THE IMPACT OF ANTENNA BEAMWIDTH ON THE CORRELATION DISTANCE
OF 60 GHZ INDOOR AND OUTDOOR CHANNELS

by

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B.A.Sc., The University of British Columbia, 2018

A THESIS SUBMITTED IN PARTIAL FULFILLMENT OF
THE REQUIREMENTS FOR THE DEGREE OF
MASTER OF APPLIED SCIENCE

in
THE FACULTY OF GRADUATE AND POSTDOCTORAL STUDIES
(Electrical & Computer Engineering)

THE UNIVERSITY OF BRITISH COLUMBIA
(Vancouver)

April 2021

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The following individuals certify that they have read, and recommend to the Faculty of Graduate
and Postdoctoral Studies for acceptance, a thesis entitled:

THE IMPACT OF ANTENNA BEAMWIDTH ON THE CORRELATION DISTANCE
OF 60 GHZ INDOOR AND OUTDOOR CHANNELS

submitted by Aidan Hughes in partial fulfillment of the requirements for

the degree of Master of Applied Science

in Electrical and Computer Engineering

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Abstract

Due to narrow antenna beamwidth and channel sparseness, millimetre-wave receivers will detect far fewer multipath components than their microwave counterparts, fundamentally changing the small-scale fading properties. By corollary, the de facto Rayleigh-Rice model, which assumes a rich multipath environment interpreted by the Clarke-Jakes omnidirectional ring of scatterers, does not provide an accurate description of this fading nor of the correlation distance that it predicts. Rather, a model interpreted by a directional ring of scatterers, recently proposed in seminal work by Va et al., theoretically demonstrated a strong dependence of correlation distance on beamwidth. To support Va’s model through actual measurement, we conducted an exhaustive measurement campaign in five different environments – three indoor and two outdoor – with our 60 GHz 3D double-directional channel sounder, compiling over 36,000 channel captures. By exploiting the super-resolution capabilities of the channel sounder, we were the first, to our knowledge, to measure correlation distance as a function of continuous beamwidth. We showed that for narrow beamwidth, correlation was maintained for much longer distances than predicted by the Rayleigh-Rice model, validating Va’s model. As the beamwidth approached omnidirectionality, with increasing number of multipath detected, the behavior indeed approached the Rayleigh-Rice model. We also demonstrated how virtual arrays implemented using a robotic arm can realize variable beamwidth antenna apertures at a fraction of the cost of the multi-element array used in this study, albeit with some restrictions on the types of the channels that can be characterized, and revealed that the relationship between estimated and actual rms delay spread is simply related to the dynamic range of the measurement.
Lay Summary

Although the use of the millimetre-wave bands for personal communications promises to unlock vast amounts of wireless bandwidth for both consumer and commercial use, the distortions and impairments introduced by over-the-air transmission of millimetre-wave signals remain a significant challenge for designers. Channel models capture our knowledge and intuition regarding the nature of such channel impairments in a form that permits designers to assess the impact of such impairments on system performance and devise potential solutions. This thesis presents the results of a measurement and modeling study involving 60 GHz indoor and outdoor channels which assessed the impact of antenna beamwidth on factors that affect the performance of multi-antenna receivers. The results show how millimetre-wave channels differ from those at lower frequencies and provide important insight to system designers.
Preface

A version of the material in Chapters 1, 2, 4 and 5, collectively, was published as A. Hughes et al., "Measuring the impact of beamwidth on the correlation distance of 60 GHz indoor and outdoor channels," in IEEE Open Journal of Vehicular Technology, doi: 10.1109/OJVT.2021.3067673. I conducted all the data analysis and wrote the majority of the initial versions of the manuscript under the supervision and guidance of Camillo Gentile while I was an international guest researcher at the Wireless Networks Division at NIST, Gaithersburg, MD. Derek Caudill conducted the measurement campaign detailed in the manuscript at NIST, Boulder, CO, campus. Jack Chuang implemented the SAGE algorithm used in the data processing pipeline for my research and assisted with various other technical tasks in the process. Sung Yun Jun, Camillo Gentile and Jelena Senic helped extensively to edit and write chapters 1, 2, 4 and 5. David Michelson participated in on-site discussions at NIST and assisted greatly with interpretation of the results and preparation of the final version of the manuscript.

Chapter 3 is based on the SAMURAI system demonstrated by Alec Weiss, Jeanne Quimby, Rod Leonhardt, Ben Jamroz, Dylan Williams. Kate Remley. Peter Vouras, and Atef Elsherbeni at NIST. I worked on the design and implementation of the system in the Radio Science Lab under the supervision of David Michelson.

Chapter 6 is based on an abstract that has been accepted for presentation at the 2021 General Assembly and Scientific Symposium (GASS) of the International Union of Radio Science (Union Radio Scientifique Internationale - URSI) and was prepared under the supervision of David Michelson.
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Acknowledgements

I offer my enduring gratitude to the faculty, staff and my fellow students at UBC for guiding me through a wide range of knowledge in the field of electrical engineering. I owe exceptional thanks to Prof. David G. Michelson for introducing me to his long-time research collaborators at the U.S. National Institute of Standards and Technology (NIST), where he also served as an international guest researcher in 2018, and opening up many incredible opportunities to contribute to the field of wireless propagation, and which I have learned from immensely. His on-site visits while I was at NIST and assistance in interpreting the results of my work were invaluable.

I am incredibly grateful for the channel modelling group in the Wireless Networks Division at NIST Gaithersburg, specifically Nada T. Golmie, Camillo Gentile, Jack Chuang and Chieping Lai, and the measurement team at NIST Boulder, for all of the support and powerful tools they provided me with for my research. I give particular thanks to Dr. Camillo Gentile for the continuous support and guidance during my time at NIST.

I give the highest praise to my parents, who have supported me in every way imaginable through my education and are the greatest role models one could have.
Dedication

This thesis is dedicated to my parents Ken and Denise, and my siblings Eamon and Carlan. I love you all.
Chapter 1: Introduction

5G millimetre-wave (mmWave) wireless channels will experience much greater path loss than sub-6 GHz channels due to the decrease in receiving antenna size, the increase in diffraction loss, and the increase in signal attenuation by vegetation and building materials at millimetre-wave frequencies [1], [2]. To compensate for the higher loss, high-gain directional antennas will be employed. Since gain is inversely related to beamwidth, beamwidths will vary from several degrees (pencil beams) to tens of degrees depending on the link budget and precise centre frequency, which alone will have impacts on various channel parameters [3][4][5]. When in motion, the transmitter (Tx) and receiver (Rx) beams will misalign quicker with narrower beamwidth, causing the received signal to decorrelate faster, translating to shorter correlation distance [6]. This in turn will necessitate a faster refresh rate for beamforming training [7], the spatial channel estimation process in which the beams are realigned. This is a very computationally costly process at ultrawide bandwidth (> 2 GHz) of 5G systems [3] and is associated with least-square minimization and convex optimization methods [8][9][10][11].

Correlation distance is commonly defined as the displacement over which the channel’s autocorrelation function (ACF) drops below 50% of its initial value [12]. The ACF is derived from measurements or models for small-scale (fast) fading, i.e., the fluctuations in the received signal that occurs over displacement on the order of wavelengths due to multipath interference [13]. The de facto model for fading is Rayleigh, in which the in-phase and quadrature components of the signal are represented as i.i.d. zero-mean Gaussian random variables, sustained by the narrowband assumption for which individual paths are irresolvable (smeared). This corresponds to a rich
multipath environment, interpreted geometrically by the Clarke-Jakes ring of scatterers [14], where the mobile receiver is surrounded by an infinite number of equidistant scatterers, hence have comparable strength when sensed by an omnidirectional antenna. The uniform distribution in angle-of-arrival (AoA) of the scattered paths gives rise to the classical U-shaped Doppler power spectrum [12]. The Rayleigh-Rice model is a popular variant in which a dominant scatterer, whose relative strength to other scatterers is quantified by the K-factor, is also present. The assumptions implicit to the Rayleigh-Rice model (and similar variants such as Nakagami [15], Suzuki [16], and Loo [17]) are valid for narrowband, omnidirectional systems, and thus form the cornerstone for recent channel models [18][19][20] to support the design of 4G LTE systems operating at sub-6 GHz.

The high directionality of 5G mmWave systems, on the other hand, results in the detection of only a few scatterers by the receiver. The few scatterers will in turn have a non-uniform AoA distribution [21][22][23][24], giving rise to an asymmetrical Doppler spectrum [25]. Furthermore, negligible diffraction at mmWave [26] makes for a sparse channel [27][28], leading to even fewer paths detected [21]. Consequently, the paths detected will exhibit less smearing, not just because there are few, but because the ultrawide bandwidth of mmWave enables resolution of individual paths, translating to highly correlated in-phase and quadrature components. By corollary, the assumptions implicit to Rayleigh-Rice fading do not apply to mmWave channels [29]. Popular channel models designed specifically for 5G systems [30][31] still adhere to these assumptions.

Measurements collected at mmWave concluded that the Two-Wave with Diffuse Power (TWDP) [32] and Fluctuating Two Ray (FTR) [33] fading models, in which the channel is dominated by two paths comparable in strength together with diffuse scattering, were more
suitable than Rayleigh. In [28][34][35], Pätzold theoretically investigated the correlation distance of mmWave channels, showing a significant variation across realizations with 2-10 non-isotropic scatterers. Beyond Pätzold’s theoretical work, there have been just a handful of studies on correlation distance at mmWave, and even fewer measurement-based work [2][21][36][37][38].

The 30 GHz measurements by Iqbal et al. [2][21] employed horn antennas with 30° beamwidth mechanically steered towards select scatterers in a small lecture room. For very narrow beamwidth ($\omega < 0.3^\circ$), the beam steered towards the AoA ($\theta_n$) of different ambient scatterers became quickly misaligned as the receiver moved, causing the coherence to fall short of its peak. Widening the beam improved alignment, reaching the coherence peak around $\omega = 0.5^\circ$, but coherence dropped off thereafter due to more and more local scatterers being admitted into the beam. The work was of notable impact because in contrast to [36][37][38], the complex form of the ACF – with no spatial averaging, neither across scenarios nor antenna elements – was computed. In addition, they demonstrated that the channels had high correlation between the in-phase and quadrature components, what is more representative of mmWave scattering conditions.

Although Iqbal’s work was indeed of notable impact, the results were specific to the 30° beamwidth of the horns employed for measurement, which is wider than beamwidths expected for mmWave [39]; in contrast, the 60 GHz phased-array antennas in [40] only had 3° beamwidths. For this reason, Va et al. [24][39] investigated the impact of beamwidth, and in particular beam misalignment, on the correlation distance of vehicular mmWave channels. In their theoretical model, the receiver in the Clarke-Jakes ring of scatterers was modified to have a synthetic horn with variable beamwidth. From the model, correlation distance was computed as a function of beamwidth.
Figure 1: Correlation distance vs. antenna beamwidth. (a) Results from Va et al.’s theoretical model [39]. For the sake of consistency, the abscissa was converted from coherence time in the publication to coherence distance by assuming a speed of $v = 3.6 \text{ km/h}$. (b) Results from actual measurements in our Laboratory for steering a beam towards select ambient scatterers, exhibiting the same behavior as Va’s model.
See the illustrative example in Figure 1(a): For very narrow beamwidth ($\omega < 0.3^\circ$), the beam steered towards the AoA ($\theta_n$) of different ambient scatterers became quickly misaligned as the receiver moved, causing the correlation to fall short of its peak. Widening the beam improved alignment, reaching the correlation peak around $\omega = 0.5^\circ$, but the correlation dropped off thereafter due to more and more local scatterers being admitted into the beam. Va’s theoretical model highlighted an important characteristic of directional mobile channels but was not supported by actual measurements.

In this thesis, we bridged the gap between Iqbal’s work, which experimentally measured the ACF/correlation distance for a fixed beamwidth, and Va’s work which modeled it theoretically for variable beamwidth. High precision measurements were taken with NIST’s state-of-the-art 60 GHz switched-array channel sounder, which estimated path gain, delay, and 3D double-directional angle, *i.e.*, azimuth and elevation angle-of-departure (AoD) from the transmitter and AoA to the receiver, of channel paths with average errors of 1.95 dB, 0.45 ns, and 2.24°, respectively. An extensive measurement campaign was run in five different environments – three indoor and two outdoor – comprising a total of 20 scenarios, each with 1801 small-scale acquisitions, for a total of 36,020 channel captures, enabling a comprehensive evaluation of correlation distance. With these measurements, to the best of our knowledge, we were the first to measure correlation distance as a function of continuous beamwidth using a single system by combining super-resolution techniques to extract paths with a synthetic horn with variable beamwidth.
In addition to using the 60-GHz channel sounder operated by NIST, the implementation of a channel sounder based on a virtual array with a maximum operating frequency of 26.5 GHz is demonstrated at UBC. The virtual array’s geometry and element spacing is highly configurable, which allows for different techniques to be used for achieving a variable beamwidth rather than those used with NIST’s 60 GHz channel sounder. The virtual array channel sounder is compared to another front-end design which operates from 40 GHz to 220 GHz and has full polarization capability, but with fixed antennas. The benefits and drawbacks of each design are considered.

The delay spread of a channel, which is a measure of the time dispersion a signal will experience, is an important parameter for channel modelling and is impacted when a minimum threshold is put on measurement data to ensure the noise floor is excluded. This threshold typically ranges from 15-25 dB above the noise floor and effectively reduces the system’s dynamic range. The fundamental relationship between root-mean-square (rms) delay spread and the dynamic range of a measurement is shown by introducing the concept of an intrinsic delay spread and demonstrated theoretically.

The thesis develops as follows: Chapter 2 describes our channel sounder and measurements. Chapter 3 describes the use of a virtual array to achieve a variable beamwidth antenna. In Chapter 4, the main scatterers per environment were identified and for each, the ACF was computed by steering a synthetic horn towards the scatterer. In Chapter 5, the correlation distance per scatterer as a function of continuous beamwidth was computed from the ACF, and representative metrics of the resultant curves were compiled over the 20 scenarios. In Chapter 6, the concept of intrinsic delay spread and the effect of restricting dynamic range are considered. In Chapter 7, conclusions and recommendations are presented.
Chapter 2: Channel Measurements

This chapter describes our channel measurements. First, our channel sounder is presented, followed by details of our measurement campaign, then by processing techniques implemented to extract channel paths and their properties from the measurements.

2.1 Channel Sounder

Figure 2 displays the 60 GHz 3D double-directional switched-array channel sounder [41]. The receiver features a circular array of 16 horn antennas with 22.5° beamwidth, rendering a synthesized azimuth field-of-view (FoV) of 360° and 45° in elevation. The transmitter was almost identical except that it featured a semicircular array of only 8 horns, limiting the azimuth FoV to 180°. At the transmitter, an arbitrary waveform generator produced a repeating M-ary pseudorandom (PN) codeword, with 2047 chips of duration $T_c = 0.5$ ns (or equivalent delay resolution), corresponding to bandwidth $B = 1/T_c = 2$ GHz. The codeword was produced at IF, upconverted to 60.5 GHz, and then emitted by a horn antenna. At the receiver, the signal is received by a horn antenna, downconverted back to IF and then sampled at 40 Gsamples/s. Finally, the sampled signal was match filtered with the codeword to generate a complex-valued channel
impulse response (CIR) as a function of delay. To reduce the effect of noise, ten CIRs are taken and averaged together. An optical cable between the transmitter and receiver was used for synchronous triggering and phase coherence [42]. For a transmit power of 20 dBm, the maximum measurable path loss of the system was 162.2 dB when factoring in antenna gain, processing gain, system noise, and remaining components of the link budget.

The codeword was electronically switched through each pair of transmitter and receiver horns in sequence, resulting in $16 \times 8 = 128$ CIRs, which is referred to as an acquisition. The Tx and Rx antennas are identified by clockwise numbering. Within a single acquisition, the individual CIRs between a pair of Tx/Rx antennas are organized by channel index, as depicted in Figure 3.

Figure 2: Photograph of the Tx array and Rx array of our 60 GHz 3D double-directional switched-array channel sounder collecting measurements in the Laboratory environment. Some of the main scatterers identified in the environment are labeled.
2.2 Measurement Campaign

Locations for the measurement campaign carried out in August 2019 were selected around NIST’s Boulder, Colorado campus, where the 60 GHz channel sounder is located. Locations were chosen such that both indoor and outdoor scenarios were investigated. Static channel conditions were imposed for the whole duration of measurements by closing indoor locations to pedestrians and outdoor locations were chosen away from vehicular traffic.

Five different environments were selected: three indoor (Laboratory, Lobby, Lecture Room) and two outdoor (Pathway, Courtyard). The laboratory environment included many metallic objects of varying size and geometries, which has similarities to an industrial case. The Lobby and Lecture Room represent typical residential or office cases, where scatterers are furniture such as chairs and tables, and flat surfaces such as televisions or drawing boards. The Pathway environment was a wide-open area approximately 30 m x 90 m in size, enclosed on three sides by
buildings 2-3 stories tall, which were expected to be the main contributors of specular scattering in the channel. Apart from the buildings, possible scatterers around the Pathway were metal lampposts and a group of large trees on the north side. There was no concern of scattering from vehicles since the Pathway was only foot-traffic accessible. The Courtyard location was chosen to represent an outdoor scenario that is more populated by scatterers than the Pathway. The Courtyard location was directly beside a building doorway with outdoor tables and above-ground planters with light vegetation nearby.

In each of the environments, four scenarios with two different Tx locations and two different Rx locations were investigated, for a total of 20 scenarios altogether. LoS conditions were maintained throughout. The Rx was mounted on a fixed tripod at 1.6 m height; the Tx was also mounted at 1.6 m on a 90 cm rail (linear positioner) whose translation was parallel to the ground. The positional tolerance of the rail was 76 μm. The measurement per scenario consisted of 1801 channel acquisitions as the Tx was translated, equivalent to a small-scale displacement of \( \frac{\lambda}{10} = 0.05 \text{ cm} \) between each acquisition, indexed as \( d = 0, 0.05, \ldots, 90 \text{ cm} \). Note that the granularity recommended in [36] to measure small-scale fading was 4-5 samples per wavelength. It required about 30 minutes to capture all 1801 acquisitions for a given scenario.

### 2.3 Path Extraction

The following steps were used to automate the extraction of all paths subject to the birth and death process [18][19][20], delimited by discrete birth and death displacements, denoted as \( d_{n}^{BIR} \) and \( d_{n}^{DTH} \), respectively:
1) **Identify all peaks in the CIRs which are assumed to be unique paths in the channel.** Peaks that are above a threshold of -55 dBm are identified, beginning with the highest amplitude peak appearing in CIR with channel index \( n \), \( CIR_n \). Time-delayed, attenuated copies of this path will be received in the channels of the neighboring Rx antennas, \( CIR_{n \pm 1} \). To ensure paths are not counted twice, a 2ns window in delay is zeroed out in \( CIR_{n \pm 1} \) at the delay of the identified peak. This process repeats until all peaks above the threshold have been identified for all 1801 rail displacements.

2) **Track peaks that appear continuously across rail displacements.** Paths extracted were tracked along the rail using a technique based on the Assignment Problem [43], yielding tracks of corresponding paths across multiple consecutive displacements. The smallest and largest displacements where track \( n \) appears are recorded as \( d_n^{BIR} \) and \( d_n^{DTH} \), respectively.

3) **Based on assumptions about the scattering surface geometry, combine tracks that may have been interrupted by blockage or fading but are still likely from the same scatterer in the environment.** A secondary track-linking algorithm was used to improve path tracking. With a given set of N tracks, the process begins with track \( T_n \) that has the lowest \( d_n^{BIR} \). If multiple tracks have the same \( d_n^{BIR} \), the track with the lowest delay value occurring at \( d_n^{BIR} \) is chosen as \( T_n \). The algorithm is as follows:

   a) Fit linear regression to delay points of trace \( T_n \).
   
   b) Extrapolate line forward from \( d_n^{BIR} \) to remaining rail displacements.
   
   c) Define delay search window of ±125 ps (at 45 GHz sample rate, 125 ps is 5 sample points, corresponding spatially to ±3.75 cm).
d) Search through $d_{m}^{\text{search}} = \{d_{1}^{\text{search}} = d_{n}^{\text{BIR}},...,d_{M}^{\text{search}} = 90\text{ cm}\}$ for next track $T_{n+1}$ such that $d_{n+1}^{\text{BIR}} = d_{m}^{\text{search}}$ within the delay window.

e) Append points in $T_{n+1}$ to the end of $T_{n}$.

f) Apply steps a) through e) with $T_{n+1}$ recursively until $d_{M}^{\text{search}} = 90\text{ cm}$ is reached.

g) Update total number of tracks in set, $N_{update}$.

h) Iterate steps a) through g) for the full set of traces until the total number of traces $N_{update}$ does not change from one iteration to the next.

4) Input all tracked paths to the SAGE super-resolution algorithm to improve estimates of the path properties [44][45]. The output from the SAGE algorithm for an acquisition at displacement $d$ was $N$ channel paths indexed through $n$, together with the path properties in the six-dimensional space: complex amplitude $\alpha_{n}(d)$, delay $\tau_{n}(d)$, and 3D double-directional angle $\theta_{n} = [\theta_{n}^{T}(d) \theta_{n}^{R}(d)]$, where $\theta_{n}^{T}(d) = [\theta_{n}^{T,A}(d) \theta_{n}^{T,E}(d)]$ denoted the AoD from the Tx in azimuth (A) and elevation (E) and $\theta_{n}^{R}(d) = [\theta_{n}^{R,A}(d) \theta_{n}^{R,E}(d)]$ denoted the AoA to the Rx. SAGE is very computationally costly, so points appearing at every $10^{th}$ slider position (0.5 cm) indoors and every $20^{th}$ position (1 cm) outdoors are input to SAGE and the six output properties are linearly interpolated.
The measurement error of the channel sounder was computed against the ground-truth properties of the LoS path, whose delay and AoA were given from the geometry between the Tx and Rx and its path gain from the delay through the Friis transmission equation. The measurement error averaged over all displacements and scenarios was determined to be 1.95 dB in path gain, 0.45 ns in delay, and 2.24° over all four angle dimensions.

Figure 4: Steps a) through e) repeated recursively on an example set of tracks \( \{T_1, T_2, T_3\} \).
Any measurement taken with a channel sounder will contain not only the response of the channel, but also the response of the sounder itself, \textit{i.e.}, the directional patterns of the antennas and the responses of the Tx and Rx front ends. While the SAGE algorithm was able to de-embed the antenna patterns, the effects of the Tx and Rx front ends were removed through pre-distortion filters from a back-to-back calibration method [41], hence the path properties reflected the “pure” response of the channel alone and not that of the measurement system.

The \textit{spatial CIR}, which incorporated the angle dimension (in addition to the delay dimension of the CIRs), offered a compact representation of the channel per scenario and is written as

\begin{equation}
    h(d, \theta, \tau) = \sum_{n=1}^{N} \alpha_n(d) \cdot p(\tau - \tau_n(d)) \cdot \delta(\theta - \theta_n(d)),
\end{equation}

where $p(\tau)$ denotes the transmitted pulse from a unity-gain omnidirectional antenna after matched filtering. It is equivalent to what a receiver (also with a unity-gain omnidirectional antenna) would detect at $d$, \textit{i.e.}, $N$ copies of the transmitted pulse, each corresponding to a different path $n$, scaled by complex amplitude $\alpha_n(d)$ and arriving with delay $\tau_n(d)$ from angle $\theta_n(d)$. As an example, Figure 5(a) displays the channel paths extracted at the first displacement ($d = 0$ cm) in a Laboratory scenario.
Angle-of-arrival, $\theta_{nR,A}$ (deg)
Delay, $\tau_n$ (ns)
Path gain, $\alpha_n^2$

(a)

Displacement, $d$ (cm)

(b)
Figure 5: Channel paths extracted in a Laboratory scenario. (a) All paths extracted at the first displacement (d = 0 cm) on the rail, displayed in azimuth AoA, delay, and path gain. (b) Persistent paths tracked over all displacements (d = 0...90 cm) displayed in azimuth AoA and path gain, labeled with the source scatterer in the environment. (c) Delay and path gain vs. displacement for persistent paths.
Chapter 3: A Virtual Array for Millimetre-Wave Channel Sounding

3.1 Introduction

The switched array channel sounder at NIST allows various beam patterns to be synthesized using a single set of measurements, but it may be prohibitively expensive for other institutions pursuing similar research. This chapter demonstrates an alternative method for variable beamwidth channel sounding by using a virtual array. The following design presents a viable alternative technique for reproducing the measurements performed with the switched array channel sounder in Chapters 2, 4 and 5.

Virtual arrays have several advantages over a conventional, fixed antenna setup in a channel modelling context: the number of actual antennas required is reduced while still benefitting from the gain that is provided by MIMO systems and element coupling is not a concern. For MIMO channel modelling, and element spacing can be changed easily with no additional hardware. The main disadvantage is that non-stationary channels are challenging to characterize as there is an unavoidable delay between measurements while the physical antenna is moved between virtual element positions, but design tradeoffs can be considered with IF bandwidth, number of points and averaging options in the VNA setup to reduce the per-element measurement time.

The Metrology for Wireless Systems Group at NIST developed the Synthetic Aperture Measurements of Uncertainty in Angle of Incidence (SAMURAI) system in 2019 [46] which enables highly flexible virtual array measurements with configurable array geometry. The SAMURAI system was replicated in the Radio Science Lab at UBC with similar hardware design and using the same software package uploaded to Gitlab by the Metrology group.
This chapter proceeds as follows: Section 3.2 describes the virtual array concepts relevant to the SAMURAI system. Section 3.3 presents the implementation of SAMURAI at the Radio Science Lab. Section 3.4 presents a basic demonstration of the system’s functionality with a linear virtual array measurement. In Section 3.5, the design of a different channel sounder front-end for the Radio Science Lab is presented that covers a range of capabilities distinct from SAMURAI’s, and the advantages and disadvantages of each design is examined. Section 3.6 is a brief discussion of the findings covered in this chapter.

3.2 Concept

A virtual array is based upon the same concepts that govern a conventional phased antenna array. A set of antennas are placed with a consistent spacing between elements, typically measured in fractions of a wavelength. Depending on the spacing and signal wavelength, the signal received or transmitted creates an interference pattern in the far-field, resulting in an angular pattern that produces an array gain when adding constructively and nulls when combining destructively. The signal at each antenna is controlled coherently using phase shifters or time delay circuits to shape the far-field interference pattern towards a given direction. Varying the amplitude of element feeds can also be used to shape the beam pattern, for example tapering the feeds at the edges of an array to reduce sidelobes. The received signal is assumed to be a plane wave which is received across all elements at excess delays related to the phase shift between elements. For a simple linear array, the phase shift $\Delta \phi$ between two neighboring elements is,

$$\Delta \phi = \frac{2\pi dsin(\theta)}{\lambda},$$  

(2)
where \( d \) is antenna spacing in metres, \( \theta \) is beam direction and \( \lambda \) is wavelength. The received signals are shifted and coherently summed for all desired \( \theta \).

The main benefit which NIST’s SAMURAI virtual array system presents is the ability to customize the array pattern to any 2-dimensional or 3-dimensional array shape with very short setup time and a positional accuracy high enough to operate reliably in the mmWave range. The MECA500 robot arm is easily programmable to move through any number of positions in 3D space for virtual element measurements. This is compared to a fixed antenna array where a mechanical design must be made for a single array configuration and individual antenna elements may be prohibitively expensive at mmWave costing upwards of \$400-\$700 or require computationally expensive measurement- and simulation-based characterization for custom antenna designs.

The MECA500 also allows for quick changes, entirely through software, between different array geometries. As an example, a linear array may be sufficient in one case where beam direction is not expected to vary much from the array boresight, leading to simpler post-processing. If measurements through a greater angular spread is desired, one could switch to a circular, cylindrical or spherical array, which avoids the limitations of a linear arrays such as the increasing of beamwidth at wide angles [47] and beam-squinting, which is the effect of the beam direction being frequency-dependent when using time delay circuits for phase control at the antenna element feeds [48]. Exploring the advantages and limitations between different array geometries is also a valuable educational exercise for wireless propagation courses.
3.3 Implementation

The virtual array is implemented using a system modeled after the SAMURAI project developed by NIST. Figure 6 shows the system diagram. The system is designed to be portable to allow both in-lab and field measurements.

![System Diagram](image)

Figure 6: System diagram of SAMURAI System replicated at UBC.

The system uses a MECA500 0.5m robot arm [49] with a 14 dBi WR-42 horn antenna mounted at the tip. The code for the project is written in Python and is downloadable via Gitlab [50]. Commands are sent from a computer through an Ethernet switch to the robot arm. The mounted antenna can be moved through any path within the arm’s range of motion, as shown in
Figure 7. The robot arm is stepped from location to location to create a virtual array. Many common antenna array configurations, such as linear, planar, circular, and hemispherical, can be traced. The mechanical specifications of the robot arm are listed in Table 1 and the maximum array dimensions are listed in Table 2. The antenna is attached to the robot using a 3D printed mount, which holds a WR-42 straight waveguide section and the horn antenna.
Table 1: Mechanical specifications for the MECA500, from user manual [49].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Position repeatability</td>
<td>0.005 mm (5 microns)</td>
</tr>
<tr>
<td>Rated payload</td>
<td>0.5 kg</td>
</tr>
<tr>
<td>Max. payload</td>
<td>1.0 kg (under special conditions)</td>
</tr>
<tr>
<td>Weight of robot arm</td>
<td>4.5 kg</td>
</tr>
<tr>
<td>Range for joint 1</td>
<td>$[-175^\circ, 175^\circ]$</td>
</tr>
<tr>
<td>Range for joint 2</td>
<td>$[-70^\circ, 90^\circ]$</td>
</tr>
<tr>
<td>Range for joint 3</td>
<td>$[-135^\circ, 70^\circ]$</td>
</tr>
<tr>
<td>Range for joint 4</td>
<td>$[-170^\circ, 170^\circ]$</td>
</tr>
<tr>
<td>Range for joint 5</td>
<td>$[-115^\circ, 115^\circ]$</td>
</tr>
<tr>
<td>Range for joint 6</td>
<td>$[-36,000^\circ, 36,000^\circ]$</td>
</tr>
<tr>
<td>Max. speed for joint 1</td>
<td>$150^\circ$/s</td>
</tr>
<tr>
<td>Max. speed for joint 2</td>
<td>$150^\circ$/s</td>
</tr>
<tr>
<td>Max. speed for joint 3</td>
<td>$180^\circ$/s</td>
</tr>
<tr>
<td>Max. speed for joint 4</td>
<td>$300^\circ$/s</td>
</tr>
<tr>
<td>Max. speed for joint 5</td>
<td>$300^\circ$/s</td>
</tr>
<tr>
<td>Max. speed for joint 6</td>
<td>$500^\circ$/s</td>
</tr>
<tr>
<td>Max. TCP linear velocity in joint mode</td>
<td>more than 2,000 mm/s</td>
</tr>
<tr>
<td>Max. TCP linear velocity in Cartesian mode</td>
<td>500 mm/s</td>
</tr>
<tr>
<td>Max. power consumption</td>
<td>200 W</td>
</tr>
<tr>
<td>Input voltage</td>
<td>24 VDC</td>
</tr>
<tr>
<td>Operating ambient temperature range</td>
<td>$[5 , ^\circ\text{C}, 45 , ^\circ\text{C}]$</td>
</tr>
<tr>
<td>Operating ambient relative humidity range</td>
<td>[10%, 80%] (non-condensing)</td>
</tr>
<tr>
<td>IP rating</td>
<td>IP 40</td>
</tr>
</tbody>
</table>
For a phased array, a wider beamwidth is preferred so that all scatterers in the scene are illuminated/viewable. The narrower the beamwidth, the smaller the area that is illuminated by each antenna, and thus the angular range of the virtual array is reduced. Omnidirectional antennas would be preferred but are much less practical at mmWave due to their low gain, small effective area, and, as a consequence, the resulting high path loss. In choosing the antenna for this channel sounder, a balanced design choice had to be made between high gain to close the link budget and lower gain to maintain a wider element beamwidth (and as a result, a wider array beamwidth).

Table 2: Physical limits of array dimensions that can be generated by the MECA500.

<table>
<thead>
<tr>
<th>Array Geometry</th>
<th>Physical Limits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear/Planar</td>
<td>Height: 368mm</td>
</tr>
<tr>
<td></td>
<td>Width: 468mm</td>
</tr>
<tr>
<td>Cylindrical</td>
<td>Height: 368mm</td>
</tr>
<tr>
<td></td>
<td>Min. radius: 113mm</td>
</tr>
<tr>
<td></td>
<td>Max. radius: 260mm</td>
</tr>
<tr>
<td></td>
<td>Azimuth: $0^\circ \leq \theta_A \leq 360^\circ$</td>
</tr>
<tr>
<td>Spherical</td>
<td>Min. radius: 113mm</td>
</tr>
<tr>
<td></td>
<td>Max. radius: 260mm</td>
</tr>
<tr>
<td></td>
<td>Azimuth: $0^\circ \leq \theta_A \leq 360^\circ$</td>
</tr>
<tr>
<td></td>
<td>Elevation: $0^\circ \leq \theta_E \leq 145^\circ$</td>
</tr>
</tbody>
</table>
Antenna gains ranging from 18 dBi to > 25 dBi are common at mmWave, for example the NIST array presented in Section 2.1 with a gain of 18 dBi and the sub-THz channel sounder in Section 3.5 with 23 dBi. A gain of 14 dBi was chosen for the SAMURAI system which corresponds to a 33° beamwidth in azimuth and elevation to sufficiently illuminate the channel scatterers.

The transmitted signal is generated by a N9918B FieldFox (which has a maximum frequency of 26.5 GHz) and sent through Junkosha phase-stable cables to the horn antenna. The FieldFox is a portable spectrum analyzer with a VNA setting that allows for frequency sweeps. This maintains the system’s versatility since VNA measurements can be made easily in the field, where conventional tabletop VNAs are unreasonable.

The same system can be refitted for frequency bands higher than 26.5 GHz in the future by mounting the appropriate waveguide antenna size and replacing the FieldFox with a higher frequency-capable VNA, as demonstrated by the SAMURAI setup at NIST which operates up to 50 GHz and plans of expanding up to 75 GHz. The NIST setup has also increased positional accuracy by using a collection of cameras to track the antenna’s position, which can be integrated into UBC’s setup in the future. The MECA500’s position repeatability of 5 microns is high enough such that upgrading to even higher frequency operation in the future is possible. For example, at 300 GHz where wavelength is 1 mm, a virtual array with the MECA500 would have a positional accuracy of λ/200. The hardware upgrades required for increasing frequency capabilities of the SAMURAI platform is discussed in Section 3.5.

3.4 Validation

The basic capability of the SAMURAI virtual array system is demonstrated with a linear array measurement. The system was set up in the UBC Radio Science Lab to measure the
transmission between two 14 dBi horn antennas in a small lab space. The environment is stationary (no moving pedestrians or vehicles). Since most applications for mmWave are typically wideband, a sweep range of 2 GHz is used. The remaining sweep parameters – IF bandwidth, number of points, and averaging, will determine the measurement noise level, frequency sampling, sweep time and total measurement time. IF bandwidth determines the amount of thermal noise in each measurement, with a smaller bandwidth decreasing noise but increasing sweep time. The number of points determines the frequency resolution of the sweep, and with applying the inverse Fourier transform, the measured delay range which limits the longest possible delayed signal that can be measured. The measurement setup and VNA settings are summarized in Table 3. For a 16-element array spaced at $\lambda/2$, the total array width is 94mm and provides an array factor gain of 24 dB. A simulation of the received angular power spectrum for the measurement is shown in Figure 9. The simulation uses a Gaussian beam to approximate the 14 dBi horn antennas, shown in Figure 8(a) compared to the actual beam pattern. Figure 8(b) shows the simulated measurement setup.
### Table 3: Summary of Simulation parameters.

<table>
<thead>
<tr>
<th>Simulation Setup Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rx array type</td>
</tr>
<tr>
<td>Number of Rx array elements</td>
</tr>
<tr>
<td>Rx Element type</td>
</tr>
<tr>
<td>Rx Element spacing</td>
</tr>
<tr>
<td>Rx Element gain, HPBW</td>
</tr>
<tr>
<td>Rx boresight angle</td>
</tr>
<tr>
<td>Tx antenna</td>
</tr>
<tr>
<td>Tx gain, HPBW</td>
</tr>
<tr>
<td>Tx boresight angle</td>
</tr>
<tr>
<td>Tx-Rx distance</td>
</tr>
<tr>
<td>Centre frequency</td>
</tr>
</tbody>
</table>
Figure 8: Simulated virtual array measurement. (a) Comparison of Gaussian beam pattern approximation to actual 14 dBi horn antenna beam pattern from the Eravant Datasheet. (b) Setup for simulated measurement with simulated Tx and Rx beam patterns.
Figure 9: Angular power spectrum received by 16-element horizontal linear array and transmitted from a 14 dBi horn antenna simulated in MATLAB, assuming no scattering.
3.5 Front-End Design Comparison

The front end of a channel sounder was designed for the Radio Science Lab and is intended for operation in the sub-THz range, defined as < 300 GHz. In comparing the front end of the SAMURAI to the sub-THz design, the benefits and drawbacks of each design can be evaluated. Similar to other sub-THz measurement projects in literature [51][52], the system uses VNA extender modules from Virginia Diodes, Inc. (VDI) to increase a VNA’s maximum frequency range. The front-end design is shown in Figure 10. The configuration in Figure 10(a) is limited by the bandwidth of the orthomode transducer (OMT), whereas Figure 10(b) can operate in the full waveguide frequency band, however with only one transmit polarization.

![Diagram](image)

(a)

(b)

Figure 10: Two configurations of the sub-THz front-end design for (a) full-polarization channel operation, and (b) a dual polarization operation.

Four VDI extender modules collectively cover the range from 40 GHz to 220 GHz. OMTs based on circular waveguide have relatively narrow single-mode frequency ranges so additional consideration must be given to OMT centre frequency selection.
FCC proposed authorizing flexible use licenses for "fixed and mobile service" in the following bands in their "Spectrum Frontiers" report in 2016 [53], all of which were identified as candidate bands for IMT-2020: 42-42.5 GHz, 47.2-50.2 GHz, 50.4-52.6 GHz, 71-76 GHz, and 81-86 GHz. Within this range, 14.75 GHz of spectrum has been harmonized worldwide to facilitate global roaming and economies of international scale [54]. For 5G within the 40 GHz range, ISED is interested in the 40-43.5 GHz, 45.5-47 GHz, and 47.2 GHz bands as mentioned in their “Spectrum Outlook” report for 2018-2022 [54], with the 42.5-43.5 GHz band also being supported by Canada at WRC-19.

ISED allows license-exempt (LE) operations in the 57-64 GHz band and made the decision in 2019 to open up the 64-71 GHz band for the same purpose [55] and was selected as a band for 5G in the US, with China and the ITU-R having published discussions on this band as well [56]. IMT-based mobile service in this range was adopted in the 2019 ITU-R Word Radiocommunications Conference (WRC-19) [57], however such applications are not being considered in Canada. 60 GHz is of interest for indoor wideband communications [58] with WiGig devices having been certified under RSS-210 (Canadian Radio Standards Specification) [59].

The 71-76 GHz and 81-86 GHz bands are designated for point-to-point backhaul systems in Canada [60] with China supporting allocation for the 81-86 GHz band for 5G and the ITU-R considering its application for space-to-earth communications [56]. ISED has identified frequencies above 95 GHz as Priority 3 and therefore has not released any specific bands yet due to uncertainty in international developments and equipment availability in this frequency range [54]. FCC Spectrum Horizons proposed the 116-123 GHz, 174.8-182 GHz, and 185-190 GHz bands for unlicensed use [61] but may require limitations due to passive Earth Exploration Satellite
Services (EESS) [62]. Considering the bands chosen for the VDI modules, several centre frequencies of interest are suggested for frequency band, shown in Table 4.

Table 4: Frequency range coverage of the four selected VDI frequency extender modules.

<table>
<thead>
<tr>
<th>Band Designation</th>
<th>Waveguide Size</th>
<th>Frequency Range (GHz)</th>
<th>Sub-bands of interest (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>U</td>
<td>WR-19</td>
<td>40-60</td>
<td>42.3, 48.7, 51.5, 58.5</td>
</tr>
<tr>
<td>E</td>
<td>WR-12</td>
<td>60-90</td>
<td>65.5, 73.5, 83.5</td>
</tr>
<tr>
<td>F</td>
<td>WR-8</td>
<td>90-140</td>
<td>93.5, 119.5</td>
</tr>
<tr>
<td>G</td>
<td>WR-5</td>
<td>140-220</td>
<td>178.4, 187.5</td>
</tr>
</tbody>
</table>

The key benefit to this front-end design which the SAMURAI system lacks is the capability to both transmit and receive at vertical (V) and horizontal (H) polarizations, enabling quad-polarized channel characterization: V-V, H-H, H-V, and V-H. All four measurements are typically used to characterize the cross-polarization and depolarization effects of a channel [63] and calculate a signal’s state of polarization using the Stokes parameters [64]. Polarization data has shown to be applicable in remote sensing applications such as target detection [64] and orientation-sensitive scattering [66] with particular interest in the THz range [67].

Combining the VNA extenders with the SAMURAI design is possible with future hardware upgrades. The high-gain antennas used in the quad-polarized setup in Figure 10 is beneficial for closing the link budget at longer Tx-Rx separations, but is too narrow in a virtual array context, and therefore lower gain (i.e., wider beamwidth) antennas at each waveguide band would be required. The VNA extension modules cannot be mounted directly to the robot arm as
they weigh 4 lbs. (1.8 kg) and 2 lbs. (0.9 kg), for the Tx and Rx modules respectively, which exceed the MECA500’s rated payload of 0.5 kg. The maximum frequency is then limited by a phase-stable cable that would run from the extender module test port to a wide-beamwidth antenna mounted on the robot arm. Junkosha offers phase-stable cables rated up to 67 GHz, and Gore’s PHASEFLEX-branded cables up to 110 GHz.

3.6 Discussion

The SAMURAI channel sounder designed by the Metrology for Wireless Systems Group is a highly versatile tool that presents a more cost-effective approach to variable beamwidth measurements than the switched array channel sounder used in Chapter 2, as it only requires a single receiving antenna. A version of the SAMURAI system was realized in the Radio Science Lab at UBC for operation up to 26.5 GHz and its advantages and capabilities were compared to a different channel sounder front-end design, intended for dual- and quad-polarized measurements in the sub-THz range. An upgrade path to combine the front-ends is available, such that virtual array measurements could be made up to 110 GHz. Basic operation of the SAMURAI system is demonstrated with a 16-element linear array simulation.
Chapter 4: Variable Beamwidth Synthesis

4.1 Introduction

In Chapter 2, the measurement process using the switched array channel sounder at NIST was detailed, with an alternate approach based on a virtual array presented in Chapter 3. This chapter details our method of generating a CIR and ACF with a variable beamwidth using the dataset of extracted paths from Section 2.3.

In Iqbal’s work, the CIR was acquired by mechanically steering the Tx horn and the Rx horn towards individual scatterers in a lecture room (Black Board, Wall, etc.) that were identified a priori. Our approach was similar, but instead 128 CIRs were acquired and coherently combined to extract individual channel paths and then reconstruct a single beamwidth-dependent CIR by applying a synthetic horn with variable beamwidth to the extracted paths. The advantage of our approach was three-fold:

1. A beamwidth-dependent ACF (and a beamwidth-specific correlation distance, described in Chapter 5) could be computed from the beamwidth-dependent CIR, rather than an ACF specific to the beamwidth of the horns used for measurement;
2. The synthetic horn could be steered towards persistent scatterers identified a posteriori from the acquisitions, rather than identifying presumptive scatterers a priori;
3. The synthetic horn could be steered towards the exact angle of the scatterer identified, enabling confirmation of Va’s theoretical model for impact of misalignment on correlation distance.
Details of our approach are provided in this chapter.

4.2 Persistent Paths

The LoS path and specular paths from ambient scatterers tended to be strong and persistent across the rail, which are desirable traits for beam steering. Diffuse paths, which clustered around specular paths in angle and delay, tended to be much weaker and could vary over fractions of a wavelength [68][69][70], causing the total number of paths $N$ extracted per displacement to vary over the rail despite a length of only 90 cm. Table 5 displays the mean and standard deviation of $N$ over the rail per environment. The mean number was greater indoors due to more scatterers, such as metallic instruments in the Laboratory. The mean number was also greater due to less free-space loss by virtue of smaller dimensions (shorter path lengths), enabling detection of more diffuse paths, in turn giving rise to a greater standard deviation. Outdoors, variation in the number

Table 5: Average* number of paths extracted per environment.

<table>
<thead>
<tr>
<th>Environment</th>
<th>All paths ($N$)</th>
<th>Persistent paths</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Mean of $N$</td>
<td>Std. dev. of $N$</td>
</tr>
<tr>
<td></td>
<td>(over rail)</td>
<td>(over rail)</td>
</tr>
<tr>
<td>Laboratory</td>
<td>74.5</td>
<td>4.3</td>
</tr>
<tr>
<td>Lecture room</td>
<td>35.5</td>
<td>5.1</td>
</tr>
<tr>
<td>Lobby</td>
<td>24.9</td>
<td>2.4</td>
</tr>
<tr>
<td>Pathway</td>
<td>11.1</td>
<td>1.5</td>
</tr>
<tr>
<td>Courtyard</td>
<td>15.3</td>
<td>2.1</td>
</tr>
</tbody>
</table>

*Average over the four scenarios per environment.
of paths was caused mostly by obstruction and scattering from foliage, but diffuse scattering from buildings was also observed.

Such variation called for a robust technique to reliably identify persistent paths. While the LoS track was detected across the whole rail in all scenarios, other tracks were subject to the birth and death process explained in Section 2.3. Some tracks extended across just a few displacements while others extended across hundreds. Upon tracking, a persistent path was identified empirically as a path that was tracked across at least one-third of the rail (30 cm). Complete details of the tracking technique are provided in [69].

The last step was to identify the source of the persistent paths. The LoS path was easily identified as earliest in delay and strongest. Specular paths, rather, were mapped against scatterers by raytracing their joint angle and delay to the incidence locations of salient objects visible in the maps, photographs, and 360° videos of the environments – walls, whiteboards, shelves, and pillars indoors; buildings and doorways outdoors. Figure 5(b) and Figure 5(c) display five persistent paths identified in the Laboratory scenario, most of which are visible in Figure 2. The azimuth AoA and delay of the paths varied gradually along the rail, attesting to the accuracy of our system¹. Of particular relevance to our work was the variance in the AoA – some paths varied up to a couple of degrees – capturing the impact of misalignment predicted by Va. Note furthermore how the Toolbox, Whiteboard, Shelf, and Far Wall exhibited more variation in path properties as the reflected paths traversed their surfaces compared to the LoS path that simply propagated through

¹ The inherent measurement error of the path properties per displacement was reduced by averaging the properties over the tracks through a sliding window.
air. The greater variation was due to their non-flat surfaces affecting delay and angle, and their composite materials affecting path gain. For example, the recessed face of the Toolbox with highly reflective door handles can be observed in Figure 2. Table 5 also shows the average number of persistent paths per environment, which ranged between 3.8 and 5.5.

4.3 Beamwidth-Dependent CIR

With the persistent paths in hand, the next step in our approach was to reconstruct the beamwidth-dependent CIR corresponding to the synthetic horn steered towards one of the paths. Consistent with the Clarke-Jakes ring of scatterers model, Va’s directional ring of scatterers model only considered receiver motion. The assumption inherent to both models is that all scatterers are illuminated by a transmitter equipped with an omnidirectional antenna. For consistency with our analysis as well, we made the same assumption that the synthetic horn was applied at the receiver only and hence was a function of AoA only. The implication is that the AoD information extracted from our measurements was not directly exploited in the analysis. The two dimensions that AoD added ([\(\theta^T_A, \theta^T_E\)]) to the six-dimensional property space of the paths was nevertheless extremely beneficial in resolving and tracking individual paths, as well as mapping specular paths against scatterers.

The synthetic horn had a 3D Gaussian pattern with unity gain and half-power beamwidth (HPBW) defined in degrees by \(\omega\) [71],

\[
g_n(\theta^R, \omega) = e^{-\left(\frac{\theta^R - \theta^R_n(\omega)}{0.6 \omega}\right)^2}
\]

(3)

Va’s model also applied a Gaussian beam pattern, which is accurate for the typical horn antennas [72] employed both in our measurements as well as in Iqbal’s. The horn was steered towards
\( \theta_n^R(d_n^{BIR}) \), the AoA of persistent path \( n \) at its birth displacement, equivalent to perfect alignment. The synthetic horn was applied to the spatial CIR \( h(d, \theta, \tau) \) in (1) to reconstruct the beamwidth-dependent CIR as,

\[
h_n(d, \omega, \tau) = \int_{\theta^R} g_n(\theta^R, \omega) \cdot h(d, \theta^R, \tau) \, d\theta^R + w(d, \tau).
\] (4)

The gain \( g_n(\theta^R, \omega) \) effectively attenuated some path \( \hat{n} \) in proportion to its angular separation, \( \theta_n^R(d) - \theta_n^R(d_n^{BIR}) \), from the steering angle as the receiver moved along \( d \), where \( \omega \) controlled the roll-off. The paths were then integrated over all arrival angles \( \theta_n^R(d) \) and white noise \( w(d, \tau) \) was added to the sample at delay \( \tau \). The sampling rate was 40 Gsamples/s, matched to the sampling rate of our channel sounder.

![Figure 11: Comparison between a CIR measured with a real horn and a CIR reconstructed with a synthetic horn whose parameters (gain, beamwidth, and steering angle/orientation) were matched to the real horn. Actual noise from the measured CIR was used to fill in samples in the reconstructed CIR that had no paths.](image-url)
Figure 11 attests to the fidelity of the reconstruction. Displayed is a CIR measured with a real horn in the Laboratory scenario together with the CIR reconstructed from the synthetic horn and the spatial CIR corresponding to the scenario. The parameters of the synthetic horn (gain, beamwidth, and steering angle, \(i.e.,\) orientation) were matched to the real horn. As can be seen, the CIR was faithfully reconstructed, where the peaks correspond to channel paths.

### 4.4 Beamwidth-Dependent ACF

Next, the complex ACF was computed from (4) as [73],

\[
R_n(\Delta d, \omega) = \frac{\int_{\tau} h_n(d_n^{BIR}, \omega, \tau) \cdot h_n^*(d_n^{BIR} + \Delta d, \omega, \tau) \, d\tau}{\int_{\tau} |h_n(d_n^{BIR}, \omega, \tau)|^2 \, d\tau}.
\]

(5)

The ACF quantifies the rate at which the beamwidth-dependent CIR steered towards persistent path \(n\), \(h_n(d, \omega, \tau)\), decorrelates with incremental displacement \(\Delta d\) along the rail. In practice, \(R_n(\Delta d, \omega)\) was computed along the path’s track, from \(d = d_n^{BIR} (\Delta d = 0)\) to \(d = d_n^{DTH} (\Delta d = d_n^{DTH} - d_n^{BIR})\). The denominator normalizes the maximum value of the ACF (at \(\Delta d = 0\)) to 1.

Figure 12(a) displays the ACF steered towards the Far Wall scatterer in the Laboratory scenario, for various beamwidths. For \(\omega = 10^\circ\), the receiver beam was highly focused on the specular path while diffuse paths clustered locally had AoAs that were slightly offset, and so were admitted into the beam only marginally. As beamwidth widened, attenuation on the diffuse paths was relaxed, giving rise to oscillations (widening the Doppler spread) due to multipath. Precisely, the ACF could be viewed as a composite of complex-valued ACFs, each from a separate path in the beam with its own AoA and in turn its own Doppler shift, \(i.e.,\) rate of phase rotation versus displacement [28]. As more and more paths were admitted, the oscillations intensified (the Doppler
spread widened); at $\omega = 360^\circ$, the ACF resembled the Rayleigh-Rice ACF, which in fact corresponds to the omnidirectional case of a rich scattering environment.

The oscillations were observable thanks to the complex-valued ACFs and, in this respect, were comparable to Iqbal’s complex-valued ACFs in Figure 12(b). Although their beamwidth was fixed at $\omega = 30^\circ$, the various scatterers (Black Board, Wall, etc.) had different oscillations nonetheless. As a reference, results are compared to the Rayleigh-Rice ACF modelled as $I_0(2\pi f_{MAX} \Delta d / \nu)$, where $I_0$ is the zeroth-order Bessel function of the first kind [21]. The maximum Doppler shift $f_{MAX}$ for both plots in Figure 12 corresponds to a receiver speed of $\nu = 10$ km/h. In both our case and Iqbal’s, the calculated ACFs exhibited much higher correlation than the Rayleigh-Rice ACF.
Figure 12: Comparison between a set of our beamwidth-dependent ACF results and that of Iqbal et al. in [21]. (a) Beamwidth-dependent ACF measured for the Far Wall scatterer in our Laboratory. (b) ACF measured by Iqbal et al. by mechanically steering narrowbeam (30°) horns towards select scatterers in a lecture room [21].
Chapter 5: Correlation Distance

In this chapter, correlation distance as a function of beamwidth was computed for all persistent paths identified. The resultant correlation profiles were then characterized through a number of salient metrics and subsequently aggregated over the 20 scenarios for a comprehensive statistical representation.

5.1 Correlation Profiles

The original use of the term coherence intended the length of time or space over which a signal did not change appreciably and fading was due to multipath effects only, that is, the coherent summation of paths with similar amplitudes and random phases [14]. Such fading is witnessed in rich, omnidirectional, wideband scattering environments, resulting in a wide-sense stationary (WSS) channel [74], meaning that the CIR’s second-order statistics, *i.e.*, its mean and variance, are constant. In turn, the Rayleigh-Rice ACF used to quantify coherence is dependent only on the incremental displacement of the receiver, not on its absolute location [11].

The spatial stationarity of a channel is a concept different than WSS and refers to the spatial stationarity of persistent paths that arises from environment geometry, forming stationary regions [75] or local regions of stationarity [19] in popular MIMO channel models. While recent measurements have demonstrated that WSS does not hold for sparse, directional, wideband channels [76] in which persistent paths are dominant, as confirmed by our own measurements, the ACF has nonetheless been employed to quantify the rate at which these channels decorrelate with displacement [11][36][37][50][76]. Although these papers still use the term coherence distance, we feel that the more appropriate term to capture both wide-sense and spatial stationarity is
correlation distance. Accordingly, the correlation distance of the signal, \textit{i.e.}, the beamwidth-dependent CIR steered towards persistent path n, was defined as the incremental displacement where the signal’s ACF first fell below 0.5, computed from (5) as,

$$\Delta d_n(\omega) = \Delta d \bigg|_{R_n(\Delta d, \omega) = 0.5}. \quad (6)$$

The upper bound on the correlation distance corresponds to the best-case scenario, for which the channel only has one path – the persistent path n and no other multipath. In this case, the upper bound is determined by two factors:

1. \textit{Bandwidth}: the wider the signal bandwidth (B), the narrower the pulse, and thus the faster the signal decorrelated when in motion;

2. \textit{Misalignment}: the greater that angle between the path and the receiver motion $[\theta_n^R, \theta_n^E]$ (the direction of receiver motion is set to $\theta^R \equiv [0° \ 0°]$ in the coordinate system), the faster the decorrelation.

Given these two factors, the correlation distance is bound by,

$$\Delta d_n(\omega) \leq 0.5 \cdot \frac{c}{B} \cdot \left| \cos \theta_n^R \cdot \cos \theta_n^E \right| \leq 0.5 \cdot \frac{c}{B} \cdot \frac{|f_n|}{f_{\text{MAX}}}, \quad (7)$$

where $c$ is the speed of light. Sometimes, the upper bound is preferred in terms of the path’s Doppler shift $f_n$ relative to the maximum Doppler shift $f_{\text{MAX}}$, where $\frac{f_n}{f_{\text{MAX}}} = \cos \theta_n^R \cdot \cos \theta_n^E$ [77]. It follows from (7) that paths more aligned with the receiver motion inherently had longer correlation distance; in particular, the upper bound $\Delta d_n(\omega) = 0.5 \cdot \frac{c}{B} = 7.5$ cm could be achieved only in the case of perfect alignment, either when the receiver moved
directly towards the scatterer ($\theta_n^R = [0^\circ 0^\circ]$), $f_n = f^{MAX}$, i.e., maximum Doppler shift) or directly away from it ($\theta_n^R = [180^\circ 0^\circ]$), $f_n = -f^{MAX}$, i.e., minimum Doppler shift).

Here we compute the ACF of the system pulse $p(\tau)$ (see Section 2.3), assuming that the channel has only one path whose propagation is represented by the pulse. The ACF has the same format as (5),

$$R_p(\Delta \tau) = \frac{\int_\tau p(\tau) \cdot p^*(\tau + \Delta \tau) d\tau}{\int_\tau |p(\tau)|^2 d\tau}. \quad \text{(8)}$$

Its computation is depicted in Figure 13, where the pulse received at incremental displacement $\Delta d = 0$ ($p(\tau)$ shown in red) is autocorrelated with the pulse received at some other $\Delta d$ along the rail ($p(\tau + \Delta \tau)$ show in green).

For the special case in Figure 13(a), the direction of propagation is aligned with the rail, so the incremental delay is given simply as $\Delta \tau = \Delta d / c$. By substituting in (8), the ACF could be written conveniently in terms of $\Delta d$ as,

$$R_p^{ALIGN}(\Delta d) = \frac{\int_\tau p(\tau) \cdot p^*(\tau + \frac{\Delta d}{c}) d\tau}{\int_\tau |p(\tau)|^2 d\tau}. \quad \text{(9)}$$

According to (5), the pulse’s correlation distance is defined as where its ACF first fell below 0.5, or,

$$\Delta d_p^{ALIGN} = \Delta d \bigg|_{R_p^{ALIGN}(\Delta d) = 0.5}. \quad \text{(10)}$$
Figure 13: Correlation distance of the system pulse $p(\tau)$. (a) Special case where the pulse’s propagation direction is aligned with rail. (b) General case where pulse propagates in direction $\theta_p = [\theta_p^A \theta_p^E]$ with respect to the rail.
The actual pre-distorted pulse was used to solve (10) numerically, resulting in,

$$\Delta d_p^{ALIGN} \approx 0.5 \cdot \frac{c}{B},$$

(11)

where the correlation distance depended only on the pulse width (given from the system bandwidth, B).

For the general case depicted in Figure 13(b), where the pulse propagates in direction $$\theta_p = [\theta_p^A \theta_p^E]$$ with respect to the rail, the incremental delay projected along the rail is $$\theta_p = [\theta_p^A \theta_p^E]$$, resulting in a generalized expression for the ACF,

$$R_p(\Delta d) = \frac{\int_{\tau} p(\tau) \cdot p^* \left( \tau + \frac{\Delta d}{c} \cdot | \cos \theta_p^A \cdot \cos \theta_p^E | \right) d\tau}{\int_{\tau} |p(\tau)|^2 d\tau}. \tag{12}$$

The general expression for the pulse’s correlation distance follows as,

$$\Delta d_p \approx 0.5 \cdot \frac{c}{B} \cdot | \cos \theta_p^A \cdot \cos \theta_p^E |. \tag{13}$$

The correlation distance was computed versus beamwidth, tracing out a correlation profile; consider the profiles in Figure 12(a) for four persistent paths identified in a Laboratory scenario. Note that this was for a different Laboratory scenario than in Figure 5, underscoring that different scatterers were observed for different Tx-Rx positions due to the limited FoV of the channel sounder. Let $$\Delta d_n(\omega^{MAX})$$ denote the maximum correlation distance computed, falling at the maximum correlation beamwidth $$\omega^{MAX}$$. The slight misalignment of the LoS path ($$\theta_{LoS}^R = [-9^\circ 0^\circ]$$) caused $$\Delta d_n(\omega_{LoS}^{MAX}) = 7.23 \text{ cm}$$ to fall short of the upper bound. Although the reflection from the wall in back of the receiver (Back Wall) had the same, but opposite, misalignment as the LoS path ($$\theta_{BW}^R = [171^\circ 0^\circ]$$), $$\Delta d_n(\omega_{BW}^{MAX}) = 2.97 \text{ cm}$$ was significantly shorter due to multipath interference from diffuse scattering, whereas the LoS path was void of diffuse scattering altogether. The other
two specular paths, from the Whiteboard ($\theta_{WB}^{R} = [32° \ 0°]$) and the Shelf ($\theta_{Shelf}^{R} = [-58° \ 0°]$), had greater misalignment than the LoS path and the Back Wall, hence their peaks were more pronounced. As beamwidth was expanded further, more and more diffuse paths were admitted, causing a further drop in correlation.

The measured correlation profiles in Figure 12(a) exhibited behavior similar to Va’s theoretical curves in Figure 12(b) when expanding the beamwidth from $0^\circ$. The behavior was characterized by an initial peak in correlation due to the improved alignment of the beam with the persistent path. The improvement was however offset by the additional scatterers admitted into the beam that disrupted the correlation. Eventually, when the beamwidth was expanded enough, the latter effect took over, dictating the drop-off.

5.2 Multiple Persistent Paths

The case investigated thus far – the same case investigated by Pätzold, Iqbal, and Va – was when the beam contained a single persistent path. In this case, the correlation distance dropped monotonically after the peak due to more and more diffuse paths being admitted to the beam. The case of a single persistent path will generally apply to antennas with narrow beamwidth, but how narrow? Moreover, what happens when other persistent paths are admitted into the beam? This section provides an explanation by expanding the beamwidth to $\omega = 360^\circ$, eventually admitting in all paths for each scenario.

Figures 14 to 18 display correlation profiles (solid curves) with the abscissa expanded to $\omega = 360^\circ$ – one illustrative scenario for each of the five environments – together with the scenario map in 2D. The direction of the receiver motion in the map is shown as a purple arrow ($\theta_{R,A} = 0^\circ$) and the azimuth AoA ($\theta_{n,A}^{R,A}$) of the persistent paths are also shown color-coded against the
scatterers in the legend. Returning to the Laboratory scenario in Figure 1(b), in the expanded view in Figure 14 it can be observed that after the initial drop-off of the Shelf due to local scattering, the LoS path was admitted into the beam around $\omega = 70^\circ$. The stronger LoS path dominated the beam, “pulling” the profile of the weaker path towards its own. Since a stronger path will generally have a longer correlation distance, this will cause the profile of the weaker path to rise back up after reaching a trough. Let $\Delta d_n(\omega^{MIN})$ denote the minimum correlation distance computed, occurring at the minimum correlation beamwidth $\omega^{MIN}$. The correlation profile of the Whiteboard was similar to the Shelf, however the Whiteboard’s admitted two other stronger paths along the rise back up, first the Shelf (around $\omega = 105^\circ$) and then the Back Wall (around $\omega = 170^\circ$), clearly indicated by positive steps in its profile.

An initial rise to maximum correlation due to misalignment, followed by a drop due to local scattering, and a subsequent rise due to other persistent paths in the beam was typical of all persistent paths per scenario, except for the strongest (LoS path). For the strongest path, rather, the initial rise was followed by a monotonic drop with negative steps along the way, each step occurring when another persistent path was encountered. Common to all persistent paths was the asymptotic correlation distance, defined as the correlation distance when $\omega \to \infty$. When all beams were omnidirectional, all $N$ paths were attenuated identically in each beam$^2$. The asymptotic correlation distance could be viewed as a composite of the correlation distances of the individual

$^2$ Note that since $\omega$ denotes half-power beamwidth, the pattern is not perfectly omnidirectional at $\omega = 360^\circ$. 

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persistent paths weighed by their amplitudes. Accordingly, the asymptotic value typically converged towards the strongest paths.

While the correlation profile of most paths followed the general trend described above, many deviations from this were observed in the actual scenarios. For example, in the Lobby scenario in Figure 15, the LoS path’s correlation distance dropped precipitously due to the strong Pillar reflection separated from it by only 12.9°, and converged to an asymptotic value of only 1.34 cm. In the Lecture Room scenario in Figure 16, the Near Wall reflection, the weakest of the three persistent paths identified, experienced two troughs due to the stronger LoS path and the Far Wall reflection. The LoS path in the outdoor Pathway scenario in Figure 17 experienced no noticeable drop at all since the other persistent paths were much weaker. In the outdoor Courtyard scenario in Figure 18, the four persistent paths were spaced far apart; the azimuth AoA between successive paths was 37°, 50°, and 50°, and so the asymptotic value was not reached by $\omega = 360^\circ$. Finally, while most persistent paths identified in the scenarios were either the LoS path or specular reflections, those from Bldg 1 Doorway in the Pathway scenario and Doorway 1 and 2 in the Courtyard scenario were actually strong diffractions from metallic door frames.

Flat segments were observed in all correlation profiles over which widening the beamwidth between persistent paths did not admit sufficient multipath to alter the correlation distance, attesting to the sparsity of the mmWave channel. In general, weaker persistent paths were more vulnerable to multipath, and hence experienced more variation across the profile.

5.3 Obstructed LoS (OLoS)

We now investigate the case in which the LoS path was obstructed. This case is important since penetration loss at mmWave can be as high as 25 dB or more [1][78], and therefore in mobile
scenarios the LoS path will likely be intermittently blocked by buildings, humans, vehicles, foliage, etc. Although all measurements were collected in LoS conditions, OLoS conditions were created synthetically by discarding the LoS path in the spatial CIR in (1). The correlation profiles corresponding to OLoS conditions were also plotted in Figure 14, in the same color as the persistent paths in LoS conditions, with only symbols (no line).

The maximum correlation distance and beamwidth were essentially unchanged between LoS and OLoS conditions in all scenarios in Figures 14 to 18 – the solid and dashes curves diverged much after the maximum correlation beamwidth – since the LoS path was never local to the other persistent paths anyway. When present, the LoS path was much stronger than the other persistent paths, pulling their correlation distances towards its own; when absent, much more variation was observed. In most scenarios, the other persistent paths actually benefited from the obstructed LoS path, maintaining correlation for a wider range of beamwidth; the only exception was for the outdoor Pathway scenario in Figure 17, in which the Bldg1 Doorway and Bldg 1 Wall had rich local scattering. When there existed a persistent path much stronger than the others in OLoS, such as the NE Building in Figure 18, the strongest path mimicked the LoS path, experiencing no trough and pulling the others towards its own correlation distance. Note that in Figure 18, the asymptotic correlation distance was actually longer in OLoS conditions whereas in the other four OLoS scenarios (Figures 14-17), there was no dominant persistent path so the asymptotic value settled somewhere between the asymptotic values of the individual persistent paths.
Figure 14: (a) Beamwidth-dependent correlation distance for the full range of beamwidths and (b) layout for Laboratory.
Figure 15: (a) Beamwidth-dependent correlation distance for the full range of beamwidths and (b) layout for Lobby.
Figure 16: (a) Beamwidth-dependent correlation distance for the full range of beamwidths and (b) layout for Lecture Room.
Figure 17: (a) Beamwidth-dependent correlation distance for the full range of beamwidths and (b) layout for Outdoor Pathway.
Figure 18: (a) Beamwidth-dependent correlation distance for the full range of beamwidths and (b) layout for Outdoor Courtyard.
5.4 Statistical Representation

Across the 20 scenarios considered, correlation profiles were generated for a total of 88 persistent paths in LoS conditions and a total of 68 (88 minus 20 LoS paths) persistent paths in OLoS. As demonstrated in Figures 14 to 18, the common trend in the profiles was a rise to maximum correlation followed by a drop to minimum correlation, both occurring at relatively narrow beamwidth. Thereafter, the profiles could vary significantly from each other, but what happened for wider beamwidth was less relevant at mmWave since it is expected that antennas will have beamwidth less than 30° [11]. As such, for the purpose of comprehensive statistical representation across all the profiles computed, the individual profiles were characterized by their maximum and minimum correlation distance and maximum and minimum correlation beamwidth. These four metrics were then compiled into Cumulative Distribution Functions (CDFs) in Figure 19. Also compiled were CDFs for the differential correlation distance, defined as the maximum minus the minimum correlation distance, and for the differential correlation beamwidth, defined as the minimum minus the maximum correlation beamwidth. The latter two metrics characterized, respectively, the correlation distance lost due to local scattering and the beamwidth over which the loss occurred. All CDFs were partitioned into indoor and outdoor scenarios, and then further partitioned into LoS and OLoS conditions.

As explained earlier, there were three factors that affected maximum correlation distance: bandwidth, misalignment, and local scattering. Because bandwidth was equal for all scenarios and misalignment was arbitrary, the longer maximum correlation distance outdoor versus indoor in Figure 19(a) could be justified by poorer local scattering in part, as explained earlier, due to greater path loss, so the weakest diffuse paths went undetected. Indoor, the multipath interference from
the richer scattering caused the detected AoA to shift from its actual AoA, requiring a wider maximum correlation beamwidth than outdoors in Figure 19(b) to capture the persistent path reliably when in motion.

Not only were the persistent paths stronger indoors, there were more of them (see Table 5); in consequence, the destructive interference between them was more severe, translating into shorter minimum correlation distance and wider minimum correlation beamwidth in Figure 19(c,d). This effect was better evidenced through the differential correlation distance and beamwidth in Figure 19(e,f), which factored out the maximum correlation distance and beamwidth. While OLoS conditions did not significantly affect the maximum beamwidth, the minimum correlation beamwidth was significantly widened in absence of the strongest (LoS) path to overtake the beam quickly in the correlation profile. The minimum correlation distance was then reached later, allowing it more time to drop, hence a shorter minimum correlation distance compared to LoS.
Maximum correlation distance, $\Delta d_n^{MAX}$ (cm)

Cumulative distribution function
Indoor
Indoor (OLoS)
Outdoor
Outdoor (OLoS)

Maximum correlation beamwidth, $\omega^{MAX}$ (deg)

Cumulative distribution function
Indoor
Indoor (OLoS)
Outdoor
Outdoor (OLoS)
Cumulative distribution function

Minimum correlation distance, \( n_{MAX} - n_{MIN} \) (cm)

Indoor
- Indoor (OLoS)

Outdoor
- Outdoor (OLoS)

Cumulative distribution function

Minimum correlation beamwidth, \( \omega_{MIN} - \omega_{MAX} \) (deg)

Indoor
- Indoor (OLoS)

Outdoor
- Outdoor (OLoS)
Figure 19: Cumulative Distribution Functions (CDFs) for salient characteristics of correlation profiles, compiled across the 20 scenarios investigated. (a) Maximum correlation distance, (b) Maximum correlation beamwidth, (c) Minimum correlation distance, (d) Minimum correlation beamwidth, (e) Differential correlation distance, and (f) Differential correlation beamwidth.
Chapter 6: Intrinsic Delay Spread

6.1 Introduction

Scattering by objects and structures in the vicinity of the propagation path gives rise to time-delayed replicas of the transmitted signal at the receiver. The excess delay refers to the difference between the propagation delay associated with the direct path or LoS component and that associated with a given multipath component. The delay profile is the expected power per unit of time received with a certain excess delay. When a large set of impulse responses is averaged over time or location, the result typically follows an exponential shape when expressed in linear units [50].

Several different metrics are used to characterize delay spread including the maximum delay spread, average delay spread (first central moment) and rms (root-mean-square) delay spread (second central moment). The latter is the most commonly used measure. The threshold for the weakest multipath component affects the value of the rms delay spread. That is, as the threshold increases, the rms delay spread decreases. In previous work, it has been common to report multipath delay spread for various thresholds but little effort has been devoted to determining if a pattern exists. Here, we show that such a pattern does indeed exist and can be applied to both non-LoS (Rayleigh) and LoS (Rician) scenarios.

6.2 Concept

Consider a delay profile of the form,

\[ P(\tau) = P_0 e^{-(\tau-\tau_0) / \tau_1} \]  

(14)
where $\tau$ is the delay, $\tau_1$ is the delay constant and $\tau_0$ represents the propagation delay associated with the direct path. We make the following observations that are not widely recognized:

1. The parameter $\tau_1$ is identical to the rms delay spread given by the second central moment of the delay profile.

2. If the minimum signal threshold is set to zero, the calculated rms delay spread achieves its maximum value. We refer to this as the intrinsic delay spread of the channel. A simple closed form expression gives the actual delay spread given the value of the intrinsic delay spread and the ratio of the peak response to the minimum signal threshold, i.e., the dynamic range of the response.

3. In the case of a Rician channel in which a line-of-sight component is superimposed on the exponential delay profile, we use results presented in [79] as the basis for an exposition of how the actual delay spread is related to the intrinsic delay spread, Rician K-factor, and the minimum signal threshold or dynamic range.

### 6.3 Results

The path gains of a channel are simulated using the delay profile model in (14) by generating log-linear path gain values with slope $1/\tau_1$ from $P_0$ to $P_{min}$, then converting to linear units for $\tau_{rms}$ calculations.
Figure 21 shows example delay profiles for dynamic ranges set at 20 dB. $P_0$ is set to 0 dB so that system dynamic range, $DR$, which is the ratio of the peak response to the minimum signal threshold in units of dB, is conveniently equal to $-P_{\text{min}}$. The rms delay spread is calculated as,

$$
\tau_{\text{rms}} = \frac{\int_0^\infty (\tau - \mu_\tau)^2 P(\tau) d\tau}{\sqrt{\int_0^\infty P(\tau) d\tau}}, \quad \mu_\tau = \frac{\int_0^\infty \tau P(\tau) d\tau}{\int_0^\infty P(\tau) d\tau},
$$

where $\mu_\tau$ is the average delay spread [11]. A family of curves displaying the results are shown in Figure 22, plotting $\tau_{\text{rms}}$ as a function of system dynamic range for $K_0 = 0$ (Rayleigh case) and
8 dB (Rician case). The calculated rms delay spread is reasonably modelled by a bounded exponential,
\[
\tau'_{rms}(DR) = \tau_1 \left( 1 - e^{\frac{DR}{\beta}} \right),
\] (16)
where \( \lim_{DR \to \infty} \tau_{rms}(DR) = \tau_1 \) and \( \beta \) is the fitted exponential coefficient.

A Rician channel’s power-delay profile consists of a strong initial signal, referred to as the main component, in addition to the exponential component. The main component is modelled as a delta function at the lowest delay \( \tau_0 \) and is characterized separately from the exponential component. The ratio of the power in the main component, \( A \), to the power in exponential component, \( B \), scaled appropriately such that \( P_0 = A + B \), is the K-factor \( K_0 \). Combining (14) with the Rician delay profile in [79],
\[
P(\tau) = A\delta(\tau) + Be^{-\frac{(\tau-\tau_0)}{\tau_1}},
\] (17)
and rewriting (16) to include the effect of the K-factor,
\[
\tau'_{rms}(DR) = \tau_1 \left( 1 - e^{-\frac{(DR-\alpha)}{\beta}} \right), \quad \forall \ DR > \alpha,
\] (18)
where \( \alpha = -20\log_{10}(A), \ K_0 = A/B \), and \( \beta \) is the fitted exponential coefficient.
Figure 21: Generated delay profiles for $P_0 = 0$ dB, $\tau_1 = 0.4 \mu s$, $DR = 20$ dB, (a) Rayleigh case ($K_0 = 0$) and (b) Rician case with $K_0 = 8$ dB.
Discussion

These results provide a particularly useful framework to which time-domain channel measurements collected in built-up areas can be compared and assessed. The relationship between rms delay spread and the delay profile slope is an intuitive relationship which has not been previously revealed. A bounded exponential model can be used to predict rms delay spread as a function of system dynamic range, and, in agreement with previous work demonstrated in [79], can be expanded to the Rician case.

Figure 22: Family of curves for calculated $\tau_{rms}$ vs. system $DR$. Dashed lines are intrinsic delay spreads $\tau_1$. Annotated stars on the purple lines correspond to datapoints collected from delay profiles in Figure 21(a) and (b).
Chapter 7: Conclusions and Recommendations

In this thesis, we combined recent seminal works by Iqbal et al. in measuring correlation distance for a fixed beamwidth and by Va et al. in order to model correlation distance versus beamwidth, both at millimetre-wave. Specifically, we were the first, to our knowledge, to measure correlation distance as a function of continuous beamwidth, using a single channel sounder rather than multiple systems with different bandwidths. We found that correlation was maintained for much longer than is predicted by the Rayleigh-Rice fading model, the de facto standard for sub-6 GHz channels still widely used for millimetre-wave. We also found that correlation was maintained for a longer distance and for a wider beamwidth outdoor versus indoor due to less multipath. Another significant finding was that obstructed line-of-sight conditions were actually conducive to correlation since the line-of-sight path behaved as a dominant interferer to beams steered in other directions.

We also demonstrated how virtual arrays implemented using a robotic arm can realize variable beamwidth antenna apertures at a fraction of the cost of the multi-element array used in this study, albeit with some restrictions on the types of the channels that can be characterized, and revealed that the relationship between estimated and actual rms delay spread is simply related to the dynamic range of the measurement. These results will be put to use in the Advanced Multi-Mode Millimetre-Wave Channel Sounder that is currently being developed by Prof. David Michelson and his team in the Radio Science Lab at UBC. Capable of characterizing wireless channels from 26.5 to 220 GHz with high precision, it will play a key role in revealing the nature of channel impairments in this vast spectrum frontier.
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