTER 3

MILLIMETER WAVE CHANNEL SOUNDER CHARACTERIZATION AND CALIBRATION

3.1 SOUNDER STRUCTURE

A block diagram of the 90 GHz channel sounding system is shown in Figure 3.1. We begin our description by summarizing the signal paths and processing from transmission source to the receiver destination.

A wideband sounding signal is generated by a vector signal generator (VSG), Rohde & Schwarz (R&S) SMW200A, at an intermediate frequency (IF) of 3 GHz. Signal BW is selectable, but we used 500 MHz; the maximum value available with our VSG is 600 MHz. The local oscillator (LO) signal is also supplied by the VSG, set to a frequency of 11 GHz, and this signal gets multiplied up to 88 GHz with a multiplier (Quinstar, model number QMM-88011508). This LO signal is then passed through an 88 to 89 GHz band pass filter 1 (BPF1) (Quinstar, model number QFB-8901W0) and the output of the BPF1 is used as the other input to the upconverter (Quinstar, QMB-9389WS model) that mixes the IF and LO signals, yielding a 91 GHz center frequency radio frequency (RF) signal. This RF signal is passed through an isolator (Quinstar, QIF-W00000 model) and amplified by two amplifiers (Quinstar, QPN-93053030-P2 and QPW-90952720-HP202) after bandpass filtering (BPF2, Quinstar QFB-9095W0 model). The amplified output is connected to a 15 dBi gain, 32-degree half-power beamwidth (HPBW) W-band pyramidal
A horn antenna, model number SAR-1532-10-S2, manufactured by SAGE Millimeter, Inc. (now ERAVANT) [53]. An identical horn antenna is used at the Rx, and this captures and passes the signal to a R&S Rx mixer (RPG FS-Z110 model number) for down conversion with another LO to convert the signal to a 1330 MHz IF. This IF signal is the input to the signal and spectrum analyzer (SSA), R&S FSW43. The SSA down converts and samples, and the baseband samples are processed by the channel sounder receiver processor (correlating with the known transmit signal), and are transferred to a laptop for analysis. Multiple numerical values for system operational parameters are listed throughout the diagram in Figure 3.1, and a channel sounder photo is provided in Figure 3.2.

*Test means manufacturer specified data.
*Experiment means measurement data in laboratory.

Figure 3.1 A block diagram of 90 GHz channel sounding system.
The total loss/gain between Tx mixer and Tx antenna is shown in Table 3.1. According to the manufacturer’s data sheet, the saturated power of the amplifier 2 is 27 dBm. The maximum IF input power at the mixer, considering different possible values of overall gain between the Tx mixer and amplifier 2, can be found using $Gain = P_{out, RF} - P_{in, IF}$; here, $P_{out, RF}$ and $P_{in, IF}$ are the RF output power at the amplifier 2 and the IF input power at the Tx mixer, respectively.

Table 3.2 shows different values of the maximum IF input power at the upconverter considering overall gain between the Tx mixer and the amplifier 2. The range of IF input power at the mixer is -24.5 to -20 dBm considering maximum and minimum gain. In our measurement, we use a value of -26 dBm for the IF signal at the Tx mixer, yielding 25.5 dBm signal at the Tx antenna.
Table 3.1. Sounder upconverter assembly components Loss/Gain.

<table>
<thead>
<tr>
<th>Components</th>
<th>Loss/Gain [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>TxMixer (Conversion Loss, minimum)</td>
<td>6.3</td>
</tr>
<tr>
<td>TxMixer (Conversion Loss, maximum)</td>
<td>7.7</td>
</tr>
<tr>
<td>BPF2 (Insertion Loss)</td>
<td>3.6</td>
</tr>
<tr>
<td>Amplifier1 (Gain @ 90 GHz)</td>
<td>36.7</td>
</tr>
<tr>
<td>Amplifier1 (Gain @ 92.5 GHz)</td>
<td>37.4</td>
</tr>
<tr>
<td>Amplifier1 (Gain @ 91 GHz, using interpolation)</td>
<td>37</td>
</tr>
<tr>
<td>Isolator (Insertion Loss)</td>
<td>2</td>
</tr>
<tr>
<td>Amplifier2 (Gain @ 90 GHz)</td>
<td>26</td>
</tr>
<tr>
<td>Amplifier2 (Gain @ 92.5 GHz)</td>
<td>23.6</td>
</tr>
<tr>
<td>Amplifier2 (Gain @ 91 GHz, using interpolation)</td>
<td>25.2</td>
</tr>
<tr>
<td><strong>Overall Maximum Gain</strong></td>
<td>51.5</td>
</tr>
<tr>
<td><strong>Overall Minimum Gain</strong></td>
<td>47</td>
</tr>
<tr>
<td><strong>Overall Maximum Gain considering amplifiers</strong></td>
<td></td>
</tr>
<tr>
<td><strong>interpolated Gain</strong></td>
<td>50.3</td>
</tr>
<tr>
<td><strong>Overall Minimum Gain considering amplifiers</strong></td>
<td></td>
</tr>
<tr>
<td><strong>interpolated Gain</strong></td>
<td>48.9</td>
</tr>
</tbody>
</table>

Table 3.2. Maximum IF input for different RF assembly gain values.

<table>
<thead>
<tr>
<th>Gain</th>
<th>IF input at the Tx mixer [dBm]</th>
<th>Estimated output power at the Tx antenna [dBm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum</td>
<td>-24.5</td>
<td></td>
</tr>
<tr>
<td>Minimum</td>
<td>-20</td>
<td></td>
</tr>
<tr>
<td>Maximum considering interpolation</td>
<td>-23.3</td>
<td></td>
</tr>
<tr>
<td>Minimum considering interpolation</td>
<td>-21.9</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>27</td>
</tr>
</tbody>
</table>

3.2 CHANNEL SOUNDER SOFTWARE

Our channel sounding measurements use the R&S TS-5GCS software [52]. This channel sounding software generates a signal using one of several signal types (chirp, or pseudo-random code). We set the signal as chirp, and BW as 500 MHz. The receiver software commands a correlation of the known signal with the incoming received signal to
estimate MPCs. The software allows two channel sounding modes: untriggered or triggered. In triggered mode, a reference signal (10 MHz clock) from the VSG to the SSA synchronizes the instruments, and a trigger signal initiates the measurements. This mode removes any frequency and sampling clock offsets between the Tx and the Rx, and allows absolute delay measurements, but requires cable connections. Trigger and reference signals are not provided in untriggered mode, which is used for most measurements, where cable connection is inconvenient or impossible. In this mode, the Rx is triggered by the initial arrival of the sounding signal itself, and the measurement results are relative to the first-arriving MPC. The primary measurement result is a relative PDP, starting with the first relevant path.

A chirp signal is known as a frequency modulated continuous wave signal. A linear frequency increase or decrease within a $B$ Hz frequency range in $T$ sec characterizes this signal. Sweep rate and dispersion factor are represented as $B/T$ Hz/s and $BT$, respectively. This signal is expressed as follows [22]:

$$S(t) = A\cos(\omega_c t + \chi t^2), \quad -\frac{T}{2} \leq t \leq \frac{T}{2} \quad (3.1)$$

Here, $\omega_c$ = (starting) carrier frequency in radian/s, and $\chi = \pi \frac{B}{T}$.

Our transmitted chirp waveform duration is 0.13107 ms, and the length is 65,536 samples [52], i.e., the signal sweeps from the minimum to maximum frequency in 0.13107 ms. In untriggered mode, we transmit the signal a selectable number of times (default is 152 channel snapshots). After correlation, in each snapshot, we have a 1001-sample CIR with delay range 2 $\mu$ sec. Using this channel sounding software, we obtain the CIR, from
which we can estimate different parameters, e.g., RMS DS, mean excess delay, Doppler, etc.

We next briefly describe the sounding software operation. When we start the channel sounding software, the main window appears as in Figure 3.3. First, we must choose the sounding waveform type by selecting “Settings > Waveform Settings;” we select “Chirp_16_bit.mat”, shown in Figure 3.4. Next, we configure the Tx and the Rx in the main application window by selecting "Settings > Measurement Settings > General", illustrated in Figure 3.5: Tx clock rate 500 MHz, Tx power level 25.5 dBm, Tx and Rx antenna type horn, Tx and Rx antenna gain 15 dBi, Rx Frequency 91000 MHz, and Rx Sampling Rate 500 MHz.

Figure 3.3 Channel sounder software main window [52].
Before performing any measurement, we conducted a calibration by connecting a reference cable and trigger cable between the VSG and the SSA and selecting "Measurement > Start Trigger Calibration" from the channel sounding software. After several minutes, for any measurement (indoor or outdoor), we disconnected both cables and started our untriggered measurement by selecting "Measurement > Start Untriggered Measurement". Once measurement starts, our SSA immediately records the samples as complex IQ data and passes this data to the channel sounding software, which then computes the number of available CIRs based on the recorded number of samples, the sampling rate and the previously set up waveform. From each CIR, the software creates a channel snapshot with a relative time delay axis, and we can observe a PDP diagram,
Figure 3.5 Tx and Rx settings in the channel sounder software [52].

Figure 3.6 An example PDP displayed in the main window [52].
We can see the intensity of the Rx signal as a function of time delay and delay Doppler by selecting "View > Power/Delay/Time Profile" and "View > Power/Delay/Doppler Profile", respectively. We can also observe the results in the Frequency domain by selecting “View > Frequency Domain”. For all measurements, we export our measured data by selecting “Export” in the main bar and storing the required file in the laptop for further processing.

3.3 SPECIFIC IF SIGNAL SELECTION AT TX MIXER

In this section we briefly describe the procedure and results for selecting the signal type. Theoretically, each of the signal types should be equivalent in terms of performance (e.g., peak-to average power ratio, spectral flatness), but practical implementations yield some differences. Here we address only spectral flatness. Based upon cable losses in our available RF cables, we decided to use a value of +5 dBm at the multiplier and -26 dBm maximum IF input at the mixer (Figure 3.1). This provides us a maximum estimated output power at the Tx antenna of 25.5 dBm considering maximum gain between the Tx mixer and the amplifier 2. To represent a near-ideal channel, the Tx and the Rx were connected via two waveguide attenuators to get 80 dB attenuation: a 30 dB attenuator (Quantum Microwave, model number QMC10-ATT30, range 75-110 GHz, power 0.3 W) and a 50 dB attenuator (ERA V ANT, model number STA-50-10-F2, range 75-110 GHz, power 0.5 W, 50 dB attenuation @ 92.5 GHz). The IF cable loss from the Rx down converter to the SSA is 6 dB. We selected the IF signal based upon the spectral flatness of the different types of IF signals: chirp 16 bit, chirp 16 bit 10x, FZC 16 bit 10x and FZC 16 bit 100x (here the “bit numbers” denote the length of the shift registers used to generate the pseudo-random sequences, e.g., $2^{16}$ for 16 bit).
Table 3.3 shows the measured spectral ripple for the different waveforms. An example SSA measured spectrum using the 16 bit chirp signal as an IF input is illustrated in Figure 3.7. All outputs from the other IF signals are shown in Appendix A. We observed a spurious output (~tone) at 90.62 GHz, but this does not affect our CIR results. We selected the chirp 16 bit signal as the IF input signal for measurements because of its spectral flatness, i.e., a ripple of only 0.35 dB.

Table 3.3 The ripple observed in different waveforms.

<table>
<thead>
<tr>
<th>Waveforms</th>
<th>Ripple [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Chirp 16-bit (waveform length: 65535)</td>
<td>0.35</td>
</tr>
<tr>
<td>Chirp 16-bit 10x (waveform length: 655350)</td>
<td>0.5</td>
</tr>
<tr>
<td>FZC 16-bit 10x (waveform length: 655350)</td>
<td>1.22</td>
</tr>
<tr>
<td>FZC 16-bit 100x (waveform length: 6553500)</td>
<td>6</td>
</tr>
</tbody>
</table>

Figure 3.7 Ripple observed in the SSA for 16 bit chirp signal spectrum.
3.4 RECEIVED POWER CALIBRATION USING WAVEGUIDE ATTENUATOR

In this section, we show a relation between measured and estimated received power at the receiver. The estimated received power was computed by using the maximum gain from Table 3.1. Two waveguide attenuators were used between the Tx and the Rx, a 30 dB and a 80 dB attenuator. We also examine the relation between IF input power at the mixer and output power at the receiver. The R&S TS-5GCS software is used to capture CIRs recorded by the SSA. Received power using these CIRs was also compared with the estimated received power.

We first used a 30 dB attenuator between the Tx and the Rx, and measured the output power. By changing the IF input power at the mixer from -60 dBm to -30 dBm, we observed the output RF power; this is shown in Figure 3.8. It is observed that the output power at the SSA does not increase linearly for the IF input power above -41 dBm. We also observed a message in the SSA, “RF OVLD”, during our measurement. This error message is displayed if there is any overload in the input mixer [58]. To solve this overload issue, we used an 80 dB attenuator instead of a 30 dB attenuator and examined the received power.

Using an 80 dB attenuator between the Tx and the Rx, the IF input at the mixer was changed from -46 dBm to -26 dBm. The IF input at the mixer and RF output power showed a linear relationship—observed in Figure 3.9. The curve is steep at low IF input levels, i.e., below -31 dBm.
Figure 3.8 Relation between input IF signal at mixer and output RF signal at the SSA using 30 dB attenuator.

Figure 3.9 Relation between input IF signal at mixer and output RF signal at the SSA using 80 dB attenuator.
The relation between the estimated and the measured RF power at the Rx was determined by changing the IF input at the Tx mixer from -46 dBm to -26 dBm in steps of 5 dB, while using an 80 dB attenuator between the Tx and the Rx. The relationship is almost linear, illustrated in Figure 3.10. We observe a significant difference between the estimated and the measured received power (approximately 5 dB), when the IF input power at the Tx mixer is maximum, i.e., -26 dBm. The amplifier 2 of the Tx may be close to the output saturation level at this input power. Moreover, we find a relation among estimated, SSA displayed, and channel sounder software determined RF received power by connecting 80 dB, 90 dB, and 100 dB of attenuation between the Tx and the Rx and by using -26 dBm 500 MHz chirp signal at the Tx mixer.

Figure 3.10 Relation between measured and estimated RF power at the SSA.
We show in Figure 3.11 that the difference between estimated received power and measured received power by using SSA is approximately 5 dB. We considered overall 51.5 dB gain from IF input at the Tx mixer to RF output at the amplifier 2. The overall gain might be smaller by 5 dB due to the saturation of output power level in amplifier 2. The difference between the two measured powers, measured power using the SSA and measured power by calculating captured CIRs using the channel sounding software, are very close for all attenuators, approximately 1 dB difference.

Figure 3.11 Relation among the estimated, SSA displayed and the sounder software calculated RF power using different attenuators.
3.5 MEASUREMENT IN TRIGGERED AND UNTRIGGERED MODE

We confirmed in our laboratory tests that when the Tx and the Rx were run “back to back” in triggered mode for calibration, after disconnecting synchronization cables, operation in untriggered mode yielded essentially identical channel measurement results, for up to 24 hours. That is, after calibration, the Tx and the Rx maintain good synchronization for up to one day.

To do this measurement, we used an 80 dB attenuator between the Tx and the Rx. A 500 MHz – 26 dBm chirp signal was used as the IF input at the Tx mixer. We connected two cables between the VSG and SSA for synchronization: the reference cable and trigger cable. We captured CIR samples when both of these cables were connected. Next, we disconnected those two cables and captured CIR samples over time: immediately after cable removal (within 1 min), 30 min, 1 hour and 23 hours after cable removal.

Figure 3.12 shows PDPs for the triggered connection and the untriggered data. It can be observed in Figure 3.12 that the untriggered PDP after 30 minutes closely follows the triggered PDP. It is also noticed that the PDP after 1 hour does change, and similarly after 23 hours. Interestingly, the PDP immediately after cable removal exhibits a different pattern from the other samples.
To quantify the significance of the change over time, we computed the DS and range between maximum and minimum DS for several cases: (i) using all PDP samples, (ii) using the internal sounding software detected threshold (determined by software detected noise threshold), (iii) PDPs resulting when removing all samples that are 50 dB, 30 dB and 25 dB samples below the peak, (iv) the delay window (the length of the middle portion of the PDP containing 90% of the total power, W90) and (v) the delay interval (the time difference between $t_1$ when the amplitude of PDP first exceeds a threshold, and $t_2$ when it falls below that threshold for the last time, for a 25 dB threshold, $I_{25}$). These DS measures and threshold values are commonly used among researchers. Results are shown in Table 3.4 and Table 3.5.
It is observed in Table 3.4 that RMS DS changes immediately when the trigger cable is removed, e.g., approximately 2% of the triggered mode for 50 dB threshold. After 30 minutes of untriggered mode, the difference between triggered and untriggered mode decreases, changing approximately 1% of the triggered mode for 50 dB threshold.

Table 3.4 RMS DS both for triggered (T) and untriggered modes (UT).

<table>
<thead>
<tr>
<th></th>
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<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>T</td>
<td>13.37</td>
<td>2.1600</td>
<td>2.12</td>
<td>1.73</td>
<td>1.52</td>
<td>0.78</td>
<td>1.53</td>
</tr>
<tr>
<td>UT 1 min</td>
<td>13.44</td>
<td>2.1675</td>
<td>2.16</td>
<td>1.79</td>
<td>1.40</td>
<td>1.06</td>
<td>1.41</td>
</tr>
<tr>
<td>UT 30 min</td>
<td>13.38</td>
<td>2.1707</td>
<td>2.10</td>
<td>1.68</td>
<td>1.58</td>
<td>0.68</td>
<td>1.59</td>
</tr>
<tr>
<td>UT 60 min</td>
<td>13.32</td>
<td>2.1640</td>
<td>2.09</td>
<td>1.65</td>
<td>1.53</td>
<td>0.48</td>
<td>1.53</td>
</tr>
<tr>
<td>UT 1380 min</td>
<td>13.15</td>
<td>2.1876</td>
<td>2.10</td>
<td>1.70</td>
<td>1.62</td>
<td>0.45</td>
<td>1.62</td>
</tr>
</tbody>
</table>
If we consider triggered RMS DS as 13 ns, then 1% of this is 0.13 ns, i.e., untriggered RMS DS can be 13.13 ns. Our delay resolution is 2 ns. The number of symbols for 13 ns and 13.13 ns RMS DS are 6.5 and 6.6. This is likely indistinguishable for any communications receiver in practice. As RMS DS does not change significantly up to 23 hours of untriggered mode, untriggered mode can be used for large link distance measurement.

Table 3.5 RMS DS range.

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<tr>
<td>Maximum</td>
<td>13.4428</td>
<td>2.1707</td>
<td>2.1603</td>
<td>1.796</td>
<td>1.5877</td>
<td>1.06</td>
<td>1.62</td>
</tr>
<tr>
<td>Minimum</td>
<td>13.3227</td>
<td>2.16</td>
<td>2.0961</td>
<td>1.6585</td>
<td>1.4059</td>
<td>0.45</td>
<td>1.41</td>
</tr>
<tr>
<td>Range</td>
<td>0.1201</td>
<td>0.0107</td>
<td>0.0642</td>
<td>0.1375</td>
<td>0.1818</td>
<td>0.61</td>
<td>0.21</td>
</tr>
</tbody>
</table>
3.6 CALIBRATED SYSTEM BACK TO BACK RESPONSE

Our 90 GHz channel sounder RF subsystems consist of a multiplier, BPFs, Tx and Rx mixers, an isolator and amplifiers. For calibration, we used an 80 dB attenuator between the Tx and the Rx. In our indoor and outdoor measurements, we used horn antennas in the Tx and the Rx. The sounder’s system response is also included in this wireless measured response. We removed this system response from our measured response to evaluate wireless channel. In this section, we describe our calibration process.

We performed calibration by removing the system response. We used a -26 dBm IF input 500 MHz chirp signal at the Tx mixer and connected the Tx and the Rx with the 80 dB broadband and low-loss waveguide attenuator. We captured the received signal using channel sounding software in triggered mode. We determined the calibrated system back to back response and observed a near Dirac delta, shown in Figure 3.13.

![Calibrated system back to back response](image)

Figure 3.13 Calibrated system back to back PDP.
We also determined the system transfer function (\(|H_{sys}|^2\)) using captured CIRs by channel sounding software in triggered mode, illustrated in Figure 3.14. It is noticed that the system frequency response is not perfectly flat. The response \(H_{sys}(f)\) is stored and saved, and used after all measurement results: we divide the measured response with the system response to get actual wireless channel response using (3.2)

\[
Y(f) = H_{sys}(f) H_c(f) \quad (3.2)
\]

Here, \(Y(f)\) = Measured response, \(H_{sys}(f)\) = system response, and \(H_c(f)\) = wireless channel response.

![System Transfer Function's magnitude squared](image-url)

Figure 3.14 Triggered system Transfer Function’s magnitude squared.
During our measurements, we follow the following steps before taking channel data:

- Operate the sounder connected back-to-back for at least 30 minutes for synchronization, connecting reference and trigger cables between the Tx and the Rx;
- Disconnect the cables after 30 minutes of synchronization;
- Conduct measurement as soon as possible (e.g., 15 minutes) after disconnection.
CHAPTER 4
MILLIMETER WAVE CHANNEL MEASUREMENT ENVIRONMENTS,
MEASUREMENT PROCEDURES AND SIMULATIONS

4.1 MEASUREMENT ENVIRONMENTS

We conducted measurements in both indoor and outdoor settings for both LOS and NLOS scenarios. For all measurements, the Tx was fixed in position, and we moved the Rx antenna (in the azimuth plane) to capture maximum received power. Our aim was to analyze the influence of surrounding objects, e.g., glass, metal, drywall, concrete wall, brick wall etc., on the channel transition characteristics and PL. We also wanted to quantify important wireless channel parameters, e.g., RMS DS, SPCC, large scale shadowing parameters. Two different environments were selected for measurements: non-aviation (hallways and streets) and aviation (an airport maintenance hangar and airport baggage areas). The measurements environments are explained in detail in the following subsections.

4.1.1 NON-AVIATION ENVIRONMENT

We conducted measurements moving from LOS to NLOS zones in the Swearingen (SWGN) Engineering Center building at the University of South Carolina (U of SC): 1st floor A wing (1A), 2nd floor D wing (2D), 3rd floor D wing (3D), and the outdoor areas near SWGN Northwest (NW) and 300 Main Street. We also measured in LOS conditions along
the 3D corridor, 1A lobby, SWGN NW outdoor and 300 Main Street outdoor; see Figures 4.1 and 4.2. The 1A lobby has a large glass wall on one side and an open area in the middle with several tables, chairs, and ceramic floor tile. The 2D hallway is a space with plasterboard walls, tile floor, acoustic tile ceilings, and multiple wooden doors with metal frames. The 3D hallway is similar to 2D, except there are also glass windows looking into adjoining rooms. These windows have metal mesh inside the glass and surrounding metal frames. In 300 Main Street, we observed multiple potential reflectors: concrete and brick walls, the car parking area surface and multiple metal sign poles. The Rx path away from the Tx here is slightly inclined. The SWGN NW area is similar to that of 300 Main Street, except there are stairs, stair rails, a bicycle rack, small pylons, light poles and only concrete walls.

![Figure 4.1 Swearingen indoor corridor measurement location photos, clockwise from upper left: 3D hallway, 2D hallway; lobby 1A (near NLOS region), and lobby 1A (LOS zone).]
4.1.2 AVIATION ENVIRONMENTS

Measurements were conducted in two airports: Jim Hamilton L. B. Owens Airport (KCUB) in downtown Columbia, SC, and Columbia Metropolitan Airport (CAE) in West Columbia, SC. We performed measurements in the airport maintenance hangar in KCUB, and in the airport baggage areas in CAE. The following subsections provide a detailed description of these environments.

4.1.2.1 AIRPORT MAINTENANCE HANGAR

The KCUB airport is a small municipal airport. Measurements were done in the indoor environments at CUB’s nearby aircraft maintenance hangar. The maintenance hangar has interior thermal insulation (foil-backed) and a concrete floor. It is a metal
building with an approximate 30 by 40 m footprint with 10 m height. In the hangar during our measurements, which were made during normal work hours, there were six or seven small-medium aircraft, large metal tool chests, tables of various sizes, metal objects, plastic cabinets, and a few other objects. During the measurement time, a few tool cabinets and tables were moved. The airport maintenance team personnel also moved about occasionally.

Figure 4.3 illustrates a CUB hangar measurement environment photo showing four radial paths: LOS (path 1), and mixed (paths 2, 3 and 4). As there were moving local obstacles, i.e., aircraft, cabinets, it was difficult to ensure long continuous LOS or NLOS conditions. For this reason, we describe our results as a mixture of LOS and NLOS, i.e., a mixed environment. This is quite practical for the maintenance hangar where local obstacles will move during the day, in contrast to other indoor environments, e.g., office buildings. We conducted our measurements for two days. On the first day, when we performed LOS and mixed setting measurements, the large doors on two of the building’s four sides were open for the entire time of our approximately 4-hour measurement period.
On the second day, we performed LOS to NLOS transition measurements, moving the Rx behind the aircraft. A plan view diagram and photos appear in Figures 4.4 and 4.5, respectively. We note some potential reflecting objects, e.g., engines, cabinets, walls. During our measurements, we observed that one of the large front doors was opened for entire measurement time. The other door was closed when we performed our measurement at Rx path 1 and opened after few minutes when we started our Rx path 2 measurement.

Figure 4.4 LOS to NLOS transition measurement at KCUB: plan view diagram.

Figure 4.5 LOS to NLOS transition measurement (KCUB): Rx path 1 at top and Rx path 2 at bottom.
4.1.2.2 AIRPORT BAGGAGE AREA

Our final indoor setting was the Columbia Metropolitan Airport (CAE) in West Columbia, SC. This is a commercial airport, and the principal airport for the Midlands region of South Carolina. We performed our measurement in the CAE baggage area, which is in the level 1 of the CAE terminal building. In this area, there are three baggage claim carousels, a baggage service area, restrooms, glass gates with surrounding metal frames, multiple metal pillars and sign frames, and a ramp. A plan view diagram of the CAE measurement environment is shown in Figure 4.6.

In Figure 4.6, we show our three Tx positions: Tx position 1 (Tx1), Tx position 2 (Tx2) and Tx position 3 (Tx3). We conducted LOS, NLOS, mixed and LOS to NLOS transition measurements, gathering PDPs for each Rx position as previously described.

In Fig. 4.6 for Tx1, we show a long LOS path, about 75 m between the Tx1 location and the final Rx position. In this LOS path, from the Tx1 perspective, there are twelve metal pillars along the left side, and three baggage carousels and twelve metal pillars on the right side. There is a ramp sloping upward starting from approximately 50 m away from the Tx1 position. There are metal rails on both sides along the ramp. Transmitter locations Tx2 and Tx3 also allow illumination of multiple reflectors, e.g., carousels, multiple pillars, metal sign holders. Table 4.1 summarizes the measurement settings and some of their characteristics.
Figure 4.6 Plan view diagram of measurement locations at CAE baggage area, showing three Tx positions (Tx1, Tx2, and Tx3) and multiple Rx movement paths (LOS, NLOS, mixed, LOS to NLOS transition).
Table 4.1 Locations and measurement setting.

<table>
<thead>
<tr>
<th>Measurement Environments</th>
<th>Measurement Locations</th>
<th>Measurement Settings</th>
<th>Probable reflectors</th>
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<tbody>
<tr>
<td>Indoor</td>
<td>Non-aviation</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>SWGN 1&lt;sup&gt;st&lt;/sup&gt;</td>
<td>LOS, LOS to NLOS</td>
<td>Metal doors and</td>
</tr>
<tr>
<td></td>
<td>floor lobby</td>
<td>transition</td>
<td>windows’ frames,</td>
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<tr>
<td></td>
<td></td>
<td></td>
<td>doors’ locks, glass</td>
</tr>
<tr>
<td></td>
<td>SWGN 2&lt;sup&gt;nd&lt;/sup&gt;</td>
<td>LOS to NLOS</td>
<td>windows, floor tiles,</td>
</tr>
<tr>
<td></td>
<td>floor hallway</td>
<td>transition</td>
<td>etc.</td>
</tr>
<tr>
<td></td>
<td>SWGN 3&lt;sup&gt;rd&lt;/sup&gt;</td>
<td>LOS, LOS to NLOS</td>
<td></td>
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<tr>
<td></td>
<td>floor hallway</td>
<td>transition</td>
<td></td>
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<tr>
<td>Aviation</td>
<td>CUB maintenance</td>
<td>LOS, mixed, LOS to</td>
<td>Wall, floor, aircrafts,</td>
</tr>
<tr>
<td></td>
<td>hangar</td>
<td>NLOS transition</td>
<td>cabinets, cylinder,</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>engines, tools, etc.</td>
</tr>
<tr>
<td></td>
<td>CAE baggage area</td>
<td>LOS, mixed, LOS to</td>
<td>Pillars, frames of sign</td>
</tr>
<tr>
<td></td>
<td></td>
<td>NLOS transition</td>
<td>holders, carousels,</td>
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<td></td>
<td></td>
<td></td>
<td>etc.</td>
</tr>
<tr>
<td>Outdoor</td>
<td>Non-Aviation</td>
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<tr>
<td></td>
<td>SWGN NW</td>
<td>LOS, LOS to NLOS</td>
<td>Concrete wall, pylons,</td>
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<tr>
<td></td>
<td></td>
<td>transition</td>
<td>bike rack, sign poles,</td>
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<tr>
<td></td>
<td></td>
<td></td>
<td>stair rails, etc.</td>
</tr>
<tr>
<td></td>
<td>300 Main Street</td>
<td>LOS, LOS to NLOS</td>
<td>Concrete and brick</td>
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<tr>
<td></td>
<td></td>
<td>transition</td>
<td>wall, sign poles, cars,</td>
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<td></td>
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<td>etc.</td>
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</table>
4.2 MEASUREMENT PROCEDURES

The Tx and the Rx were placed on wheeled carts, and both Tx and Rx antennas were mounted at heights of 137 cm. For both LOS and NLOS measurements, the Tx was stationary. The Rx antenna was mounted on a manual adjustment positioner that allowed lateral movement and azimuth rotation. Both antennas used vertical polarization. For LOS measurements, a boresight alignment was made by adjusting the azimuth angle, while the elevation angle was fixed (0°) for all measurement locations. For LOS to NLOS transition measurements, we started recording measurements in the LOS area before passing the corner of the transition region, then proceeded to the obstructed NLOS region and captured the received signal at the azimuth angle that yielded the largest received power. At each measurement location, three or four measurements at one wavelength apart were recorded around the best boresight alignment position to average over small scale fading due to MPCs caused by reflection, diffraction, and scattering. We define our measured paths as LOS when the Tx and the Rx are in optical LOS. If the 1st Fresnel zone is obstructed, then we define this path as partially obstructed path, otherwise NLOS.

For indoor and outdoor LOS measurements, we moved the Rx away from the Tx in steps of 1 or 2 m at each measurement location, as shown in Figures 4.1 and 4.2. We used a distance increment of 2 cm in the 3D, 2D hallways, SWGN NW outdoor and 300 Main Street, and an increment of 50 cm in the 1A lobby for LOS to NLOS transition measurements. The Rx antenna was 75 cm and 1 m away from the obstructing wall in all indoor and outdoor settings, respectively. We did the same small-scale fading averaging at each location. For the SWGN NW measurement, as the Tx was above the top stairs; there is an extra 73 cm difference between the Rx at ground level and the Tx at the top of the
Figure 4.7 shows a general NLOS measurement scenario for both indoor and outdoor settings. As noted, in the NLOS regions, we captured the received signal at the azimuth angle that yielded the largest received power, approximating the expected operational approach of a working mmWave system. The direct Tx-to-Rx signal is obstructed by a wall. The antenna reference position (0 degrees) for 3D and 2D hallways, 1A lobby, SWGN NW, and 300 Main Street outdoor are shown in Figure 4.7. The reference Rx antenna position is a straight-line connecting Tx and Rx. If the antenna moves clockwise from this line, the angle at maximum received power $\theta_{\text{max}}$ is positive, otherwise it is negative, e.g., the antenna aims at the opposite wall of the NLOS path and toward the open area in D wing (2D and 3D hallway) and 1A lobby, respectively, when the azimuth angle is positive.

Figure 4.7 Typical NLOS scenarios, left to right: 2D and 3D hallway floor plans (indoor), 1st Floor lobby (indoor), SWGN NW and 300 Main Street (outdoor).
In the SWGN NW outdoor and 300 Main Street settings for LOS to NLOS transition measurements, we averaged over small-scale fading by taking received power at different antenna rotation positions (reference, clockwise, counterclockwise, i.e., $\theta_{\text{max}}$, $\theta_{\text{max}} - \theta_r$ and $\theta_{\text{max}} + \theta_r$), without changing the antenna position horizontally (right or left). We calculated the rotation angle $\theta_r$ using,

$$\frac{\text{arc length}}{\text{radius}} = \theta_r.$$  \hspace{1cm} (4.1)

Here, the radius of the antenna movement is the distance between the antenna's phase center and the rotation point, approximately equal to 8 cm in our setup, illustrated in Figure 4.8. One wavelength movement along the arc for 91 GHz RF signal is equivalent to a 2.15-degree rotation. For some locations, where we observed an equal maximum received power at more than one azimuth angle, more small-scale averaging resolved this ambiguity. We note that the rotation of 2.15 degrees is much smaller than the antenna’s main lobe beamwidth (approximately 30 degrees).

Figure 4.8 Distance between antenna’s phase center and rotation point.
When we conducted our measurements at KCUB, we moved the Rx away from the Tx in steps of 4 m for path 1 (LOS) and 3 m for paths 2 (mixed), 3 (mixed) and 4 (mixed). The Tx and the Rx were boresight aligned for all LOS and mixed measurements. We observed maximum received power by rotating the Rx antenna from reference position in the azimuth plane in NLOS (path 4) and a few mixed measurements (paths 2 and 3). Again, the antenna reference position is a straight-line connecting the Tx and the Rx. The azimuth angle is positive when the Rx antenna moves counterclockwise (looking from the Rx towards the Tx), otherwise negative. Figure 4.9 shows LOS, and mixed environment measurement photos.

Figure 4.9 LOS, and mixed environments measurement photos from the KCUB maintenance hangar, clockwise from top left: LOS path 1, mixed path 2, mixed path 3, and mixed path 4.

We also conducted LOS to NLOS transition measurements within the CUB maintenance hangar; photos are shown in Figures 4.10 and 4.11. We performed these transition measurements in two paths, Rx path 1 and Rx path 2. In both paths, at first, we conducted reference measurements in the LOS zone, then moved the Rx behind the aircraft.
for conducting LOS to NLOS transition measurements. We took measurements at intervals of 50 cm up to a maximum distance of 9 m from the reference.

In receiver path 2, we observed two LOS zones: one between 0 to 1.5 m from the reference and other between 8.5 to 9 m from the reference, but the latter area was beyond HPBW of the Tx antenna, illustrated in Figure 4.11.

![Figure 4.10 LOS to NLOS transition measurement at CUB maintenance hangar, Rx path 1.](image)

![Figure 4.11 LOS zone beyond HPBW of the Tx antenna at CUB maintenance hangar, Rx path 2.](image)
Our last indoor area at which we performed LOS, NLOS, obstructed LOS, and LOS to NLOS transition measurements was the CAE baggage area. Here, we conducted LOS measurements at intervals of 5 m up to 50 m from the reference for Tx position 1 (Tx1). We also performed a LOS to NLOS transition measurement at 75 m from the Tx on the top of the ramp, by moving the Rx behind the wall to yield NLOS conditions. The Rx antenna was rotated in azimuth plane to capture the maximum received power, illustrated in Figure 4.12.

Figure 4.12 Measurement in CAE baggage area: at top LOS and at bottom LOS to NLOS transition.
When the Tx was in Tx1, the Rx was also moved behind each carousel from reference position and measurements were taken for three different conditions: LOS, NLOS and partially obstructed. We placed the receiver behind a metal pillar and a metal sign holder for capturing NLOS and partially obstructed CIRs, respectively. The reference positions were considered as 15 m, 30 m and 45 m from the Tx1, near carousels 3, 2 and 1, respectively. We performed our measurements up to maximum 13 m from these reference positions.

Figure 4.13 LOS, partially obstructed and NLOS measurement at CAE baggage area for Tx1: clockwise from upper left at baggage area 1, 2 and 3, respectively.
The Tx was placed at baggage area 2 (Tx2), on the opposite side of our CAE LOS measurement path. We conducted a LOS (boresight) measurement first by placing the Tx and the Rx at two sides of the carousels, shown in Figure 4.14. This was our reference Rx position. Then the Rx was moved to the right behind the carousel and toward the left behind the pillar from this reference LOS position to get NLOS and partially obstructed CIRs, respectively.

![Figure 4.14 Photo showing LOS, NLOS and partially obstructed measurement areas near CAE baggage area 2 for Tx2.](image)

The transmitter positions 3 and 2 are actually the same location, except the transmitter’s antenna was rotated 45° clockwise from the reference antenna position, aiming at one of the pillars, in Tx position 3 (Tx3). We mentioned that the reference antenna position is a straight line between the Tx and the Rx, when both the Tx and the Rx antennas are boresight aligned. At Tx3, we conducted measurements at similar values of distance as those used for Tx2: LOS (beyond Tx antenna HPBW), NLOS (behind carousel), partially obstructed (behind pillar and beyond Tx antenna HPBW); these are illustrated in Figure
4.15. We also performed LOS and partially obstructed measurements by moving the Rx behind the carousel in baggage area 1.

![Figure 4.15 LOS, NLOS and partially obstructed measurement at CAE baggage area for Tx3: at top near baggage area 2 and at bottom behind baggage area 1.](image)

4.3 RAY TRACING SIMULATIONS

Both indoor and outdoor non-aviation environments were modeled by WI for LOS settings. We selected features first for our indoor and outdoor simulations, e.g., city, floor plan, objects, and assigned materials, e.g., dry wall, wood, glass, concrete, brick, etc., according to [22], for doors, windows, ceiling, floors, and walls. To compare with our measurements, we used a 91 GHz center frequency chirp signal as a waveform and fixed the Tx and the Rx horn antennas with 15 dBi gain. The Tx and the Rx were positioned at
the same distances and heights as used in our measurements. The software requires us to select the number of reflections, transmissions, and diffractions, which we set as 6, 0, and 1, respectively. For indoor simulations, we imported floor plans of U of SC, Columbia, SC, using the DXF file format. Figure 4.16 shows an example illustration from WI of the setting and a set of propagation paths for 3rd floor hallway.

![Wave propagation paths for SWGN 3rd Floor indoor, from WI.](image)

In addition, we created walls, stairs, cars etc., as objects for outdoor simulation: SWGN NW and 300 Main Street. For our actual simulations, when estimating PL, we averaged over small scale fading for all indoor and outdoor simulations, just as in our measurements. Propagation paths for LOS settings of SWGN NW and 300 Main Street are shown in Figures 4.17 and 4.18, respectively. We did not include all potential objects, e.g., we excluded sign poles in 300 Main Street and small pylons in SWGN NW.
Figure 4.17 Wave propagation paths for 300 Main Street outdoor, from WI.

Figure 4.18 Wave propagation paths for SWGN NW outdoor, from WI.
CHAPTER 5
MEASUREMENT AND SIMULATION RESULTS FOR INDOOR CORRIDORS AND OUTDOOR STREET ENVIRONMENTS

5.1 MILLIMETER WAVE CHANNEL CHARACTERIZATION AND PL MODEL

The presence of different building materials indoors, e.g., glass, drywall, metal, etc., and outdoors, e.g., concrete wall, brick wall, metal stair rail, bike rack, etc., have varying effects on mmWave signals, producing both attenuation and MPCs. Millimeter wave signals experience less diffraction than at lower frequencies. It is worth noting that the objects traditionally represented as scattering objects at lower frequencies behave as reflectors at mmWave frequencies, which often yield measurable MPCs. The RMS DS in LOS scenarios can increase from the energy reflected from building materials. In addition, RMS DS can decrease in NLOS regions because weaker reflections are highly attenuated. We analyze the effect of reflecting objects on PL parameters, received power, AoA, RMS DS, and channel stationarity distance in this chapter.

5.2 PATH LOSS MODEL

We computed the CI model parameters for both indoor and outdoor settings. The CI model is a common PL model, whose equation is provided in (5.1). This model is parameterized by PL exponent \( n \) and large-scale standard deviation (SD) \( \sigma \).
\[ PL^{CI}(f, d)[dB] = FSPL(f, d_0) + 10n\log_{10}\left(\frac{d}{d_0}\right) + X^{CI}_\sigma \quad (5.1) \]

for \( d \geq d_0 \), where \( d_0 = 1 \) m. Here, large-scale channel fluctuations (due to shadowing in NLOS conditions) are contained within the variable \( X^{CI}_\sigma \), typically modeled as a zero mean Gaussian random variable with SD \( \sigma \) in dB. In (5.1), \( d \) is the 3D Tx to Rx separation distance, \( FSPL(f, d_0) \) denotes the free space PL in dB between Tx and Rx at distance \( d_0 \) at the carrier frequency \( f \), and we can determine \( FSPL(f, d_0) \) using (2.1) in chapter 2. The PL exponent \( n \) and the SD are obtained from a classical least-squares fit to the data. Here, we discuss PL for both indoor corridors and outdoor street environments.

5.2.1 Measured Results

5.2.1.1 Indoor Corridor Environments

Path loss versus link distance for the 1A lobby LOS setting is shown in Figure 5.1. Free-space PL is shown for reference. The CI PL model slope is 2.21 and standard deviation is 1.58 dB. In Figure 5.2, we observe that the 1 A lobby has an open space with large glass windows on one side and a ceramic tile floor. The Tx signal gets reflected from the glass window and the floor tile. We also observed that measured PL is larger than FSPL. Although the glass is a good reflecting material, the Rx receives an attenuated signal outside the antenna’s HPBW.

It is evident from Figure 5.2 that the Rx gets reflected signals from the floor within the antenna HPBW, if the link distance is more than 10 m. The Rx also captures the reflected signal from the large glass window within the HPBW of the antenna beyond 15 m link distance, see Figure 5.2. Hence, we noticed in Figure 5.1 that the PL decreases beyond 10 m link distance.
We also conducted measurement in a corridor environment, SWGN 3\textsuperscript{rd} floor indoor, and determined PL versus link distance and CI model parameters, shown in Figure 5.3. The PL model slope is 1.85 and the SD is 2.98 dB. A waveguiding effect can result for the MPCs in indoor environments [17] [65]. For distances less than 10 m, measured path loss
results are larger than free space. Rays are reflected at small incident angle by mostly dry
down, and are beyond the antenna HPBW within this distance. Beyond 10 m, we observe
PL decreases, because many rays are likely to be received within the antenna HPBW due
to reflective materials, e.g., metal-mesh-loaded glass, door and window metal frames; see
Figure 5.4.

Figure 5.3  PL vs. link distance for 3rd floor hallway, showing measured data and model fits.

Figure 5.4  Plan view diagram for 3rd floor hallway.
5.2.1.2 Outdoor Street Environments

Several reflecting materials, e.g., concrete, metal, brick, etc., are present in the outdoor environment. Path loss versus link distance for the outdoor LOS setting of 300 Main Street is shown in Figure 5.5. The CI model slope is 1.61. One reason for our slopes less than 2 is the directional antennas, but we also expect some waveguiding by the reflectors, e.g., concrete ground, parked cars, brick and concrete wall, and metal poles. The plan view diagram, showing the wall and car reflections, is illustrated in Figure 5.6.

![Figure 5.5 PL vs. link distance for 300 Main Street outdoor, showing measured data and model fits.](image-url)
Measurements were also conducted in a rich reflecting environment at the outside of the U of SC building, SWGN NW. We illustrate PL versus link distance in Figure 5.7 and computed the CI model parameters. In addition, we show two ray model path loss considering different incident angles and reflection coefficients. The PL model slope is 1.67 for SWGN NW. Our directional antenna captured strong reflected signals from nearby potential reflectors: concrete ground, stair rails, bike rack, pole, concrete wall, and pylon. For this reason, the PL exponent is appreciably smaller than the FSPL exponent. Figure 5.8 shows the plan view diagram. Rays shown by dotted green and orange are the rays outside the antenna HPBW, and the rays illustrated by dotted red are inside the antenna HPBW.

We observe that two ray model shows fairly good agreement with our measured data. The Rx likely gets a strong reflected signal from the ground within the antenna
HPBW, when the link distance is larger than 12 m, illustrated in Figure 5.8.

**Figure 5.7 PL vs. link distance for SWG NW outdoor**, showing measured data and model fits.

**Figure 5.8 SWGN NW outdoor**, left and right: plan view diagram and ground reflection, (red within antenna HPBW, green and orange are outside antenna HPBW).
5.2.2 WI Ray Tracing Model Results

All indoor and outdoor environments were modeled directly in WI. Materials are defined in WI for walls, ceiling tiles, floor tiles, doors, and windows from [22]. Antenna specifications were obtained from [53], where as noted, our 91 GHz system uses horn antennas. The Tx is set at a fixed point whereas the Rx moves along the route that follows the measurements. Three Rx positions at each distance were used to account for small scale fading, emulating our measurement procedure. The X3D model is the most comprehensive “flagship propagation model” in WI, which is used for simulation; this is valid for both indoor and outdoor environments. We set the following parameters: a maximum of 25 stored propagation paths, 0.25-degree ray spacing for azimuth angle, 6 reflections, 0 transmissions, and 1 diffraction, as suggested by WI [27].

5.2.2.1 Indoor Corridors

The PL model using WI was computed and is illustrated in Figures 5.9 and 5.10, for 3rd floor hallway and 1st floor lobby, respectively. We observe a good agreement in CI model PL between WI and the measured data. However, at distances larger than 10 m, the WI PL fluctuates from measured PL for 3rd floor hallway; a maximum of 10 dB larger in WI (see Figure 5.3 and 5.9). It is also observed that the WI PL is larger than measured PL at 2 m, approximately 5 dB, in 1st floor lobby (see Figure 5.1 and 5.10).
Figure 5.9  PL vs. link distance for 3rd floor hallway, showing WI data and model fits.

Figure 5.10  PL vs. link distance for 1st floor lobby, showing WI data and model fits.
5.2.2.2 Outdoor Streets

We also computed WI models for outdoor street environments in SWGN NW and 300 Main Street, shown in Figures 5.11 and 5.12, respectively. A large PL is estimated by the WI model compared to the measured PL in SWGN NW outdoors. We did not consider all potential reflectors, e.g., pylons, metal poles, bike rack, stair rail, etc., in the WI model whose inclusion could reduce PL. The effect of the number of cars on PL in 300 Main Street was examined via the WI model and results are illustrated in Figure 5.12. When we included more than one car between the Tx and the Rx, PL decreased for distances larger than 50 m: PL exponent decreased by 0.6 and SD by 1.72 dB. The effect of including the 2nd and 3rd cars is approximately the same as we conducted measurements up to a maximum 78 m link distance.

![Swearingen Northwest Outdoor](image.png)

Figure 5.11 PL vs. link distance for SWGN NW outdoor, showing WI data and model fits.
5.2.3 Comparison Between Measured and WI PL

PL exponents and standard deviations of CI models for both indoor and outdoor environments are shown in Table 5.1. It is worth noting that the difference between the measured and WI PL exponent is small for all environments, except SWGN NW. A possible reason for this may be that we have not yet accounted for all possible reflectors in that setting. The WI PL SD is larger than the measured SD. The outdoor PL exponents are smaller than indoor PL exponents, because most of the indoor materials absorb rays more than the outdoor material, e.g., dry wall, ceiling tile, i.e., the reflected signal from indoor dry wall is weaker than the outdoor concrete wall reflection. The SD for all environments is less than 3 dB.
Table 5.1 PL CI model parameters using measured and WI data in different environments.

<table>
<thead>
<tr>
<th>Environments</th>
<th>Measured</th>
<th>WI</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>n</td>
<td>σ [dB]</td>
</tr>
<tr>
<td>Indoor</td>
<td></td>
<td></td>
</tr>
<tr>
<td>1st Floor Lobby</td>
<td>2.21</td>
<td>1.58</td>
</tr>
<tr>
<td>3rd Floor Hallway</td>
<td>1.85</td>
<td>2.98</td>
</tr>
<tr>
<td>Outdoor</td>
<td></td>
<td></td>
</tr>
<tr>
<td>300 Main Street</td>
<td>1.61</td>
<td>2.21</td>
</tr>
<tr>
<td>SWGN NW</td>
<td>1.67</td>
<td>1.26</td>
</tr>
</tbody>
</table>

5.3 RECEIVED POWER FOR LOS TO NLOS TRANSITION

As noted in the prior chapter, once beyond the optical LOS, we rotated the Rx antenna in azimuth to the angle at which received power was maximum. We found maximum received power and azimuth angle at each distance when the Rx was moved from the LOS to NLOS zone for both indoor and outdoor environments. Geometric analysis has also been done for all environments to support the measured results, e.g., maximum received power and AoA.

5.3.1 Indoor Corridors

Figures 5.13, 5.15 and 5.17 show (maximum) received power and azimuth angle of maximum power vs. distance into the NLOS region, for corridors 1A lobby, 2D hallway, and 3D hallway, respectively. Received power generally decreases with NLOS distance,
and this holds for all indoor and outdoor settings. For the 1A hall (Figure 5.13), we observe a substantial increase in received power at a NLOS distance of 3.5 m. This is where a large (~1.5 m) trash canister became “visible” to the Rx antenna’s main beam, hence allowing a strong reflection to be captured, increasing power by approximately 14 dB. This reflection carries the maximum power for a distance approximately equal to its size of 1.5 m, showing the “nearly optical” behavior of smooth metal at this frequency. The azimuth angle of maximum received power also indicates this reflection for the 1A hallway.

In Figure 5.14, we observed Rx antenna azimuth angle orientation for maximum received power in the 1st floor lobby at different Rx distances: at 250 cm, 300 cm and 350 cm. We noticed that at 350 cm Rx distance the antenna main beam aimed at the metal canister.

![Graph](image)

Figure 5.13 Hallway 1A azimuth angle and maximum received power vs. NLOS distance.
For the 2D hall, the azimuth angle tends to increase as distance into the NLOS region increases. This indicates that the Rx antenna should aim away from the wall that is in the direct line to the Tx, and instead aim toward the open hall toward the LOS region, see Figure 5.15.

In Figure 5.16, we observe a +45-degree azimuth angle when the receiver is 90 cm away from the reference, i.e., the receiver is 70 cm away from the obstructing wall edge of the NLOS path. We believe that the strongest components come from a reflection from the metal door frame which is located on the opposite wall of the NLOS path.
Figure 5.15 Hallway 2D azimuth angle and maximum received power vs. NLOS distance.

Results for the 3D hall in Figure 5.17 are also augmented by geometric analysis. Specifically, the colored blocks or bands denote geometrically estimated “shadows” by the metal-framed, wire-mesh loaded windows and reinforced fire doors along the long wall of the 3D hall. This wall is the one between the Tx and the Rx when the Rx is in the NLOS zone. Based on attenuation values from [32], the azimuth pointing angle results lead us to hypothesize “through-drywall” transmission for the non-shadowed regions of Figure 5.17. In the shadowed regions, the Rx antenna cannot receive the shadowed through-wall signals, and hence must point toward the hall opening for maximum received power. The penetration loss of plywood (8.10 dB/cm for 73 GHz, [33]) and metal mesh [67] are very large.
Figure 5.16 Hallway 2D azimuth angle at 90 cm from reference, showing probable reflection from metal door frame.

A plan view diagram is illustrated in Figure 5.18, and from this diagram we were able to estimate the shadow regions attributable to different components, made of different materials.
materials, in the obstructing wall. We used this geometric-structural approach to analyze the “shadow diagram” along with the Rx position at different azimuth angles, shown in Figure 5.18 and Figure 5.19, respectively. The Rx at 32 and 86 cm received maximum power due to reflection from the metal door frame or metal door lock on the opposite wall, and at 74 cm, 88 cm, and 110 cm, the Rx received maximum power due to through-wall transmission. Moreover, we observed some azimuth angles close to zero degrees. We expect there may be metal objects within the utility room at the LOS-NLOS corner, but these are not yet accounted for by our geometric analysis. We also note additional MPCs are present at angles corresponding to shadows, thus even though the azimuth angle of maximum power can change abruptly, received power does not always do so. Regardless of the physical mechanism behind the effect, it is clear that small obstacle attenuations can cause fairly abrupt or rapid changes in the angle of maximum received power, which adaptive antenna systems will need to track. 

Figure 5.18 Plan view diagram (using AutoCAD) for 3D hallway at right and material shadow through NLOS wall at left.
Figure 5.19 3D hallway reflection and through wall transmission clockwise: at 86 cm, 110 cm, 88 cm, and 32 cm from reference at $29^0$, $33^0$, $-7^0$, and $3^0$ azimuth angles, respectively.

5.3.2 Outdoor Streets

Figure 5.20 shows (maximum) received power and azimuth angle of maximum power vs. distance into the NLOS region for SWGN NW outdoors. It is worth noting that the outdoor received power is approximately -32 dBm up to 56 cm from the reference, although the distance from the obstructed wall edge and the reference is 20 cm. In this outdoor case, we expect that reflections from objects, e.g., pylon, account for this larger distance for maximum received power. The azimuth angle should be negative if we move into the NLOS region. Moreover, at approximately 56 cm into the NLOS region, received power drops by over 17 dB in approximately 15 cm ($> 1$ dB/cm).
Figure 5.20 SWGN NW outdoor azimuth angle and maximum received power vs. NLOS distance.

Figure 5.21 shows antenna azimuth angle orientation for different received power conditions i.e., LOS, NLOS considering reflection and NLOS considering diffraction. Although the distance from the obstructed wall edge and the reference is 20 cm, the LOS zone is 33 cm; this is explainable from geometry, as the Rx is 1 m behind the wall and the antenna HPBW is 32 degrees. For this reason, if the Rx antenna rotates counterclockwise, it achieves a LOS path with the Tx antenna. We observed approximately constant received power up to 56 cm from the reference due to reflection from the small concrete surface, i.e., pylon, illustrated in Figure 5.21. In addition, beyond 56 cm we observed a substantial drop in received power as the Rx might receive only a weak diffracted signal from the sharp edge of concrete wall.
In 300 Main Street, illustrated in Figure 5.22, we observed a large value of received power: -26.8 dBm at 8 cm and 3 degrees azimuth angle. This is due to reflection from the metal drain pipe attached to the wall. We also observed a substantial drop in Rx power, e.g., a 25 dB drop within 28 cm (again, ~ 1 dB/cm). There is no reflecting material in this open space. For this reason, Rx power decreased nearly monotonically when we moved the Rx into the NLOS zone. Figure 5.23 shows the LOS scenario and NLOS scenario when the Rx obtained a weak diffracted signal from the sharp edge of the brick wall.

Figure 5.21 SWGN NW outdoor Rx antenna position clockwise: LOS, NLOS (reflection), NLOS (diffraction).
Figure 5.22 300 Main Street outdoor azimuth angle and maximum received power vs. NLOS distance.

Figure 5.23 300 Main Street outdoor Rx antenna position Left and Right: NLOS and LOS.
5.3.3 Comparison Between Indoor and Outdoor Received Power and Azimuth Angle

We observed that maximum received power in the first part of the NLOS region stays relatively close to that in the LOS region in the outdoor scenarios. Figure 5.24 shows that the LOS zone is only 20 cm. The drop in NLOS power after that is relatively steep (~15 dB/10 cm) for outdoors. In contrast, we noticed a more gradual decrease in received power (~15 dB/20 cm), when the Rx moved from LOS to NLOS regions in the indoor environment. For both indoor and outdoor transition settings, we obtained maximum received power when we turned the antenna away from the attenuating wall and aimed towards the open space (see Figure 5.25), which allows reception via reflections.

Figure 5.24 Normalized maximum Rx power vs. NLOS distance for all indoor hallways and outdoors street environments.
5.4 POWER DELAY PROFILES

Example sequences of PDPs for both indoors and outdoors are illustrated in Figure 5.26. For all environments, as expected, the number of strong MPCs decreased when we moved the Rx from the LOS to NLOS zone. We removed all MPCs 40 dB or more below the peak. It is worth noting the strong reflected MPC arising at approximately 3.5 m for the 1st floor lobby. Figure 5.14 shows the existence of a metal canister at the opposite of the NLOS path and clearly illustrates the evidence of receiving a strong signal. In addition, for SWGN NW outdoors, we noticed a large number of MPCs up to 56 cm and beyond that, the number decreases substantially. This result conforms to the Rx power and Rx antenna orientation results of Figures 5.20. We expect this due to some reflectors, e.g., pylon. Root mean square DS and SPCC, two crucial channel parameters, have been computed from
measured PDPs and are described in the following subsection.

Figure 5.26 PDP along LOS to NLOS transition, clockwise from upper left: 1st Floor Lobby, 2nd Floor and 3rd Floor Hallway, SWGN NW and 300 Main Street Outdoor.
5.4.1 Root Mean Square Delay Spread.

Fading effects in wideband channels are described as “frequency selective fading” because the BW is sufficiently large so that different PL and fading effects can occur for each frequency component within the transmission band. Coherence BW is defined as the maximum separation in frequency over which the propagation channel amplitude response is essentially flat. Coherence BW is inversely proportional to DS, where RMS DS defines the spread of propagation delays of a multipath channel. The Rx should be designed to account for DS effects because once the range of delays becomes large enough, this yields frequency selectivity, which if not compensated for can cause errors at the receiver. Here, we quantify RMS DS for different indoor and outdoor settings, for both LOS and NLOS environments.

We observe the typical result that RMS-DS increases significantly when moving from the LOS to a NLOS region, from a few ns to several tens of ns. We applied a 6 dB minimum MPC signal-to-noise ratio (SNR) threshold, and a 100 ns maximum MPC delay threshold for all PDPs before computing RMS-DS. We found RMS DS using [52]:

\[ \sigma_r = \sqrt{\tau^2 - \bar{\tau}^2} \] (5.2)

Here, mean excess delay, \( \bar{\tau} = \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)} \), and the second moment of the delay, \( \tau^2 = \frac{\sum_k P(\tau_k) \tau_k^2}{\sum_k P(\tau_k)} \) where \( P(\tau_k) \) is the power at delay \( \tau_k \).

For the 1A lobby, in addition to increasing RMS-DS as we transition from LOS to NLOS, Figure 5.27 shows the sharp decrease in RMS-DS when the strong MPC from the metal canister becomes visible. Figure 5.28 illustrates the RMS DS statistics for 2nd floor and 3rd floor hallway, SWGN NW and 300 Main Street outdoor environments. It is worth noting that the RMS DS are less than 5 ns up to the NLOS distance of 56 cm in SWGN
NW outdoors, although the LOS distance is 20 cm from the reference position.

Figure 5.27 RMS-DS vs. distance in 1st Floor Lobby.

Figure 5.28 RMS-DS vs. distance in indoor and outdoor settings.

In both indoors and outdoors, RMS DSs are less than 30 ns except for the 1st floor lobby, observed in Table 5.2. For all indoor and outdoor environments, NLOS mean RMS
DS is less than 15 ns whereas the LOS mean RMS DS is less than 7 ns. We observe the largest overall DS SD $\sigma_{rms}$ in the 1st floor (10.2 nsec) and the smallest overall $\sigma_{rms}$ in SWG NW outdoor (2.8 nsec).

Table 5.2 RMS DS in indoor and outdoor settings.

<table>
<thead>
<tr>
<th>Environments</th>
<th>RMS DS [ns]</th>
<th>$\sigma$ (ns)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Min</td>
<td>Max</td>
</tr>
<tr>
<td>Indoor</td>
<td></td>
<td></td>
</tr>
<tr>
<td>1A</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LOS</td>
<td>5.9</td>
<td>5.9</td>
</tr>
<tr>
<td>NLOS</td>
<td>4.9</td>
<td>38.6</td>
</tr>
<tr>
<td>Overall</td>
<td>4.9</td>
<td>38.6</td>
</tr>
<tr>
<td>2D</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LOS</td>
<td>6.2</td>
<td>8.6</td>
</tr>
<tr>
<td>NLOS</td>
<td>5.9</td>
<td>23.7</td>
</tr>
<tr>
<td>Overall</td>
<td>5.9</td>
<td>23.7</td>
</tr>
<tr>
<td>3D</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LOS</td>
<td>4.3</td>
<td>10.5</td>
</tr>
<tr>
<td>NLOS</td>
<td>4.4</td>
<td>28.6</td>
</tr>
<tr>
<td>Overall</td>
<td>4.3</td>
<td>28.6</td>
</tr>
<tr>
<td>Outdoor</td>
<td></td>
<td></td>
</tr>
<tr>
<td>300 Main</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LOS</td>
<td>4.4</td>
<td>5.2</td>
</tr>
<tr>
<td>NLOS</td>
<td>4.5</td>
<td>26.7</td>
</tr>
<tr>
<td>Overall</td>
<td>4.4</td>
<td>26.7</td>
</tr>
<tr>
<td>SWGN NW</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LOS</td>
<td>4.4</td>
<td>4.8</td>
</tr>
<tr>
<td>NLOS</td>
<td>4.5</td>
<td>15.8</td>
</tr>
<tr>
<td>Overall</td>
<td>4.4</td>
<td>15.8</td>
</tr>
</tbody>
</table>
We also determined delay spread using two different metrics that are sometimes
used: the $X\%$-energy delay window $W_X$, and the $Y$ dB delay interval $I_Y$ [54]. These were
computed at each distance as we moved from LOS to NLOS for SWGN NW outdoor.
Results are shown in Figure 5.29. We determined the delay window containing $X=90\%$ and
$85\%$ of the total energy in the middle portion of the PDP, and the delay interval considering
$Y=20$ dB and $25$ dB thresholds. All these measures of DS are less than $5$ ns up to NLOS
distance of $56$ cm; we observed approximately constant maximum received power up to
this distance (cf. Figure 5.20). Delay spread using $I_{25}$ increases to about $80$ ns, and all other
DS metrics increase to ~ $25$ ns beyond $56$ cm.

![Swearingen LOS to NLOS transition DS measures vs. distance.](image)

Figure 5.29 SWGN NW outdoor LOS-to NLOS transition DS measures vs. distance.
5.4.2 Spatial PDP Correlation Coefficients

The spatial correlation coefficient is a very useful quantity to show the change of the channel properties at different antenna positions and frequency bands [55], as a function of distance. We computed sets of PDP correlation coefficients, via the method in [56], to assess the rate of spatial channel variation. We compute the SPCC $c(\Delta x, x_i)$ using our average PDP $P_{avg}(\tau, x_i)$ at a given location $x_i$ as follows:

$$c(\Delta x, x_i) = \frac{\int P_{avg}(\tau, x_i)P_{avg}(\tau, x_i+\Delta x)d\tau}{\max\{\int [P_{avg}(\tau, x_i)]^2 d\tau \int [P_{avg}(\tau, x_i+\Delta x)]^2 d\tau\}^{1/2}}$$  \hspace{1cm} (5.3)

The similarity between the average PDP at locations $x_i$ and $x_i+\Delta x$ is quantified by this metric. It is noted that $0 \leq c(\Delta x, x_i) \leq 1$. Figure 5.30 shows the SPCC at different Rx positions for both indoor and outdoor environments when the Rx was moved from LOS to fully obstructed zone. Results in Figure 5.30 agree with those for RMS-DS in Figure 5.28: specifically in the 3D hallway, the SPCC is “broader” when RMS-DS is approximately constant or very slowly changing, from ~5-20 cm, 27-45 cm, 50-60 cm, and 90-110 cm. Alternatively stated, the SPCC is very narrow, with large values only on the diagonal, when RMS-DS changes rapidly. For SWGN NW outdoor, the SPCC is again “broader” when DS is approximately constant or very slowly changing, up to 56 cm.

For a threshold $c(\Delta x, x_i) \geq 0.7$, we can define a “stationarity distance”: the NLOS stationarity distance ranges from as small as approximately 2 cm up to 16 cm, or approximately only 6 $\lambda$ to 54 $\lambda$, observed in Table 5.3. We noticed this small stationary distance due to the presence of few reflectors, and MPCs might change rapidly in space. The short correlation distance in most cases is favorable for spatial multiplexing in MIMO, since it allows uncorrelated spatial data streams to be transmitted from closely spaced (a
fraction to several wavelengths) antennas [57]—as long as several MPCs are actually available at the Rx.

Figure 5.30 Spatial PDP correlation coefficient vs. distance along LOS to NLOS transition, clockwise from upper left: 1st Floor Lobby, 2nd Floor and 3rd Floor Hallway, SWGN NW and 300 Main Street Outdoor.

In addition, we observe a “broadening” of the region of large values of SPCC when the Rx moved away from the Tx in both indoor and outdoor LOS scenarios, illustrated in Figure 5.31. For a threshold $c(\Delta x, x_i) \geq 0.7$ (at and above which the CIR is changing only by
a small amount), the stationarity distance ranges from as small as approximately 2 m up to 8 m, or approximately 600 $\lambda$ to 2667 $\lambda$. The larger values of stationarity distance are expected in LOS conditions: with a strong LOS component and only few and often weak MPCs, the channel’s CIR changes only very slowly.

We observed the largest stationarity distance in the 3D hallway, shown in Table 5.3. In 3D hallway, there is a large number of reflectors: metal door frame, metal door lock, metal framed windows, metal mesh inside glass window, etc., which can cause a large number of MPCs, but the rate of change of these large number of MPCs should be small because their angle of arrival changes only very slowly in this corridor LOS environment. The minimum NLOS stationary distance is approximately a factor of 100 smaller than that for the LOS measurements, except for the 1st floor lobby. This means the channel varies much more slowly in space in LOS than in NLOS scenarios.

Table 5.3 SPCC length (for $c \geq 0.7$) in indoor and outdoor settings.

<table>
<thead>
<tr>
<th>Environments</th>
<th>SPCC Length [cm]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LOS</td>
</tr>
<tr>
<td></td>
<td>Min</td>
</tr>
<tr>
<td>Indoor 1st Floor Lobby</td>
<td>200</td>
</tr>
<tr>
<td>2nd Floor Hallway</td>
<td>-</td>
</tr>
<tr>
<td>3rd Floor Hallway</td>
<td>200</td>
</tr>
<tr>
<td>Outdoor 300 Main Street</td>
<td>200</td>
</tr>
<tr>
<td>SWGN NW</td>
<td>200</td>
</tr>
</tbody>
</table>
Figure 5.31 Spatial PDP correlation coefficient vs. distance, LOS settings, clockwise from upper left: 1st Floor Lobby, 3rd Floor Hallway, SWGN NW and 300 Main Street Outdoor.

5.5 COMPARISON WITH RESULTS FROM THE LITERATURE

In this section, we compare our 90 GHz band wireless channel characterization parameters for LOS and NLOS settings with few existing works in other bands. For an indoor office building, the authors of [15] found the PL exponent at 28 GHz and 142 GHz as 1.7 and 2.1, respectively for LOS settings. The HPBW of Tx/ Rx is 30° for 28 GHz, and
8° for 142 GHz. The authors of [66] and [69] determined PL exponent 1.45 and 2 in indoor hotspot scenario at 28 GHz, and indoor long hallway and moderate-length corridor at 60 GHz, respectively. In our 90 GHz setup, we calculated PL exponent ~2 for indoor LOS settings.

For LOS urban microcellular environment, the authors of [61] found a PL exponent of 2.1 for 28 GHz and 38 GHz. The PL exponent was also estimated as 1.9 for an urban scenario in [63]. We calculated PL exponent for outdoor LOS settings less than 2 (~1.6). Concrete and brick wall, concrete pylons, concrete ground, metal bike rack, metal stair rail, metal sign poles and cars are observed in our outdoor environment. As reflection coefficient of concrete [60], metal and glass are large, we observe smaller PL than FSPL.

In an urban environment at 28 GHz, the authors observed that the propagation loss during the LOS to NLOS transition scenario increases rapidly, about 30 dB [62]. We observed 17 dB increase in PL for our 90 GHz transmission when we moved the Rx behind an obstacle in NLOS zone.

RMS DS for LOS settings in a library environment [68] and office room measurement [70] were maximum 44.4 ns and mean 13 ns at 28 GHz and 60 GHz, respectively. The maximum RMS DS is 62.21 ns at 28 GHz for NLOS setting in library environment [68]. For our 90 GHz measurements, the maximum RMS DS for indoor LOS and NLOS settings are 10.5 and 38.6 ns, respectively. As the PL is larger at our higher frequency, few dominant MPCs are captured by the Rx with directional antennas, yielding a smaller RMS DS in our 90 GHz results.

Authors of [61] determined median RMS DS values for an urban microcellular environment at 28 GHz and 38 GHz: LOS 10.8 ns at 28 GHz, and 8.4 ns at 38 GHz, and
for NLOS, 44.6 ns and 40 ns, for the two respective frequencies. Another NLOS urban scenario in [63], the authors found less than 60 ns average DS at 28 GHz. We observe smaller RMS DS values in our 90 GHz band: LOS 4.4 to 5.2 ns, and NLOS 4.5 to 26.7 ns.

Based on the similarity of PDPs, the authors of [62] determined the average correlation distance at 28 GHz in urban environment. For the LOS route, the correlation distance is 0.9 m. At the beginning section of the LOS to NLOS transition route, the correlation distance reached up to 4 m, and afterwards dropped to about 1.26 m. For our 90 GHz sounder setup, we observed comparatively large stationarity distance for LOS settings, maximum 6 m. However, for NLOS settings, the stationarity distance is only 16 cm. Hence in summary, our results are comparable to those we have found in the literature, with differences as expected due to larger path loss or smaller beamwidths for our 90 GHz case.
CHAPTER 6
AIRPORT ENVIRONMENT MMWAVE CHANNEL MEASUREMENT RESULTS

6.1 MMWAVE CHANNEL CHARACTERIZATION AND PL MODELS

Path loss measurement and corresponding model construction are essential for link budget and coverage analysis. As the wavelength is small in the 90 GHz band (~3.33 mm), potential reflectors can be used to overcome PL of an obstructed direct component. Indoor office environments are quite different from indoor airport areas, in our case an airport maintenance hangar and an airport baggage area. Passengers move about in the airport terminal building, and the airport baggage area is also periodically crowded during baggage claims. The airport maintenance hangar is full of many potential movable and fixed reflectors, e.g., tool cabinets, aircraft, etc. Because of this environment uniqueness, mmWave channel measurements and characterization in airport environments are crucial for both LOS and NLOS scenarios.

We performed measurements within the CUB airport maintenance hangar in Columbia, SC, and in the CAE airport baggage area in West Columbia, SC. In this chapter, we provide empirical PL models of the CUB maintenance hangar and CAE baggage area. Some important channel parameters, e.g., RMS DS, rate of change of maximum received power, rate of change of antenna azimuth angle, channel stationarity distance, etc., are quantified and discussed in this chapter.

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6.2 PATH LOSS MODEL

Path loss was computed using (5.1) for our two different airport environments: the maintenance hangar in CUB and the airport baggage area in CAE. In the maintenance hangar, we determined PL parameters for LOS, and mixed paths. We computed CI model PL parameters only for an LOS path within the CAE baggage area.

6.2.1 CUB Maintenance Hangar

The CI model parameters were computed for all 4 radial paths we measured within the CUB maintenance hangar. Figures 6.1-6.4 show the PL versus link distance for radial paths 1, 2, 3 and 4, respectively. We show FSPL as a reference in these figures, and observe that measured PL is larger than FSPL—expected of course for the mixed paths. The Rx antenna azimuth angle was adjusted for obtaining the largest Rx power at NLOS locations and in partially obstructed paths. The PL exponent \( n \) in path 1 is close to the FSPL exponent value of 2 because the Tx and the Rx were always in LOS. When the Rx was moved away from the Tx at distances of 18 m it was in front of, and at 24 m it was behind a V560-N577XW aircraft (approximate height 4.6 m, length 14.9 m, and wing area 31.8 m\(^2\)). The measured PL is small at these distances because the Rx was still in LOS and additional reflected signals were captured from nearby aircraft.
Figure 6.1 PL vs. link distance for path 1 at CUB maintenance hangar.

Paths 2 and 3 are the same path but with Tx and Rx locations interchanged. The Rx was mostly blocked by a wing of an aircraft in path 2 at 6 m link distance. We find large PL (~106 dB) at this distance. When we moved the Rx away from the Tx from this distance along this path, we observed a partially obstructed scenario and the PL decreased. This is due to reflected signals from nearby metal obstacles, in addition to the LOS signal. We observed a similar semi-obstructed scenario when the Rx was at 12 m link distance in path 3, behind an aircraft wing. A large PL, approximately 101 dB, was recorded at this distance. At 15 m link distance in both paths 2 and 3, we steered our Rx antenna to obtain maximum received power, and observed approximately 6 dB decrease in PL.
Figure 6.2 PL vs. link distance, path 2 at CUB maintenance hangar.

Figure 6.3 PL vs. link distance, path 3 at CUB maintenance hangar.
The Rx was fully obstructed by an aircraft at 9 m link distance in path 4. We placed the Rx behind the aircraft’s wing and steered our Rx antenna to obtain maximum received power. The Rx was aimed at a metal cylinder to yield maximum Rx power at this distance, for which we still observed large PL, 106 dB, shown in Figure 6.4. It is also worth noting that the maximum-power signal captured by the Rx was actually transmitted from outside the HPBW of the Tx antenna.

![Figure 6.4 PL vs. link distance, path 4 at CUB maintenance hangar, Columbia, SC, USA.](image)

The CI model parameters for the different paths are shown in Table 6.1. The PL exponents in LOS and mixed paths are \( \sim 2 \) and \( \sim 3 \), respectively. We also observed large standard deviation in path 2 (\( \sim 6 \) dB) due to large blockage attenuation. In addition, the PL exponent decreased by up to 0.2 when the Rx searched for the maximum Rx power in the mixed environment. The large number of scattering objects within the maintenance hangar
environment provide a smaller \( n \) and smaller \( \sigma \) than the typical office environment’s NLOS scenarios.

Table 6.1 CI Model Parameters, PL exponent and SD, for 4 radial paths at CUB maintenance hangar.

<table>
<thead>
<tr>
<th>Paths</th>
<th>PL exponent (n)</th>
<th>Standard deviation [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LOS</td>
<td>Partially Obstructed</td>
</tr>
<tr>
<td>1</td>
<td>2.1</td>
<td>-</td>
</tr>
<tr>
<td>2</td>
<td>-</td>
<td>3.1</td>
</tr>
<tr>
<td>3</td>
<td>2.1</td>
<td>2.8</td>
</tr>
<tr>
<td>4</td>
<td>2.2</td>
<td>-</td>
</tr>
</tbody>
</table>

6.2.2 Airport Baggage Claim Area

Measurements were taken in LOS, and mixed environments in the CAE baggage claim area for three transmitter positions; see plan view diagram in Figure 4.6 (chapter 4). We quantified CI model parameters for LOS path for Tx1: PL exponent 2.3 (close to FSPL) and standard deviation is 3.5 dB, illustrated in Figure 6.5. At some link distances, e.g., 20 m, 25 m, 40 m and 50 m, the PL is smaller than other distances. This is due to the capture
of both LOS and reflected signals from metal pillars and potentially a ceiling-mounted metal speaker grill. It is also noticeable that the PL is large, approximately 114 dB, at 35 m and 45 m link distances. The Rx was close to carousel 1 at those distances, and could receive weak reflected signals from the carousels, pillars, or sign holders. It is evident in the Figure 6.5 that the measured results show approximately two ray type effect, a reflected ray might be from the floor.

Figure 6.5 PL vs. link distance, LOS path at CAE baggage area.

Measurements for mixed paths were also taken near CAE baggage areas 1, 2, and 3, moving the Rx behind the carousels, for Tx location 1. At each path, the measurements were performed for LOS, NLOS and partially obstructed settings, starting from a reference LOS position, and again we captured maximum Rx power by steering the Rx antenna along the azimuth plane. It is illustrated in Figure 6.6 that PL is smallest in LOS and largest in NLOS, as expected. The semi-obstructed scenario PL is always in between the LOS and NLOS PL. The range between the LOS and the NLOS PL is 15 dB at path 3, 7 dB at path
2, and 18.9 dB at path 1. Path 2 is surrounded by two carousels, several pillars, and sign holders. Strong reflected signals could be received in these NLOS positions, yielding a small PL range at path 2.

Measurements were also taken for the LOS to NLOS transition path for Tx1, at the top of the ramp, by moving the Rx behind a drywall corner. The reference link distance is 75 m, and the Rx was moved behind the corner as the hallway made a right angle. As the Rx was totally obstructed by the wall, we captured the maximum received power by aiming the Rx antenna at the pillar of the 1\textsuperscript{st} carousel. The PL range is 11 dB at the ramp. The FSPL values for paths 1, 2, 3 and top of the ramp are approximately 105 dB, 102 dB, 96 dB and 109 dB, respectively. We observed for all LOS cases (except path 3) the measured PL is smaller than FSPL, due to the geometry allowing reception of strong reflections.

![PL for CAE Baggage area, Tx1 position](image)

Figure 6.6 PL vs. link distance for three different paths, LOS, NLOS and partially obstructed path, at CAE baggage area, for 4 paths from the left: behind carousels 3, 2, 1 and at the top of the ramp.

In addition, we conducted measurements at Tx2, near carousel 2 for three settings: LOS, NLOS and partially obstructed. The FSPL at these distances is approximately 98 dB.
It is illustrated in Figure 6.7 that all LOS PL values are smaller than FSPL. The maximum Rx power at the NLOS distance was obtained by aiming the Rx antenna at the metal pillar, near carousel 1. We observed a large PL, approximately 109 dB, in the partially obstructed location, with obstruction caused by the pillars near carousel 2. Measurements for Tx3 used the same Tx position as Tx2. The Tx antenna was placed parallel to the carousels for Tx2, whereas the Tx antenna was rotated 45 degrees anti clockwise for Tx3, standing behind the Tx. The Tx antenna was directed at one of the pillars of the carousel 1 for Tx3. Measurements were made at the same distances as for Tx2, but near carousel 2 for Tx3. The LOS PL values are smaller for Tx2 than for Tx3, shown in Figure 6.8. For Tx3, both LOS and partially obstructed paths, near carousel 2, were beyond the HPBW of the Tx antenna. Next, we conducted measurements behind carousel 1 for Tx3. A large PL, approximately 106 dB, was observed as it was partially obstructed by sign holders. The range of the PL for Tx3 near baggage area 2 is 3.7 dB and behind baggage area 1 is 7.2 dB, whereas a large PL range is observed in baggage area 2 for Tx2, 16.5 dB.

![PL for CAE Baggage Area 2, Tx2 position](image)

Figure 6.7 PL vs. link distance for three different scenarios, LOS, NLOS and partially obstructed, at CAE baggage area, for Tx2.
Figure 6.8 PL vs. link distance for three different scenarios, LOS, NLOS and partially obstructed, at CAE baggage area, for Tx3.

6.3 RECEIVED POWER FOR LOS TO NLOS TRANSITION

Measurements were taken in the CUB maintenance hangar by moving the Rx from LOS to NLOS, i.e., behind an aircraft, and by steering the Rx antenna to get maximum Rx power. Figures 6.9 and 6.12 illustrate received power at different distances for paths 1 and path 2, respectively. The Tx was 11 m away from the reference position of the Rx, and both the Tx and the Rx were boresight aligned. The Tx was fixed, and we moved the Rx by up to 9 m from the reference Rx position. The measurements were taken at intervals of 50 cm and the Rx was obstructed by the aircraft beginning at 150 cm. The reference antenna position is a straight line which connects the Tx and the Rx antennas in boresight. We could not conduct any measurement between 150 cm to 400 cm in path 1 because a large number of tools prevented access.
In path 1, when the Rx was fully obstructed by the aircraft, the Rx could capture reflected signals from the metal cabinets and the aluminum foil wall, 2 m behind the Rx. We observed substantial decrease in received power, approximately 17 dB, when the Rx moved from 100 cm LOS to 150 cm NLOS. A similar 17 dB decrease in received power was also noticed at 650 cm to 750 cm. The Rx antenna orientation for seeking the best received signal in NLOS is shown in Figure 6.10. We observed a large received power at 600 cm, approximately 3 dB less than LOS, reflected by the wall or cabinet. The Rx antenna azimuth angle should be less than -90 degrees if the reflected signal comes from the cabinet or wall. Figure 6.9 illustrates that the angle is less than -90 degrees at most of the NLOS distances, as expected. The received power range is 20.1 dB, and the range of angle is 155 degrees.

Figure 6.9 Maximum received power and azimuth angle vs. distance for path 1 at CUB maintenance hangar.
In path 2, we also moved the Rx up to 9 m from a reference LOS position behind an aircraft; see Fig. 6.11. The Rx was in NLOS between 150 cm and 850 cm. The LOS signal, which was beyond the HPBW of the Tx antenna, was captured by the Rx after 850 cm, shown in Figure 6.11. The Rx could capture reflected signals from the wall or engine, when it moved along the NLOS path, and we searched for the maximum received power by steering the Rx antenna. The received signal and azimuth angle for different distances are illustrated in Figure 6.12. As in path 1, a sharp decrease in received power was also observed when the Rx entered NLOS from LOS, approximately 14 dB. In contrast to path 1, the reflecting wall in path 2 was beside the Rx, not behind the Rx. When the Rx entered LOS to NLOS zone, the received power decreased 0.3 dB/cm for path 1, and 0.14 dB/cm for path 2.
Figure 6.11 Tx and Rx antenna position in reference, NLOS and LOS (beyond HPBW), path 2 at CUB maintenance hangar.

Figure 6.12 Maximum received power and azimuth angle vs. distance for path 2 at CUB maintenance hangar.
6.4 POWER DELAY PROFILES

6.4.1 Example Power Delay Profiles

6.4.1.1 Maintenance Hangar

Example sequences of PDPs for LOS, and mixed paths at the CUB maintenance hangar are illustrated in Figure 6.13. We removed all MPCs 40 dB or more below the peak. We observe that MPC powers decrease when we move the Rx far away from the Tx due to large PL; this is as expected, and was observed in all paths. We also observe that MPC powers decrease substantially when the Rx is partially (paths 2 and 3) or fully blocked by the aircraft (path 4), at 6 m, 12 m, and 9 m for paths 2, 3, and 4 respectively. In path 4, the Rx was fully obstructed by the aircraft and the Rx antenna was focused on a metal cylinder to obtain maximum received power.

Figure 6.13 Sequence of PDP at CUB maintenance hangar, clockwise from left, path 1, 2, 3 and 4.
We illustrate example PDPs for paths 1-4, in Figures 6.14-6.17. In the LOS path, when we moved the Rx behind an aircraft at 24 m, we observed many MPCs due to the reflections from nearby aircraft. For this reason, small PL and large RMS DS are expected at this distance; see Figure 6.1 for PL. When the Rx was in partially-obstructed and NLOS environments, the Rx antenna was rotated from the reference position to capture maximum received power. In path 2, at a link distance of 15 m, we received maximum power when the antenna was aimed at metal reflectors (+12 degree angle, anticlockwise, from the Rx perspective).

Figure 6.14 Example PDP for path 1 at CUB maintenance hangar, placing Rx at 6 m, in front of an aircraft (18 m) and behind the aircraft (24 m).
In path 3 (mixed path), maximum received power was obtained at 15 m link distance by rotating the Rx antenna clockwise from reference antenna position to +9 degrees. For this angle, the Rx was aimed at a metal reflector and could receive stronger MPCs than at the reference antenna position, yielding small PL, and smaller RMS DS. We also steered the Rx antenna when the Rx was fully obstructed by an aircraft in path 4. The maximum received power was obtained at azimuth angle -9 degrees by rotating the antenna clockwise, aiming at a metal cylinder.
Figure 6.16 Example PDP for path 3 at CUB maintenance hangar, showing reference and best antenna azimuth position.

Figure 6.17 Example PDP for path 4 at CUB maintenance hangar, showing reference and best antenna azimuth position.
Power delay profiles for the Rx along LOS to NLOS transition paths 1 and 2 are illustrated in Figures 6.18 and 6.19, respectively. The Rx was obstructed by the aircraft in paths 1 and 2, starting at 1.5 m from the reference. We observe diffracted MPCs from the aircraft tail at 1.5 m in path 1. When the Rx was moved along this NLOS path, strong reflected signals could result from the wall or cabinet behind the Rx. At 8.5 m from the reference, in path 2, the Rx was in LOS and the geometry is such that the received signal emanates from beyond the HPBW of the Tx antenna. For this reason, we observe both weak LOS and reflected components in the PDPs.

Figure 6.18 Example PDP for LOS to NLOS transition, path 1 at CUB maintenance hangar.
6.4.1.2 CAE Baggage Area

Power delay profiles were also obtained for LOS paths at CAE baggage area for Tx 1, illustrated in Figure 6.20. A reflection from a ceiling-mounted metallic speaker grill or pillars was observed. Specifically, at 20 m link distance, we obtained a probable reflection from a speaker grill, which is located approximately 2.5 m above the Tx and the Rx. It is shown in Figure 6.21 that the transmitted and received signals were within the HPBW of the antenna at this distance. The Rx could capture many weak MPCs reflected from pillars or carousels at 35 m, resulting in a large RMS DS, e.g., 22 ns.
Figure 6.20 PDPs in CAE Baggage area for Tx1.

Figure 6.21 Tx and Rx position at 20 m link distance in CAE Baggage area for Tx1.
Power delay profiles for the LOS to NLOS transition path, moving the Rx behind the wall corner at the top of the ramp, were computed, and are illustrated in Figure 6.22. A strong LOS component is observed at the reference position. The Rx antenna was steered and focused to an open area to capture reflected signals. The transmitted signals might be reflected from pillars or sign holders, located nearby the baggage area 1. We observe a weak LOS component and dominant reflected MPCs in Figure 6.22, when the Rx was in NLOS condition.

![Power delay profile for CAE Ramp @ reference (LOS)](image1)

![Power delay profile for CAE Ramp @ 1.6 m from reference (NLOS)](image2)

![Power delay profile for CAE Ramp @ 2.2 m from reference (NLOS)](image3)

Figure 6.22 Example PDPs for LOS to NLOS transition, at top ramp in CAE Baggage area for Tx1.

The Rx was moved behind each carousel for Tx1, and PDPs were determined for LOS, NLOS and partially obstructed conditions in baggage areas 1, 2 and 3, illustrated in Figures 6.23, 6.24 and 6.25, respectively. The largest power is observed in the LOS
component, as expected. In Figure 6.23, we observe PDPs for two LOS distances: one at reference point (45 m link distance), and another 9.9 m from the reference point (46.1 m link distance). The Rx could catch strong reflected signals from metal sign holders and metal pillars along with the LOS component at 46.1 m link distance.

![CAE Baggage Area 1 LOS, NLOS & Partially Obstructed PDP](image)

Figure 6.23 Sequence of PDPs for LOS, NLOS, and partially obstructed paths in CAE Baggage area 1 for Tx1.

In NLOS paths, the range of power among MPCs is small, compared to LOS and partially obstructed paths. Both boresight and reflected MPCs were found in partially obstructed paths. It is worth noting in Figure 6.25 that there is a 56 ns delay difference, approximately 17 m path length difference, between significant MPCs at the partially obstructed location in baggage area 3. It is illustrated in Figure 6.26 that the Rx could capture a reflected signal from the metal pillar located at the opposite side of the carousel 3.
Figure 6.24 Example PDPs for LOS, NLOS, and partially obstructed condition in CAE Baggage area 2 for Tx1.

Figure 6.25 Example PDPs for LOS, NLOS, and partially obstructed condition in CAE Baggage area 3 for Tx1.
Figure 6.26 The Tx and the Rx position behind carousel 3 for Tx1, showing pillar reflection at partially obstructed path.

For Tx2 and Tx3, the measurements were taken near baggage area 2 at the same distances, and PDPs were computed for LOS, NLOS and partially obstructed locations. Although the Rx was in NLOS at 21.93 m link distance, the Rx antenna for Tx2 could receive significant reflected signals from the pillar, located nearby baggage area 1, shown in Figure 6.27. It is also noticed in Figure 6.28 that 41 MPCs, within 20 dB from the peak, were captured in partially obstructed path for Tx3. It is mentioned in chapter 3 that any resolvable MPC must be 2 ns in delay from any other MPC. The Rx was placed between baggage areas 1 and 2 in this path, surrounded by many reflectors, e.g., pillars, sign holders, etc., yielding these MPCs.
Figure 6.27 Sequence of PDPs for LOS, NLOS, and partially obstructed paths near CAE Baggage area 2 for Tx2.

Figure 6.28 Example PDPs for LOS, NLOS, and partially obstructed paths near CAE Baggage area 2 for Tx3.
6.4.2 Root Mean Square Delay Spread

6.4.2.1 CUB Maintenance Hangar

As in our other measurement environments, for the airport settings we applied a 6 dB minimum MPC SNR threshold, and a 100 ns maximum MPC delay threshold for all PDPs before computing RMS-DS for all measurements. RMS DS for four radial paths (LOS and mixed) for the CUB maintenance hangar is shown in Figure 6.29. We observe the typical result that RMS-DS increases significantly when the Rx was partially or fully obstructed by an aircraft. Table 6.2 shows the RMS DS for LOS and mixed paths. The RMS DS is less than 8 ns for LOS path 1. It is also noticed that the maximum RMS DS is less than 17 ns for mixed paths. When we directed the Rx antenna to the largest received power direction, RMS DS decreases, as expected. For instance, RMS DS for paths 2 and 3 at 15 m reference antenna position are 11.5 ns and 16.13 ns, respectively. When we rotated the Rx antenna to the largest received power position, the RMS DS decreased to 8.26 ns and 11.4 ns for path 2 and 3, respectively. RMS DS decreases at the “best” antenna position because most of the MPCs become weaker relative to the dominant path.

![RMS delay spread for 4 radial paths](image)

Figure 6.29 RMS DS for four radial paths in LOS, NLOS and partially obstructed environments at CUB maintenance hangar.
Table 6.2 RMS DS statistics at CUB maintenance hangar for 4 radial paths.

<table>
<thead>
<tr>
<th>Paths</th>
<th>LOS [ns]</th>
<th>Mixed [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Min</td>
<td>4.54</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>7.62</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>5.62</td>
</tr>
<tr>
<td>2</td>
<td>Min</td>
<td>4.8</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>4.8</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>4.8</td>
</tr>
<tr>
<td>3</td>
<td>Min</td>
<td>4.71</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>4.76</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>4.73</td>
</tr>
<tr>
<td>4</td>
<td>Min</td>
<td>4.55</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>4.7</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>4.62</td>
</tr>
</tbody>
</table>

Root mean square DS values for the CUB maintenance hangar were computed for LOS to NLOS transition paths. It is observed in Figure 6.30 that RMS DS is small in the LOS region for both paths 1 and 2, less than 6 ns. We found a large RMS DS, approximately 16 ns, between 850 cm to 900 cm, although the Tx and Rx were in LOS. This is because the Rx captured a relatively strong signal that was beyond the HPBW of the Tx antenna. It is also worth noting that the RMS DS decreased to approximately 6 ns at 600 cm in path 1 and 300 cm in path 2, for NLOS settings; the Rx captured a large value of power, shown in Figures 6.9 and 6.12, at these distances. Table 6.3 shows the RMS DS statistics for LOS
to NLOS transition paths in the CUB maintenance hangar. The maximum and mean RMS DS are less than 24 ns and 14 ns, respectively.

Figure 6.30 RMS DS for path 1 and 2, LOS to NLOS transition, at CUB maintenance hangar.

Table 6.3 RMS DS statistics for the CUB maintenance hangar LOS to NLOS transition.

<table>
<thead>
<tr>
<th>Paths</th>
<th>LOS [ns]</th>
<th>NLOS [ns]</th>
<th>LOS (beyond HPBW) [ns]</th>
<th>Combined LOS &amp; NLOS [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Min</td>
<td>4.98</td>
<td>5.34</td>
<td>4.98</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>5.49</td>
<td>15.25</td>
<td>13.95</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>5.97</td>
<td>23.95</td>
<td>23.95</td>
</tr>
<tr>
<td>2</td>
<td>Min</td>
<td>4.87</td>
<td>5.88</td>
<td>4.87</td>
</tr>
<tr>
<td></td>
<td>Mean</td>
<td>5.23</td>
<td>10.75</td>
<td>10.29</td>
</tr>
<tr>
<td></td>
<td>Max</td>
<td>5.73</td>
<td>15.43</td>
<td>16.12</td>
</tr>
</tbody>
</table>
6.4.2.2 CAE Baggage Claim Area

We calculated RMS DS for the LOS path in the CAE baggage area, illustrated in Figure 6.31. RMS DS is less than 10 ns up to 30 m link distance, close to baggage area 2. RMS DS is also less than 10 ns at 40 m and 50 m link distance. It is observed in Figure 6.1 that measured PL is close to FSPL at these distances, which means a few dominant MPCs that are close in delay exist, resulting in a small RMS DS. A large RMS DS (approximately 22 ns) was found at 35 m, close to baggage area 1. The link distance was relatively large, and hence the Rx could receive multiple reflected signals with long delays in this position. In Figure 6.5, we observe a large difference between measured PL and FSPL at this distance, as expected. Table 6.4 shows the mean, minimum, maximum, and standard deviation of calculated RMS DS. The mean RMS DS is 8 ns, and the range is 16 ns. The relatively large RMS DS of 16 ns results because there are many potential reflectors, e.g., pillars, sign holders and carousel, near this LOS path.

![RMS delay for CAE LOS path](image)

Figure 6.31 RMS DS for LOS path at CAE.
Table 6.4 RMS DS for LOS path at CAE baggage area.

<table>
<thead>
<tr>
<th></th>
<th>RMS DS [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Min</td>
<td>4.03</td>
</tr>
<tr>
<td>Max</td>
<td>20.46</td>
</tr>
<tr>
<td>Mean</td>
<td>8.44</td>
</tr>
<tr>
<td>Standard deviation</td>
<td>5.44</td>
</tr>
</tbody>
</table>

As for PL, we moved the Rx behind each carousel and determined RMS DS at LOS, NLOS and partially obstructed paths. The Rx was also placed behind the wall at the top ramp. For three paths, i.e., behind carousels 2, 3 and behind the wall at the top of the ramp, RMS DS is maximum when the Rx is fully obstructed, as expected, illustrated in Figure 6.32. At LOS positions, we observed small RMS DS, close to 10 ns. The range of RMS DS values is the smallest behind carousel 2 and small PL values are also observed (Figure 6.6). Table 6.5 shows the RMS DS statistics considering all LOS, NLOS and partially obstructed locations for Tx1. A relatively large SD of RMS DS is observed for NLOS paths, 3.7 ns. The mean RMS DS of the partially obstructed path is approximately double that of the LOS paths. The minimum RMS DS for all paths is less than or close to 10 ns.
Figure 6.32 RMS DS for CAE baggage area Tx1, behind Carousel 1, 2, 3 and at the top of the ramp (LOS to NLOS transition).

Table 6.5 RMS DS for CAE baggage area Tx1 at LOS, NLOS and partially obstructed paths.

<table>
<thead>
<tr>
<th></th>
<th>LOS [ns]</th>
<th>NLOS [ns]</th>
<th>Partially Obstructed [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Min</td>
<td>5.21</td>
<td>10.1</td>
<td>10.26</td>
</tr>
<tr>
<td>Max</td>
<td>10.67</td>
<td>22.6</td>
<td>26.8</td>
</tr>
<tr>
<td>Mean</td>
<td>7.5</td>
<td>17.2</td>
<td>16.9</td>
</tr>
<tr>
<td>Standard Deviation</td>
<td>1.92</td>
<td>3.7</td>
<td>6</td>
</tr>
</tbody>
</table>

The CIRs were captured for Tx2 position near carousel 2 at CAE baggage area. The Rx was placed near the same carousel but on the opposite side of the Tx. Figure 6.33 illustrates RMS DS for Tx2. All LOS RMS DS values are less than 6 ns. We observe less
than 8 ns RMS DS in NLOS paths because the Rx could capture reflected signals from a nearby pillar. The Rx was partially obstructed by the pillar near carousel 2, and a large RMS DS of nearly 16 ns is observed.

![Figure 6.33 RMS DS versus link distance for CAE baggage area Tx2](image)

Figure 6.33 RMS DS versus link distance for CAE baggage area, near carousel 2, Tx2.

At the Tx3 position, we only changed the azimuth angle of the Tx antenna, keeping the Tx at the same position as Tx2, and the Tx antenna was aimed at one of the pillars, near carousel 1. The Rx was moved between carousels 1 and 2, and behind carousel 1. We observed a slightly larger RMS DS, more than 10 ns, in the LOS path near carousel 2, because the Rx could capture weak signals that were beyond the HPBW of the Tx. Similar observation is noticed for the partially obstructed paths. The Rx was partially obstructed by a metal pillar. The RMS DS in NLOS locations is smaller than in the LOS paths, since the Rx could obtain the strong reflected signal from the carousel 1 pillar. We also observed our expected small RMS DS behind carousel 1 in the LOS paths.
Figure 6.34 RMS DS for CAE baggage area, near carousel 1 and 2, Tx3.

We list the minimum, mean, maximum and standard deviation of RMS DS for all paths in Table 6.6 for Tx3. It is worth noting that the mean RMS DS for NLOS path is close to 10 ns. Both the Tx and the Rx were aimed at the same pillar near carousel 1 in this NLOS path and could get a strong reflected signal.

Table 6.6 RMS DS for CAE baggage area Tx3 at LOS, NLOS and partially obstructed paths.

<table>
<thead>
<tr>
<th></th>
<th>LOS [ns]</th>
<th>NLOS [ns]</th>
<th>Partially obstructed [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Min</td>
<td>6.14</td>
<td>9.7</td>
<td>9.7</td>
</tr>
<tr>
<td>Max</td>
<td>12.63</td>
<td>11.5</td>
<td>16.3</td>
</tr>
<tr>
<td>Mean</td>
<td>9.27</td>
<td>10.6</td>
<td>13</td>
</tr>
<tr>
<td>SD</td>
<td>2.9</td>
<td>0.92</td>
<td>2.9</td>
</tr>
</tbody>
</table>
6.4.3 Spatial PDP Correlation Coefficients

We computed sets of PDP correlation coefficients, via the method in [56], to assess the rate of spatial channel variation. The SPCCs were computed using our average PDP at a given location. For this case, they are large only on the diagonal. A threshold of SPCC was selected as 0.7 for our analysis, and stationarity distance was calculated if the SPCC is greater than or equal to 0.7. The following subsection describes the SPCC for both the airport maintenance hangar and the airport baggage area.

6.4.3.1 CUB Maintenance Hangar

The Rx was moved from LOS to NLOS paths (path 1 and 2) at CUB maintenance hangar. Channel impulse responses were captured at each distance and SPCCs were computed. It is shown in Figure 6.35 that the SPCC is greater than or equal to 0.7 when the Rx is at 6 m from the reference. The stationarity distance for NLOS is 0.5 m. We also noticed large SPCC values at 0.5 m, 1 m and 8.5 m from the reference in path 2, shown in Figure 6.36. The Tx and the Rx were in LOS in all cases, but the Rx captured MPCs with AoD beyond the Tx HPBW at 8.5 m. The maximum and minimum LOS stationarity distance is 1 m and 0.5 m, respectively. We determined larger stationarity distance in LOS than in NLOS settings, when we conducted measurements in U of SC indoor and street environments. However, in the airport maintenance hangar, the minimum stationarity distance for both LOS and NLOS are same, approximately 0.5 m. We did not observe a long LOS zone in the maintenance hangar as it was crowded with multiple objects, so we have less data to compute the SPCC here, which may at least partially explain the unexpected similarity of the LOS and NLOS SPCC values.
Figure 6.35 Spatial PDP correlation coefficient vs. distance along LOS to NLOS transition, path1, at CUB maintenance hangar.

Figure 6.36 Spatial PDP correlation coefficient vs. distance along LOS to NLOS transition, path2, at CUB maintenance hangar.
6.4.3.2 CAE Baggage Claim Area

The SPCC was also computed for LOS path at CAE baggage area for Tx1. When we rotated the Rx antenna 2 degrees anticlockwise from the reference antenna position, a large SPCC, close to 0.8, is observed at 25 m, and at 40 m link distance. It is also shown in Figure 6.37 that the average SPCCs at these two distances are approximately 0.6. The Rx was placed close to carousel 2 and carousel 1 in these distances. Although there are pillars in both sides of the LOS path, but we noticed more reflectors, e.g., more pillars, sign holders and carousels, at the left side of the Rx. For this reason, a large SPCC was observed when we rotated the Rx antenna anti-clockwise.

Figure 6.37 Spatial PDP correlation coefficient vs. distance for LOS Tx1, at CAE baggage area, clockwise from left: at reference, clockwise from reference, average and anti-clockwise from reference.
6.5 COMPARISON WITH RESULTS FROM THE LITERATURE

In this section, we make summary comparisons on the wireless channel characteristics in our aviation environments with results from other more commonly measured settings. We found very few existing articles in mmWave band channels for airport environments. Here we compare with a few indoor office environments.

In indoor office LOS environments, the authors of [64] found PL exponent as 1.3 to 1.7 from 26 to 38 GHz. In [61], the authors observed PL exponent in the passenger terminals of Seoul Railway Station and Incheon International Airport terminal for indoor hotspot: 1.8 (28 GHz) and 1.9 (38 GHz). At 60 GHz measurement, the authors of [42] determined PL exponent at the airport baggage area and the airport gate area as 1.12 and 1.38, respectively. In [43], the authors observed PL exponent as near FSPL exponent for small terminal building at 31 GHz. At 28 GHz, the authors of [49] observed PL exponent approximately equal to the FSPL exponent inside the terminal hall at Johnston Regional Airport (JNX), a local airport near Raleigh, NC. For our 90 GHz setup, we observed PL exponent as 2.3 at airport baggage area and approximately 2 at the airport maintenance hangar, slightly larger than all these other environments and frequency bands.

In obstructed LOS (OLOS) indoor office environments, the authors determined PL exponent as 1.8 to 2 from 26 to 38 GHz [64]. Authors of [43] observed PL exponent to be 3 for 31 GHz in an airport maintenance hangar environment for mixed paths. We observed PL exponent approximately 3 for both mixed and partially obstructed paths at 90 GHz in the airport maintenance hangar.

In [61] the authors measured median RMS DS in the passenger terminals of Seoul Railway Station and Incheon International Airport terminal (indoor hotspot) environment:
LOS– 42.5 and 32.6 ns at 28 and 38 GHz, respectively, and NLOS– 108.3 and 79.3 ns at 28 and 38 GHz, respectively. In our measurement, we observed smaller RMS DS for both LOS and NLOS settings. In an airport maintenance hangar environment, RMS DS is 4.5 to 7.6 ns for LOS settings, and RMS DS is 5.34 to 24 ns for NLOS settings. We also found small RMS DS at the airport baggage area: 4 to 20 ns and 10 to 27 ns, for LOS and NLOS settings, respectively. Our hypothesis for the smaller values of RMS-DS are that attenuation is larger at 90 GHz than in these other bands, generally enabling only shorter link distances, for which RMS-DS is generally smaller.
CHAPTER 7
CONCLUSIONS AND FUTURE WORK

7.1 DISSERTATION CONCLUSION

In this dissertation, we reported on measurements at 90 GHz for LOS, NLOS, and mixed paths, quantifying channel parameters for LOS-to-NLOS transitions. Measurements were conducted in various environments: non-aviation (indoor hallways and outdoor streets), and aviation (airport baggage claim area and airport maintenance hangar). We created path loss models, and also quantified several other important channel parameters: RMS DS, channel stationarity distance, AoA, and rates of change of attenuation. We also performed geometric analysis to validate our measured results, and compared measured results with those generated via ray tracing.

In indoor hallways and outdoor street environments, we conducted our LOS measurements up to a maximum link distance of 76 m. The LOS to NLOS transition measurements were made up to a maximum of 5.5 m from the LOS reference. We quantified PL, compared with WI ray tracing results, and computed additional channel characteristics: maximum received power and corresponding AoA, RMS DS, and the SPCC for LOS-to-NLOS transitions. We observed that the LOS PL exponent in all indoor and outdoor environments lies in the range 1.6 - 2.2. A comparison between measured and WI ray-tracing estimated PL was made using the CI PL model, for which we observed a maximum difference of 0.4 in slope and 1.46 dB in standard deviation. We observed
significant and rapid changes in channel characteristics in all environments, when we
moved the Rx from LOS to NLOS, e.g., an increase in average PL of ~13 dB for indoor
and ~17 dB for outdoor scenarios within several cm beyond the LOS zone, 1.5 dB/cm for
outdoor, and 0.8 dB/cm for indoor.

It is worth mentioning that the maximum received power decreases almost instantly
indoors when the Rx was moved from the LOS to the NLOS zone. In contrast, the
maximum received power maintained its initial LOS zone’s value for a larger distance
(maximum 56 cm from the reference) beyond the optical LOS point in outdoor settings.
An abrupt and substantial change in the arrival angle of maximum-power MPCs was also
observed. This is attributable to the various types of reflecting objects, i.e., low-loss, poorly
reflecting drywall and high-loss reflective metal indoors, and concrete and brick walls,
stairs, and metallic stair railings, poles, bike rack, cars, and the ground outdoors. During
the LOS-to-NLOS transitions, we observed a larger change in AoA in indoor NLOS areas
than in outdoor: greater than 60 degrees for indoor and greater than 30 degrees for outdoor.
Finally, we quantified RMS DS and channel stationarity distance in all scenarios: RMS DS
values were less than 30 ns, and minimum stationarity distance was as small as six
wavelengths (2 cm) for NLOS and six hundred wavelengths (2 m) for LOS.

We conducted measurements in the Jim Hamilton L. B. Owens Airport, in
Columbia, SC, to analyze the propagation of 90 GHz mmWave signals: we quantified PL
and power delay profile statistics. The change of maximum received power with antenna
azimuth angle was also analyzed for LOS to NLOS transition regions, by moving the Rx
behind an aircraft. We performed our measurements in a maintenance hangar, an
environment with many metallic obstacles including aircraft, as well as foil-insulation-
lined metal walls. We could not easily separate LOS from NLOS conditions on many paths due to the frequently-moved objects during our measurement. Hence, we reported our analysis for these mixed conditions. The hangar maximum LOS link distance was 24 m, and we conducted LOS to NLOS transition measurements up to 9 m from the reference. The measurement was also conducted within 15 m in a mixed environment. We observed that LOS PL exponent $n$ is near the free-space value of two, and SD is a few dB, and NLOS exponents are larger (~3), with larger SD (~6 dB) as well. It is worth noting that if we rotate the antenna in azimuth to obtain maximum received power, PL decreases as expected, and so does RMS DS (by approximately 25%) due to the surroundings of rich reflecting objects. The RMS DS in the mixed environment is less than 17 ns, whereas in LOS environments, it is less than 8 ns.

We observed that the received power dropped essentially instantly, as in indoor hallways, when we moved the Rx behind an aircraft in two receiver paths: ~17 dB in path 1 within 50 cm and ~14 dB in path 2 within 100 cm into the LOS to NLOS transition region. The Rx in path 2 captured more power than Rx in path 1 in this transition region because of reflected signals from the wall beside Rx path 2. We observed maximum 20 dB and 17 dB variation in the maximum received power between LOS and NLOS zones, in path 1 and path 2, respectively. Our AoA results confirm that the Rx captured maximum received power from multiple metal cabinets behind the Rx path 1 and the foil-lined wall or engines beside the Rx path 2. In LOS to NLOS transition settings for both paths, a maximum 24 ns RMS DS was obtained. The RMS DS in boresight is smaller than 6 ns, whereas RMS DS in LOS but beyond the Tx antenna HPBW is as large as 16 ns; such behavior is as expected. The stationarity distance for NLOS settings is 50 cm in Rx path 1.
In Rx path 2, the maximum and minimum stationary distances in the LOS setting are 100 cm and 50 cm, respectively.

In Columbia metropolitan airport, measurements were performed in a baggage claim area for different settings: LOS, NLOS, partially obstructed, LOS to NLOS transition. The maximum link distance of LOS and NLOS paths were 50 m and 75 m, respectively. We conducted measurements for three different Tx positions in the airport baggage area and performed LOS to NLOS transition measurement by moving the Rx behind the carousels and a wall. The PL exponent of the LOS path is 2.3, slightly larger than the free space PL exponent. As the Rx is surrounded by many reflectors, e.g., metal pillars, metal sign holders, carousels, etc., relatively large maximum RMS DS values were observed in both LOS and NLOS paths, 20.46 ns and 24 ns, respectively.

We make the following observations for this unique aviation environments:

- Long LOS link distances were not possible. So, mixed environments analysis is more realistic than only LOS, NLOS or obstructed paths analysis.

- In some cases, because of large metal objects, it is possible to obtain a fairly large received power in an NLOS zone, almost as large as in the LOS zone. Coverage range can be increased in NLOS zones because of this.

- As there exists several objects having strong reflection coefficients, RMS DS in NLOS zone can also be as small as in the LOS zone.

- The range of angle of arrival should be large as the Rx is surrounded by several potential reflectors.
As the environment is crowded by many obstacles, the number of MPCs and their strength vary significantly when the Rx is moved from one position to another, yielding generally small stationarity distance.

We include a summary of our aviation and non-aviation setting results in Table 7.1. It is observed in Table 7.1 that the range of AoA for aviation NLOS settings is approximately two times larger than found in the non-aviation NLOS settings. The Rx is surrounded by several reflectors with large reflection coefficients, e.g., walls, carousels, cabinets, engines, sign holders, pillars, etc., in aviation environments. Due to similar reasons, we observed a large RMS DS in LOS settings, i.e., approximately double in the aviation scenario compared to non-aviation scenarios. The maximum stationarity distance is larger in non-aviation environments than in CUB maintenance hangar, more than eight times in LOS settings.

Table 7.1 90 GHz channel characteristics parameters for both aviation and non-aviation environments.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Environments</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Aviation (Indoor)</td>
</tr>
<tr>
<td></td>
<td>Indoor</td>
</tr>
<tr>
<td>Maximum Link distance [m]</td>
<td>50</td>
</tr>
</tbody>
</table>
### 7.2 FUTURE WORK

We provide a few recommendations for work that could be conducted in the future:

1. The MIMO technique is one of the crucial techniques for 5G and beyond 5G communications. Data rates, spectrum efficiency, coverage, NLOS connectivity can all be increased by using MIMO. This technique also improves overall performance by reducing interference with its surrounding users. Hence,
measurements and simulation using MIMO antennas should be conducted in the 90 GHz band.

2. Exact values of material parameters, e.g., permittivity, surface roughness, conductivity, reflection coefficient, etc., accurate antenna pattern, and accurate dimensions of the floorplans and obstacles are vital for the analysis and modeling of mmWave channels using ray tracing simulation software. So, further analysis should be done for the improvement of simulation inputs.

3. The mobility of the Tx, the Rx and scatterers directly influences the behavior of the channel. The WSSUS properties should be investigated in pedestrian and vehicular environments for the 90 GHz band.

4. As the mmWave propagation system uses directional antennas, characterization of XPD of antenna system and radio channels is important for proper development of PL models. The received signals may not be exactly linearly polarized due to scattering effects, even though transmitted signals are linearly polarized. The XPD shows its dependencies on distance, azimuth, elevation and delay spread for different mmWave bands. XPD analysis should be investigated for both LOS and NLOS settings in 90 GHz band.

5. RIS may improve wireless communication performance when the Rx is fully or partially obstructed for both indoor and outdoor settings. An RIS can maximize spatial signal to interference plus noise ratio by decreasing interference. Aerial user communications are expected to be supported in B5G and 6G communications. Measurements and analysis using RIS on air ground communication near airports could be performed for the 90 GHz band.
6. Measurements and simulation should be performed for additional environments not fully characterized, e.g., factory settings, parks, airport surfaces, airport terminal buildings, etc.

7. Multi-band mmWave measurements can be conducted (e.g., 30 GHz and 90 GHz simultaneously) to enable design of future multiband mmWave systems.
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APPENDIX A

90 GHZ CHANNEL SOUNDER CALIBRATION

In this Appendix, we describe the RF signal spectrum at the Rx for different types of IF signals input at the Tx mixer: FZC, and chirp. Triggered and untriggered CIRs were captured by channel sounding software, and PDPs were computed for analyzing the difference between triggered and untriggered mode results. Our aim is to analyze the channel sounder behavior when we switch it to untriggered mode (for measurements) after beginning in triggered mode (for calibration). We also computed the calibrated system back to back response for untriggered mode. Finally, our 91 GHz channel sounder’s RF output at the SSA from 75.2 to 95.2 GHz was also examined.

To select the type of IF signal at the Tx mixer, we analyzed our sounder output at the SSA by using four different types of IF signals: chirp 16-bit, chirp 16-bit 10x, FZC 16-bit 10x, and FZC 16-bit 100x. All spectra are illustrated in Figures A.1-A.3. A large ripple, approximately 6 dB, is observed at the SSA if we used FZC 16-bit 100x as an IF input signal at the Tx mixer. Periodicities are also evident from the spectral lines, see Figure A.3. The ripple of all the other signal spectra is between 0.5 to 1.22 dB. In chapter 3, we described that the chirp 16-bit signal showed least ripple, i.e., 0.35 dB. Due to spectral flatness, the chirp 16-bit signal was used as an IF input signal at the Tx mixer in our 91 GHz channel sounder system.
Figure A.1 Measured sounder signal power spectrum, at SSA, for chirp 16-bit 10x IF input at the Tx mixer.

Figure A.2 Measured sounder signal power spectrum, at SSA, for FZC 16-bit 10x IF input at the Tx mixer.
Using the R&S channel sounding software, CIRs were captured for both triggered and untriggered modes. We used a -26 dBm level IF input signal at the Tx mixer, and an 80 dB attenuator was used between the Tx and the Rx. In untriggered mode, we captured CIRs after different periods of removal of both trigger and reference cables: immediately, 5, 10, 15, 20, 25, 30, 60, 1380 minutes. Here, we illustrate PDPs in Figure A.4, computed by CIRs, for both triggered and untriggered modes (removal of cables up to 30 minutes in each 5 minutes step). PDPs computed by “immediately after untriggered” CIRs, are distinctly different from the triggered PDPs. It is evident in this Figure that the untriggered PDPs follow the triggered PDPs after approximately 15 minutes of operation in the untriggered mode. Hence we must wait at least 15 minutes to start our untriggered measurement.
The RMS DS for triggered and untriggered modes, for different time periods, i.e., immediately, 30 minutes, 60 minutes, and 23 hours, were computed from measured PDPs. We determined RMS DS using different thresholds: removing samples after 50, 30 and 25 dB from the peaks, taking samples within a delay interval, and a delay window. A few examples are illustrated in Figures A.5 and A.6. For all settings, a large RMS DS is observed immediately after untriggered mode, maximum approximately 36% larger than in triggered mode (Figure A.6). It is also observed in Figure A.5 that RMS DS does not vary significantly, i.e., 1.7% from triggered mode, after 23 hours of operation in the untriggered mode; this implies that after calibration in triggered mode, we can conduct
measurements from approximately 15 minutes after calibration to 23 hours after calibration.

Figure A.5 RMS DS for triggered and untriggered modes removing samples 30 dB from peak.

Figure A.6 DS for triggered and untriggered modes considering 90% delay Window.
The calibrated system back to back response was computed for different time periods in untriggered mode: immediately, 5, 10, 15, 20, 25, and 30 minutes after removal of the trigger and reference cables. Figure A.7 illustrates this result. We observed in chapter 3 that our calibrated system back to back triggered response show a near Dirac delta. After 15 minutes of untriggered mode, the calibrated system back to back response shows an approximate Dirac delta, but with a wider pulse width.

Figure A.7 Calibrated system back to back response for untriggered modes.
The RMS DS for the untriggered calibrated system back to back response was computed for different time periods of untriggered mode, from 1 minute to 30 minutes in each 5-minute interval. The range and standard deviation of RMS DS for different time periods are 0.4 ns and 0.15 ns, respectively. This fraction of a nanosecond variation does not appreciably affect our channel DS values (with minima ~ 6 ns for some LOS cases).

Figure A.8 RMS DS for triggered and untriggered modes (calibrated) removing samples 30 dB from peak.

The transmitted signal spectrum was observed at the SSA from 75.2 to 95.2 GHz, when we transmitted our 91 GHz signal. We observed approximately similar signals at 91 GHz and at 94 GHz center frequencies. Spectrum at 91 GHz is our desired signal because 11 GHz LO is multiplied to 88 GHz, and then an 88 GHz signal is mixed with a 3 GHz IF signal. Spurious emissions are also observed at other few frequencies, e.g., 75.3, 81, 83.3,
88, 90.3, 91.7, 92.9, 94.4 GHz. The 94 GHz spurious signal is evidently the result of non-ideal mixing and multipliers, but this is fully suppressed in our receiver processing.

Figure A.9 Observed spectrum in the SSA within 75.2 to 95.2 GHz band by transmitting 91 GHz signal.
APPENDIX B

TWO RAY MEASUREMENT CALIBRATION

In this Appendix, we describe our two ray measurement calibration, and corresponding results. For this experiment, we set up a simple channel that can be analyzed theoretically, to compare with our measured results. As in all measurements, CIRs were captured by the channel sounding software and PDPs were computed. We explain our theoretical analysis, and compare these results with our computed PDPs.

Figure B.1 shows the plan view diagram of the measurement. We performed our measurements on the 3rd floor balcony (courtyard) in Swearingen Engineering Center, Columbia, SC. This is a half-circle shaped balcony with a radius of 21.6 m. We placed our Tx and Rx near balcony railing at 1.37 m height from the floor. On the opposite side of the railing, i.e., diameter of the half circle, there are glass windows with metal frames.
Our target is to observe primarily two rays in PDPs: one direct ray and another reflected ray from windows. We used -26 dBm chirp signal as an IF input at Tx mixer in our 90 GHz channel sounder. The distance between Tx and Rx was 37.4 m. According to the geometry, the difference between direct and reflected path is 5.8 m. The difference in delay between the direct and reflected rays should be 19.3 ns. Figure B.2 illustrates PDPs of two ray measurements. We observed in the Figure that there is a 20 ns difference in delay between direct and reflected rays, in good agreement with the estimated 19.3 ns. Both Tx and Rx were boresight aligned in direct path.

Figure B.2 PDPs for two ray measurements, direct ray and reflected ray.

The antenna pattern (for both antennas) is illustrated in Figure B.3. The ray gain should be 30 dB for the direct path. In the reflected path, the ray gain should be 5 dB at 30°
AoA and AoD. However, our Tx antenna is connected to a mounting structure that produces an approximate $3^\circ$ offset (counterclockwise) from the LOS path (dotted green line in Figure B.1). Hence the AoD of the transmitted ray is $27^\circ$ instead of $30^\circ$. So, the gain of the reflected path should be $\sim 9.5$ dB. The difference in FSPL between the direct and the reflected path is 1.2 dB at 91 GHz. If we consider metal frame reflection, then the reflected ray should be $21.7$ dB ($30-9.5+1.2$) down from that of the direct ray, otherwise $\sim 23.6$ dB down if the reflection is considered from glass window. Here, we consider the reflection coefficient of glass as 0.8. Our measured value of approximately 22.2 dB is within this range, confirming analysis with measurements for this 2-ray channel.

Figure B.3 Tx and Rx antenna pattern [53].
APPENDIX C

AIRPORT BAGGAGE CLAIM AREA POWER DELAY PROFILES

In this appendix, we illustrate additional example PDPs for the CAE baggage area, for three Tx positions: Tx1, Tx2 and Tx3. For Tx1, measurements were taken in LOS, at the top of the ramp (LOS to NLOS transition), and behind three carousels (LOS, NLOS and partially obstructed locations). We also performed measurements near baggage area 2 (LOS, NLOS, and partially obstructed paths) for Tx2. Our last measurements were done near baggage area 2 and behind carousel 1 (LOS, NLOS and partially obstructed settings) for Tx3. In general, a few significant MPCs were observed in LOS paths, and weak MPCs were observed at large link distances. Figure C.1 shows a sequence of PDPs for a LOS path in the CAE baggage area, Tx position 1,

Figure C.1 Sequence of PDPs for LOS path at CAE Baggage area for Tx1.
Figures C.2 shows example PDPs at the top of the ramp area for Tx1. It is noticed that the Rx received strong reflected signal although it is in NLOS zone, at distances of 1.6 m and 2.2 m from the reference. Hypothesized reflections are from a metal pillar near baggage area 1. Measurements were performed behind each of the three carousels, and PDPs were computed for baggage areas 1, 2 and 3, illustrated in Figures C.3, C.4 and C.5, respectively. Strong MPCs are often observed in NLOS settings, behind a pillar, for the best (strongest received power) antenna position. In NLOS paths, the Rx often received MPCs from nearby reflectors. The signals were partially obstructed by sign holders in semi-obstructed path. The link distance of LOS, NLOS, LOS and partially obstructed settings were 15 m, 16.6 m, 17.5 m and 19.7 m for baggage area 3, see Figure C.5.

Figure C.2 Sequence of PDPs for LOS to NLOS transition path at the top of the ramp.
Figure C.3 Example PDPs in dB, in CAE Baggage area 1 for Tx1.

Figure C.4 Example PDPs in linear scale, in CAE Baggage area 2 for Tx1.
Power delay profiles were also computed for two transmitter positions, Tx2 and Tx3. The antenna of the Tx2 was aimed parallel to the carousel 2 whereas the Tx3 antenna was pointed toward one of the pillars near carousel 1. Figures C.6 and C.7 show the PDPs near baggage area 2 for Tx2 and Tx3, respectively. The last Figure C.8 show PDPs near baggage area 1 for Tx3. Strong MPCs are observed in all LOS settings. We also observe dominant MPCs in NLOS settings as the Rx caught reflected signals from potential reflectors. When the Rx was partially obstructed by pillars or sign holders, we observe both direct and reflected MPCs.
Figure C.6 Example PDPs in dB, near baggage area 2 for Tx2.

Figure C.7 Example PDPs in linear scale, near baggage area 2 for Tx3.
Figure C.8 Example PDPs in linear scale, near baggage area 1 for Tx3.