

Beyond-Line-of-Sight Communications and Networks with Troposcatter and Atmospheric Ducts

by

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**Beyond-Line-of-Sight Communications and Networks with Troposcatter
and Atmospheric Ducts**

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This is to certify that I have examined this copy of a doctoral dissertation by

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*To my parents Ahmet and Meral Dinç
and my brother Özgün Dinç...*

ABSTRACT

Network centric operations (NCO) aim to provide a competitive advantage through information superiority. Modern military communication systems require communication links having low transmission delays and high data rates in order to provide effective command and control in challenging warfare areas. Available wireless communication systems are inefficient at b-LoS ranges. For this reason, new b-LoS communication techniques are required to provide high data rates, reliability and security. The main motivation of the proposed thesis is to analyze b-LoS wireless communications and networks with troposcatter and atmospheric ducts in order to tackle these problems.

Tropospheric scatter that is widely known as troposcatter, is the scattering of the propagating signals in all directions due to the irregularities in the troposphere. The scattered signals are directed especially in the forward direction, but some of the signal power is directed to the receiver. As a result, the effects of troposcatter can be utilized as a b-LoS communication medium. To this end, we investigate the viability of troposcatter channel as a b-LoS communication medium through physical layer channel modeling studies. In addition, we propose models for the fading correlation and rain attenuation in troposcatter links. At the end, our models show that troposcatter communication is especially promising for fixed links with low probability of detect/intercept, low transmission delays and high data rates.

Near-surface wave propagation at microwave frequencies especially 2 GHz and above shows significant dependence on atmospheric ducts that are the layer in which rapid decrease in the refractive index occurs. The propagating signals in the atmospheric ducts are trapped between the ducting layer and the sea surface, so that the power of the propagating signals do not spread isotropically through the atmo-

sphere. As a result, these signals have low path-loss and can travel over-the-horizon. Since atmospheric ducts are nearly permanent at maritime and coastal environments, ducting layer communication is a promising method for beyond-Line-of-Sight (b-LoS) communications especially in naval communications.

Although the wave propagation in atmospheric ducts has been studied, there are only a few papers which consider the ducting channel as a high data rate communication medium. Therefore, the communication aspects of this channel is mostly missing. To this end, we assess the viability of ducting communications for modern naval communications. Our results show that ducting communication can be utilized in challenging naval warfare conditions with secure and high data rate communications. To this end, we develop a channel modeling approach based on parabolic equation and ray-tracing methods. We analyze the viability of diversity techniques and investigate physical layer parameters of the ducting channels. At the end, we introduced a beam-forming method for ducting channels and investigate the feasibility of ducting ad hoc networks.

ÖZETÇE

Ağ merkezli harp uygulamaları bilgi üstünlüğü ile avantaj sağlamayı amaçlanmaktadır. Bundan dolayı, modern askeri haberleşme sistemleri özellikle ufuk ötesi mesafelerde düşük gecikme ve yüksek veri hızları sağlayabilen haberleşme kanallarına ihtiyaç duymaktadır. Klasik kablosuz haberleşme sistemleri ufuk-ötesi mesafelerde yetersiz olmaktadır. Bu kapsamda, önerilen tezin temel amacı; troposferik saçılım ve atmosferik oluklarının ufuk ötesi mesafelerde haberleşme kanalı olarak kullanımının incelenmesidir.

Troposferik saçılım atmosferdeki düzensizlikler nedeniyle gönderilen sinyallerin troposfer katmanında saçılma uğramasından kaynaklanır ve sinyal enerjisinin saçılmasına neden olur. Saçılımın tüm yönlere olmasına rağmen alıcı tarafında saçılan sinyalleri almak mümkündür. Böylece, aralarında açık görüş bulunmayan alıcı-verici arasında haberleşme sağlanabilir. Önerilen tez kapsamında, troposferik saçılım kanal modeli geliştirilerek haberleşme kanalı olarak kullanımının fizibilite analizleri yapıldı. Elde edilen kanal modeli geliştirilerek, troposferik saçılım kanalının fiziksel katman parametreleri ve yağmurun haberleşmeye etkileri incelendi. Ayrıca, çeşitlilik tekniklerinin incelenmesi amacıyla sönümleme korelasyonunun incelenmesi için analitik bir model önerildi. Çalışmalarımız sonucunda, troposferik saçılım kanalının özellikle sabit haberleşme uygulamalarında yüksek güvenlik, düşük gecikme ve yüksek veri hızı sağlayabileceği gösterildi.

Havanın kırıcılık indisindeki ani değişimler nedeni ile oluşan atmosfer olukları, yakın yüzey elektromanyetik sinyalleri yeryüzüne geri yansıtarak sinyallerin atmosfer olukları ile yeryüzü arasında hapsolmesine neden olur. Atmosfer oluklarında hapsolan sinyaller normal atmosfer koşullarına göre daha az dağılma göstererek daha düşük sinyal gücü kayıplarına uğramaktadır. Bu durum sinyallerin ufuk ötesi mesafelere

ulařmasını saęlar ve atmosfer olukları ufuk ötesi iletişim için haberleşme kanalı olarak kullanılabilir. Bu nedenle, atmosfer olukları özellikle deniz, okyanus ve kıyılarda ufuk ötesi haberleşme kanalı olarak kullanılabilir, çünkü bu bölgelerdeki yüksek nem oranları sayesinde atmosfer olukları hemen hemen her zaman oluşmaktadır.

Atmosfer olukları içinde dalga yayılması ile ilgili çalışmalar olmasına rağmen; atmosfer oluklarının yüksek veri hızı sağlayan bir kanal olarak kullanılmasını amaçlayan sadece birkaç çalışma bulunmaktadır. Bu nedenle, atmosfer oluklarının haberleşme yönünden incelenmesi konusunda pek çok sayıda açık problem bulunmaktadır. Bundan dolayı, önerilen tez kapsamında atmosfer olukları için parabolik denklem ve ışın izleme metotları kullanılarak bir kanal modeli geliştirildi. Bu kanal modeli kullanılarak, atmosfer olukları için çeşitlilik metotlarının fizibilite analizi ve fiziksel katman parametreleri incelenmiştir. Ayrıca, atmosfer oluklarına yönelik hüızmeleme teknikleri ile tasarsız aęlar için çözümler sunulmuştur.

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NOMENCLATURE

AoA:	Angle-of-Arrival
AoD:	Angle-of-Departures
AREPS:	Advanced Refractive Effects Prediction System
ASE:	Area Spectral Efficiency
AWGN:	Additive-White-Gaussian-Noise
b-LoS:	Beyond-Line-of-Sight
CDF:	Cumulative Distribution Function
CSI:	Channel State Information
CTS:	Clear-to-Send
dB:	Decibel
dBm:	Decibel-Milliwatt
DFCF:	Distance-Frequency Correlation Function
DMAC:	Directional Medium Access Control
FD:	Finite Difference
FE:	Finite Element
FFT:	Fast Fourier Transform
GPS:	Global Positioning System
HF:	High Frequency
ISI:	Inter-Symbol-Interference
ITU:	International Telecommunication Union
KLD:	Kullback-Leibler Divergence
LIDAR:	Light Detection and Ranging
LoS:	Line-of-Sight

MAC:	Medium Access Control
MIMO:	Multiple-Input-Multiple-Output
MRC:	Maximum Ratio Combining
NCO:	Network Centric Operations
OFDM:	Orthogonal-Frequency-Division-Multiplexing
PDS:	Power Delay Spectrum
PE:	Parabolic Equation
PL:	Path-loss
PPP:	Poisson Point Process
RMS:	Root-Mean-Square
RO:	Ray-optics
RSM:	Ring Scatter Model
RTS:	Request-to-Send
SATCOM:	Satellite Communications
SIMO:	Single-Input-Multiple-Output
SINR:	Signal-to-Interference-plus-Noise Ratio
SISO:	Single-Input-Single-Output
SNR:	Signal-to-Noise Ratio
SSF:	Split Step Fourier
TEMPER:	Tropospheric Electromagnetic Parabolic Equation Routine
UAV:	Unmanned Aerial Vehicle
W:	Watt
WTS:	Wait-to-Send

Chapter 1

INTRODUCTION

Modern military communication applications require high data rate applications, and available wireless communication systems are inefficient at beyond-Line-of-Sight (b-LoS) ranges. Therefore, new b-LoS communication techniques are required to provide high data rates, reliability and security. Therefore, the main motivation of the proposed thesis is to analyze b-LoS wireless communications and networks with troposcatter and atmospheric ducts.

Available b-LoS communication systems utilize either HF (high frequency) radio systems that are band-limited to support high data rates, or satellites that have intrinsic limitations due to the excessive transmission delays and capacity problems at regions having low satellite coverage, or relay nodes that can be exposed to hostile attacks and waste considerable amount of mission time. Thus, there is a significant need for a direct transmission channel between b-LoS units to empower strategical and tactical decisions. To this end, direct communication methods are required for b-LoS communications, and this thesis investigates troposcatter and ducting effects of the lower troposphere as b-LoS communication media.

1.1 Troposcatter

Troposcatter, that is also known as tropospheric scatter, is the scattering of the propagating signals due to the imperfections in the troposphere. Most of the scattered power is directed in the direction of the incoming ray. However, some portion of the scattered signals can be received at the receiver side by pointing the transmitter and receiver antennas to the radio horizon as in Figure 1.1. The intersection of the beam-

widths of the antennas forms the troposcatter common volume where the scattered signals can be directed to the receiver. In this way, the scattering effect can be utilized as communication medium for b-LoS links up to 300 km. Although 2 GHz is well-suited for the troposcatter communications, the modern troposcatter systems generally utilize 4.4 – 5 GHz spectrum due to licensing problems [1].

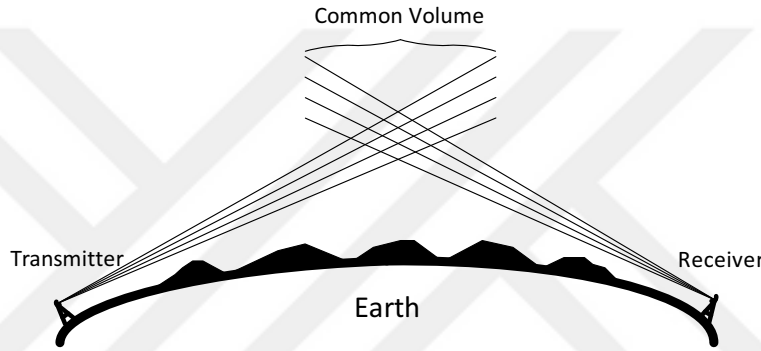


Figure 1.1: Troposcatter b-LoS paths.

In the past, troposcatter communication was widely used for both civilian and military b-LoS applications. However, due to very large antenna sizes and high transmission-losses, troposcatter applications were replaced by Satellite Communications (SATCOM). On the other hand, SATCOM has intrinsic limitations due to the excessive transmission delays and capacity problems at regions having low satellite coverage. Therefore, b-LoS communication requires alternative techniques especially for the military applications. For these reasons, troposcatter communication stands as a promising candidate for b-LoS communications with the advancements in the troposcatter equipment and communication techniques. Available state-of-the-art solid-state amplifiers and troposcatter modems can reach 2 kW and 22 Mbps for 4.4 – 5 GHz up to 250 km, respectively [2, 3]. Thus, troposcatter systems can be used in b-LoS links requiring high data rates especially with the employment of multiple-input-multiple-output (MIMO) systems.

Available troposcatter studies generally focus on the prediction of the annual and the worst month transmission-losses with statistical methods [4, 5]. Although these

statistical methods can capture the seasonal changes in the channel, fast variations and non-homogeneities in the channel cannot be modeled with such statistical approaches. In addition, [6] provided analytical expression for power delay spectrum (PDS) of the troposcatter links. In addition, there exists some experimental studies [7, 8]. In [8], the correlation between vertically spaced antennas are presented with real world measurements, and the distribution of the fading in the troposcatter channel is reported as Rayleigh. Based on the analytical and experimental studies, there are also some studies that focus on the performance metrics in the troposcatter channels [9–11].

In the available analytical studies, the scattering is modeled only with the scattering angle [6]. For this reason, these models are inefficient to derive PDS, coherence bandwidth, and root-mean-square (RMS) delay spread which show significant dependence on the channel conditions. We develop a comprehensive MIMO troposcatter channel modeling approach which considers the fluctuations and non-homogeneities in the air turbulence. In our approach, the troposcatter is modeled with Rayleigh scattering approximation [12], and the Kolmogorov spectrum technique is utilized to model the atmospheric changes with the water vapor mixing ratio datasets. Furthermore, we utilize the ray-based channel modeling approach to analyze the coherence time of the channel to determine the effects of the mobility in the troposcatter links for the first time in the literature.

The troposcatter communication introduces high losses at b-LoS distances, and the fluctuating atmospheric turbulence results in short term fading. To alleviate these effects, diversity systems can be utilized to improve the robustness of the communication systems. Troposcatter systems can employ space, frequency, angle and polarization diversity techniques. However, the correlation coefficient should be low for a diversity technique to be effective. Fading correlation is well studied for conventional wireless communication links. In addition, there are available experimental studies and empirical suggestions which consider the fading correlation in the troposcatter links [1, 5, 8, 13]. However, there is no theoretical study which considers the fading

correlation in the troposcatter links. To this end, we develop an analytical model for the fading correlation of the troposcatter antennas utilizing space, frequency and angle and space-frequency diversity techniques for the first time.

Troposcatter communication has high losses due to high path lengths and scattering. Therefore, path lengths should be minimized by making elevation angles of antennas as small as possible. In this way, troposcatter common volume, which is the intersection of transmitter and receiver antenna beams formed at the lower troposphere, and scattered rays are exposed to the effects of hydro-meteors. Since modern troposcatter equipments utilize frequency range between 4.7 – 15 GHz to provide more bandwidth, the most important hydro-meteor for troposcatter communications becomes rain because the effects of cloud and fog become considerable for carrier frequencies higher than 50 GHz [14]. There are many rain attenuation models in the literature [15–18]. Many of them consider rain attenuation in SATCOM [17, 18]. However, the geometry of the troposcatter significantly differs compared to SATCOM. Therefore, we develop a rain attenuation model for troposcatter communications which considers the distribution of rainfall rates and multiple rain cells.

1.2 Atmospheric Ducts

Atmospheric ducts that are caused by the rapid decrease in the refractive index of the lower atmosphere have tremendous effects on the near-surface wave propagation. In the presence of atmospheric ducts, signals below the atmospheric duct are trapped between sea surface and duct. Therefore, these signals do not spread isotropically through the atmosphere. Instead, the signals spread mostly in the ducting layer which is the area between the sea surface and the duct as in Fig. 1.2. As a result, the spreading loss decreases considerably compared to the standard atmosphere and there is a significant amount of increase in the received signal power. In this way, trapped signals can travel over-the-horizon and it makes the ducting layer convenient for b-LoS communications.

The formation of atmospheric ducts entirely depends on the refractivity of the

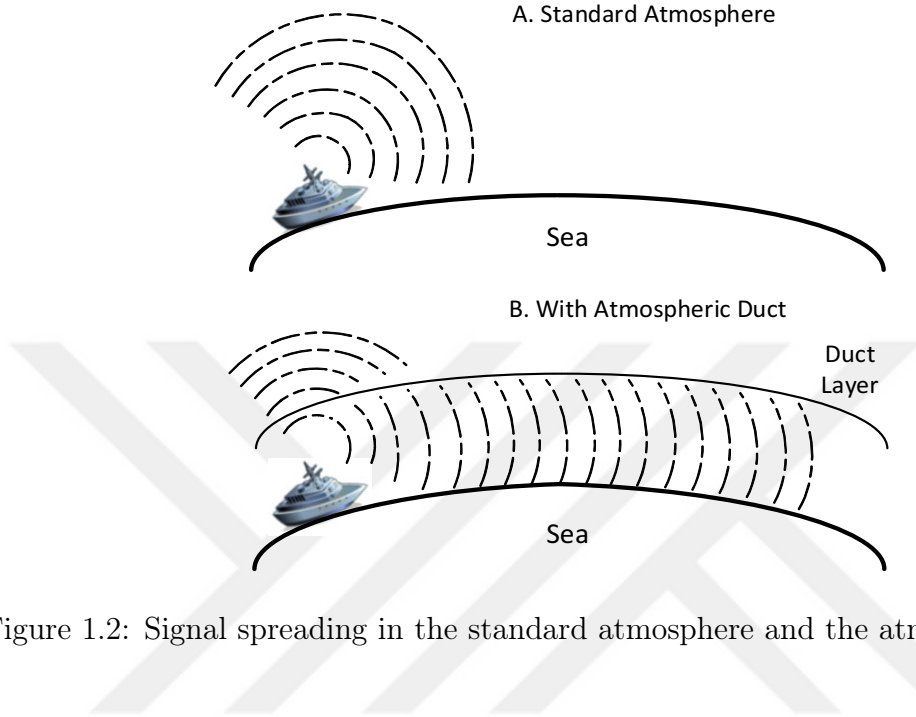


Figure 1.2: Signal spreading in the standard atmosphere and the atmospheric duct.

lower atmosphere that is determined by atmospheric parameters such as wind, pressure, temperature and most importantly humidity that is the key and essential factor in the duct formation as will be further discussed in Section 6.2 [19]. Experimental studies also show that atmospheric ducts are nearly permanent in the coastal and maritime environments due to high evaporation rates. As a result, ducting layer communication becomes the dominant propagation mechanism at the lower troposphere especially between 2 to 20 GHz. Since transmitter and receiver antennas should be located within the ducting layer, especially naval communications can utilize b-LoS communications with the ducting layer.

Ducting layer studies mainly focus on refractivity estimation techniques and radar path-loss calculations [19, 20]. However, there are a few studies which focus on the utilization of the atmospheric ducts as a communication medium. To this end, [21] provides a high data rate employment of the evaporation ducts for b-LoS sensor networks where continuous satellite communication is expensive and high data rate cellular communication is unavailable. In [21], ducting layer is utilized to monitor a reef site in Great Barrier Reef Australia. According to their results, the atmospheric

ducts can connect a 78 km link at 10.5 GHz with 10 Mbps at 80% of the time. In addition, atmospheric ducts are also utilized for television broadcasting in tropical regions.

Although wave propagation and refractivity profile modeling of atmospheric ducts are well-studied topics [22], there is no work which considers physical layer channel modeling for high data rate communication with atmospheric ducts. Therefore, we first examine the-state-of-the-art studies on the ducting communication links. Then, we develop a large-scale path-loss model for the surface ducts based on parabolic equation (PE) and ray-tracing methods. The proposed channel model is utilized to analyze delay spread and angle-of-arrival (AoA) of the surface ducts for the first time in the literature. By extending the proposed channel model, we compare space and polarization diversity systems in terms of achievable data rates. At the end, our results suggest that ducting channel is suitable for high data rate applications and the link can provide a few tens of Mbps data rate for 10 MHz bandwidth. In addition, we propose a slant-path rain attenuation model for ducting channel based on our previously introduced rain attenuation model for troposcatter [23].

In ducting layer, trapped signals are concentrated such that their AoA spread is very narrow. For this reason, antennas with sharp beam-widths can be used to provide high antenna gains, and large planar antenna arrays can be utilized to generate sharp electronically steerable beams. Thus, planar arrays are promising for ducting channel to detect the trapped signal by adjusting the signal beam in both elevation and azimuth direction. Although sharp angular spread of ducting channel can be exploited via large antenna arrays, there is no work which considers beam-forming for b-LoS ducting channels. To this end, we develop a limited-feedback multistage beam-forming algorithm for ducting links. Modern military applications require infrastructure-free communication between units. Thus, ad hoc networking provides flexible networks without requiring any infrastructure. Although atmospheric modeling and propagation within the duct layer have been attracted significant research attention, there is no work which considers ducting ad hoc networks. Therefore, we provide a system

model for ducting b-LoS ad hoc networks to perform coverage probability and area spectral efficiency (ASE) analysis.

To sum up, the contributions of this thesis can be listed as:

- Investigating the state-of-the-art for troposcatter and its communication aspects, discussing its viability and outlining open research areas
- Developing a ray-based MIMO troposcatter channel modeling approach to model fluctuations and non-homogeneities in the air turbulence
- Designing an analytical fading correlation model for troposcatter systems utilizing space, frequency, angle and space-frequency diversity techniques
- Investigating effects of rain attenuation in troposcatter links and modeling multiple non-uniform rain cells in the communication path
- Examining the state-of-the-art for ducting communication, discussing its viability and outlining open research areas
- Developing physical layer channel model for surface ducts based on ray-tracing and PE methods
- Analyzing physical layer channel parameters for ducting channels links such as coherence bandwidth, angle-of-arrival and delay spread
- Investigating correlation coefficient for space and polarization diversity in ducting links, and comparing their performances in terms of achievable data rates
- Design of a limited-feedback multistage beam-forming algorithm for ducting links by using planar antenna arrays
- Discussion of viability of ad hoc ducting networks, analyzing the coverage probability and ASE for ducting networks

Chapter 2

B-LOS COMMUNICATIONS WITH TROPOSCATTER

2.1 Introduction

Troposcatter is the scattering of the signals due to the irregularities at the troposphere, and it can be used as a b-LoS communication medium up to 300 km. Although the troposcatter communication was a popular b-LoS communication technique, SATCOM has been dominated the b-LoS communication applications. On the other hand, the troposcatter links have transmission delays on the order of milliseconds [1]. Therefore, the troposcatter channel is almost instantaneous. Since the troposcatter communication systems utilize narrow beam antennas, it provides point-to-point communication link. Thus, it has very low detect/intercept probability compared to SATCOM.

The remainder of the chapter is organized as follows. Section 2.2 provide the channel modeling techniques for the troposcatter communications. Section 2.3 includes the discussion of the diversity techniques for the troposcatter communications. The possible application areas are summarized in Section 2.4.

2.2 Channel Modeling Techniques

The geometry of the troposcatter communications is quite different from the conventional wireless communication systems. In addition, the changes and fluctuations in the atmospheric turbulence introduce variability in the scattered power according to the environmental and atmospheric parameters such as wind, temperature and climate. To this end, we review the channel modeling approaches that can be utilized for the b-LoS troposcatter communications.

2.2.1 Statistical Methods

ITU-R P.617 [5] provides the estimation of the transmission-losses (received power over transmit power) for the troposcatter communications according to the path geometry and climate. This method predicts the annual and the worst month transmission-losses for defined non-exceeding percentage of the time. Since ITU-R's (International Telecommunication Union - Recommendations) model is an empirical method, it provides reliable estimates for the transmission-losses. This model is most accurate for 5 GHz and below. Since the modern troposcatter applications mostly utilize 4.4 – 5 GHz band to provide high bandwidths and data rates [1–3], ITU-R P.617 is a promising method to model the troposcatter links in terms of the transmission-losses. In addition, the equations and parameters of ITU-R P.617 model depend on the climate of the region because the atmospheric turbulence strongly changes with atmospheric conditions. ITU-R P.617 [5] provides parameters and equations for 6 different climates: dry, humid, and etc.

We generate transmission-loss results for 4.7 GHz carrier frequency with 10%, 50% and 90% non-exceeding rates in Figure 2.1 by using ITU-R P.617 for dry climate (climate 3 in [5]). Although the results of the ITU-R transmission-loss prediction method are most accurate for 50% non-exceeding and above, the lower time percents can be calculated by assuming the transmission-losses are symmetrical as suggested in [5]. Since the troposcatter introduces high transmission-losses, the troposcatter systems utilize high power amplifiers up to 2 kW [2]. Assume that the minimum received power for the high data rate troposcatter link is -90 dBm [3] and the transmit power is 60 dBm. In this case, the transmission-loss threshold becomes 150 dB for the troposcatter link. As noticed from the Figure 2.1, the troposcatter link can provide the communication link up to 350 km at 90% of the time with this channel parameters. For lower ranges, the time percent of the communications link further increases, and the troposcatter channel can be utilized for b-LoS communication for 250 km and below almost all the time as also suggested in [1].

ITU-R model is employed in several studies to analyze the troposcatter channels.

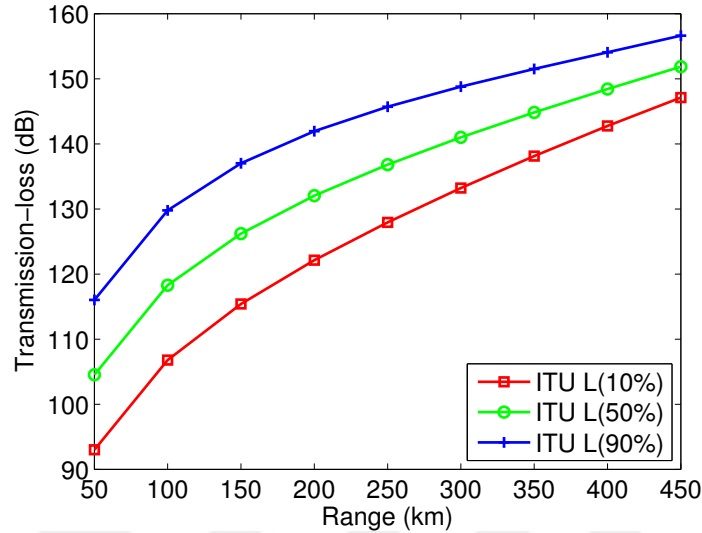


Figure 2.1: Path-loss vs. range for 4.7 GHz with ITU-R model.

For example, [24] utilizes the statistical method to estimate the data rates that can be provided for different time percents with the troposcatter communications. Although ITU-R P.617 provides reliable estimates for the transmission-losses, this model do not provide any information about other important channel parameters such as delay spread, coherence bandwidth, coherence time, correlation between antennas and PDS of the channel. Therefore, advanced methods are required for the troposcatter communication to analyze its channel characteristics.

2.2.2 Experimental Studies

There are several measurement campaigns for the troposcatter communications [8, 25, 26]. These studies provides significant parameters for the troposcatter links: path-loss, fade duration, fading distribution, delay spread and the correlation results for different diversity techniques. In [8], the employment of the space diversity in the troposcatter channel is analyzed. In addition, the experimental results for the path-loss, fade duration and fading distribution are provided. According to the experimental results, the power changes in the channel for small integration times shows Rayleigh

distribution. Therefore, the b-LoS troposcatter channel can be modeled as *Rayleigh fading channel*.

In [26], the angle diversity for the troposcatter links is analyzed and the correlation results for different beam separations are presented. [25] includes the experimental correlation results for the frequency diversity. The correlation results for frequency separations can be also used to calculate the coherence bandwidth of the channel because the 90% and 50% coherence bandwidth of the channel can be directly estimated with the correlation results for the frequency diversity [27]. [28] provides Doppler spread results for the fixed troposcatter communication, and the Doppler spread of the channel is reported as 15 Hz according to their measurements. Since this study considers the fixed communication channel, the time variations of the channel is caused by the atmospheric fluctuations. The results of [28] shows that the channel is slowly changing with the atmospheric variations by comparing MHz level bandwidths. Available experimental results can be utilized to verify the developed theoretical models, but advanced theoretical troposcatter models are required because it is very expensive to make measurement for each site. For this reason, it is required to estimate the troposcatter link characteristics with theoretical models.

2.2.3 Analytical Methods

There are also some analytical studies that models the troposcatter channel. The most important one is provided in [6] in which an analytical expression is provided for the PDS of the channel. However, this method only depends on the scattering angle. Thus, this model is seldom to include the effects of the atmospheric turbulence. [26] clearly shows that Bello's model [6] is inefficient to model the angle diversity in the troposcatter channel.

The available analytical models for the troposcatter communication depends only in the scattering angle. Therefore, these models are not able to estimate the PDS, coherence bandwidth, and RMS delay spread which show significant changes according to the path geometry and atmospheric conditions. For this reason, the ray-based

methods are promising to model the troposcatter communications.

2.2.4 Ray-based Methods

Since available troposcatter models cannot fully model the troposcatter channel, the ray-based methods are proposed for the troposcatter communications. To this end, we develop a powerful ray-based troposcatter channel modeling approach in [29]. In our ray-tracing approach, the power and delay of the each ray is calculated by using the developed differential scattering cross-section calculations. In this way, the PDS of the channel is estimated by using the path geometry and the refractivity profile of the lower troposphere. Thus, the ray powers are is calculated with bistatic radar equation [29]

$$P_r = \frac{P_t G_t G_r \sigma_V \lambda^2 \varrho}{(4\pi)^3 R_t^2 R_r^2}, \quad (2.1)$$

where P_t is the transmit power, $G_{t,r}$ are the transmitter and receiver antenna gains, ϱ is the polarization mismatch factor, σ_V is the differential scattering cross-section, and $R_{t,r}$ is the distance between the scattering point and transmitter/receiver.

In the ray-based approaches, the atmospheric profiles are included with the scattering cross-section. In [29], the scattering cross-section is developed by using Rayleigh scattering approximation [12]. To model the atmospheric turbulence characteristics, the Kolmogorov spectrum technique is utilized [12], and the refractivity profile of the lower atmosphere is generated by using real world measurements. Therefore, the ray-based methods can generate the realistic estimates of the PDS of the channel. The PDS of the channel can be utilized to predict important channel parameters such as delay spread, coherence bandwidth, coherence distance and the correlation between antennas for different diversity techniques.

2.3 Diversity Techniques

The troposcatter communication introduces high losses at b-LoS distances as discussed in Section 2.2, and the fluctuating atmospheric turbulence results in short

term fading. The effects of the fading becomes worse with the mobile b-LoS employments. Therefore, the employment of diversity techniques is required to alleviate the effects of the short term fading. To this end, we discuss the possible diversity techniques which are suitable for the troposcatter communications.

2.3.1 Space Diversity

In the troposcatter systems, the space diversity can be provided with either horizontally or vertically spaced antennas. Although the vertical placement of large parabolic reflector antennas (2.4–3 m diameter) does not seem practical, the vertical placement of the antennas can provide four different troposcatter common volume because the elevation angles of the vertically spaced antennas differ as seen in Figure 2.2(a). For this reason, the employment of vertical space diversity may be preferable. Based on the available experimental studies [1, 8] and empirical recommendations by ITU-R [5] the required spacing for the space diversity in the troposcatter should be $70\lambda - 100\lambda$ for the antennas to have low correlation values.

2.3.2 Angle Diversity

For angle diversity, the same parabolic reflector antenna with two beams can be utilized. For angle diversity, both vertical and horizontal beam separations can be utilized. However, the amount of angle separation for the horizontal angle diversity is higher than the vertical case [1, 26]. Since the troposcatter communication power is highly sensitive to angle, the employment of vertical angle diversity is more promising. In addition, vertical angle diversity can generate four different troposcatter common volumes as in Figure 2.2(b). According to [26], vertical angle diversity provides uncorrelated antennas for 1 – 1.5 beam-width beam separations. Since there will be significant path difference between angle diversity antennas, the upper beam will have significantly higher path-loss values up to 10 dB [26]. Thus, the gains of the angle diversity systems will be lower compared to the space diversity, but the cost of the angle diversity is lower because it requires only one parabolic reflector compared

to space diversity.

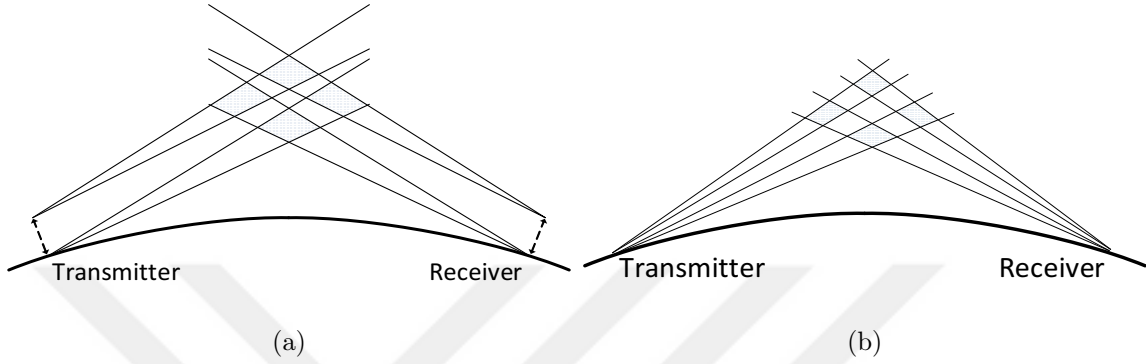


Figure 2.2: (a) Vertical space diversity and, (b) angle diversity.

2.3.3 Frequency Diversity

In frequency diversity, the signals are transmitted by different carrier frequencies that are adequately separated from each other. The required frequency separation is reported as 1% of the carrier frequency in [1]. However, experimentally measured results in [25] suggest that the troposcatter link with 192 km range and 4.7 GHz carrier frequency becomes uncorrelated at 6 MHz frequency separation. Frequency diversity receivers can be also mounted on the same parabolic reflector, but the utilization of the frequency spectrum increases with the frequency diversity systems.

2.3.4 Polarization Diversity

Polarization diversity can be provided with dual-polarized antennas without a need for additional parabolic reflector. In the white chapter provided by Comtech [1], it is pointed out that the vertical and horizontal polarizations have low correlation and the polarization diversity is valid for the troposcatter communications. Therefore, the polarization diversity can be utilized with space or frequency diversity to provide quad diversity in the troposcatter links. Therefore, the polarization diversity has high potential to improve the system performance.

The employment of diversity techniques in troposcatter links has high potential to increase the system performance. However, the comparison and analysis of different diversity techniques is required. Especially, the combination of diversity techniques may be preferable in the troposcatter links such as space-frequency, space-polarization diversity, frequency-polarization and etc. In addition, the performance of these diversity techniques under mobility is also an open research area. Thus, analytical models are required for these diversity techniques to investigate their correlation coefficients under different atmospheric and geographical conditions. As discussed in Section 2.2, the ray-based techniques are promising to calculate the correlation between antennas.

2.4 Application Areas

There is a significant demand for the high data rate b-LoS communication links in both military and civilian applications [22]. Therefore, we investigate the possible application areas of the troposcatter b-LoS communications in these categories.

2.4.1 Military Applications

Military communication applications are sensitive to many parameters because network-centric operations require timely and accurate transmission of the information between units at all the time with high reliability and security [22]. Therefore, military applications require low latency, high data rate, and high security. All of these concerns can be well-satisfied with the employment of troposcatter systems.

- **Latency:** The troposcatter communication provides near instantaneous point-to-point communication channel because the latency of the troposcatter channel is in the order of microseconds [1]. However, SATCOM applications may introduce 500 ms delays.
- **High data rate:** Modern troposcatter systems are able to provide 22 Mbps with high reliability [1, 2]. Thus, troposcatter communication can provide the data rate levels that are required for the military applications with low latency.

- **Security:** Troposcatter systems utilize low beam antennas. Therefore, the transmitted signal is sent to the atmosphere with very narrow beam. For this reason, the probability of detect/intercept is very low for the troposcatter channel. On the other hand, satellites transmit the signal to very large area. Therefore, SATCOM communication links can be determined by hostile units.

Thanks to these reasons, troposcatter is a promising candidate for the modern military applications where high data rate point-to-point communication is required. With the technological advancements, the troposcatter equipments are shrunk to the sizes that can be carried with the military trucks. Therefore, the mobile b-LoS links with troposcatter channel can be also considered as we analyzed in Section 3.7. We only consider the mobility on the 2D communication path. However, the mobility in the other directions requires the adaptive adjustment of the troposcatter antennas or adaptive troposcatter antenna array design. Thus, mobility in the troposcatter communications stand as a significant open research area.

2.4.2 Civilian Applications

In addition to the military applications, civilian applications also require high data rate b-LoS links. The most important civilian applications is off-shore gas and petroleum production platforms [30]. These platforms are generally located at the open seas where the line-of-sight microwave communication is not possible.

Off-shore platforms generally utilize underwater fiber or SATCOM to communicate with the land stations. However, underwater fiber is too expensive, but it provides highest capacity and highest communication quality compared to all other methods. High capacity SATCOM has high operational cost compared to underwater fiber and troposcatter because both troposcatter and underwater fiber has low operational costs. Even if the initial cost for the troposcatter systems are higher than SATCOM, troposcatter becomes cost effective in the long run compared to SATCOM at high data rate levels [30]. Therefore, the troposcatter is an also promising candidate for the civilian b-LoS applications.

Chapter 3

A RAY-BASED CHANNEL MODELING APPROACH FOR TROPOSCATTER

3.1 Introduction

To evaluate the performance of the troposcatter systems, it is required to investigate the characteristics of the troposcatter channel. However, available troposcatter channel models are inefficient to model the complex and time-varying air turbulence [4–6, 9, 10]. Therefore, we have introduced a ray-based troposcatter channel modeling approach in our preliminary work [29]. In our approach, the beam-width of each antenna is divided into small parts, and transmission-loss calculations are performed for each ray to estimate PDS of the channel. We introduce the differential scattering cross-section calculations by combining the available scattering studies [12, 32] and the meteorological refractive index structure constant studies [33, 34]. With this approach, we are capable of utilizing the real world water vapor mixing ratio measurements conducted by NASA [35] to consider atmospheric changes. In our preliminary work [29], this approach is proposed and simulated for different instantaneous bandwidths to show that troposcatter systems can achieve high data rates. In this chapter, we further extend our approach and analysis to derive the coherence bandwidth, RMS delay spread and correlation between vertically spaced antennas for the first time in the literature.

The main contributions of this chapter are four-fold. Firstly, we extend our proposed ray-based approach for realistic b-LoS troposcatter path geometries which considers antenna height and obstacles in the communication path. Secondly, the results of our proposed approach are compared with the ITU-R’s empirical model for predict-

ing annual and seasonal transmission losses in the troposcatter links [5] for verification. In addition, we provide link budget calculations for high frequency troposcatter systems. Thirdly, the coherence bandwidth and RMS delay spread of the channel are derived to estimate the fading behavior of the channel for given channel parameters for the first time in the literature. Lastly, we introduce vertical diversity and analyze the correlation between vertically spaced antennas for the first time. The simulation results are presented for the capacity of the MIMO troposcatter links by using real world measurements and our proposed ray-based approach.

The remainder of the chapter is organized as follows. Section 3.2 presents the related work. Section 3.3 provides troposcatter channel geometry. In Section 3.4, troposcatter transmission-loss calculations are introduced. In Section 3.5, we provide the comparison of the ITU-R's model and our approach and link budget calculations. Section 3.6 includes coherence bandwidth and RMS delay spread results. We provide the coherence time and bandwidth analysis in Section 3.7. In Section 3.8, the correlation between vertically spaced antennas is analyzed and capacity results are presented in Section 3.9.

3.2 Related Work

Available troposcatter studies generally focus on the prediction of the annual and the worst month transmission-losses with statistical methods [4, 5]. Although these statistical methods can capture the seasonal changes in the channel, fast variations and non-homogeneities in the channel cannot be modeled with such statistical approaches. In addition, [6] provided analytical expression for PDS of the troposcatter links. The analytical studies show that the envelope of PDS obeys the Rayleigh distribution. In addition, there exists some experimental studies [7, 8]. In [8], the correlation between vertically spaced antennas are presented with real world measurements, and the distribution of the fading in the troposcatter channel is reported as *Rayleigh*. Based on the analytical and experimental studies, there are also some studies that focus on the performance metrics in the troposcatter channels [9–11].

In the available analytical studies, the scattering is modeled only with the scattering angle [6]. For this reason, these models are inefficient to derive PDS, coherence bandwidth, and RMS delay spread which show significant dependence on the channel conditions. To this end, the main motivation of our work is to develop a comprehensive MIMO troposcatter channel modeling approach which considers the fluctuations and non-homogeneities in the air turbulence.

In our approach, the troposcatter is modeled with Rayleigh scattering approximation [12]. Since the scattering particles in the troposphere is much smaller than the wavelength of the signals, Rayleigh scattering approximation is valid for the troposcatter. To model the atmospheric turbulence, there are three major techniques: Booker-Gordon, Gaussian model and Kolmogorov spectrum technique. Booker-Gordon method uses variance in the exponential correlation however, the correlation is based on mathematics not the physics [12]. Furthermore, Gaussian model uses Gaussian correlation function. Therefore, these two models cannot effectively explain the troposcatter. However, Kolmogorov spectrum technique is based on the actual turbulence characteristics and it assumes the existence of locally homogeneous medium within the outer scale turbulence. Thus, this technique allows us to model the atmospheric changes with the real world water vapor mixing ratio measurements [35] in Section 3.4.

3.3 Troposcatter Geometry

In [29], we have utilized simple channel geometry which assumes that both of the antennas are placed on the ground. In this chapter, we extend this channel model to include the effects of antenna heights and obstacles in the b-LoS links.

Troposcatter channel geometry can be seen in Figure 3.1 and the description of the parameters that are used in Figure 3.1 can be found in Table 3.1. The remainder of the section covers the investigation of beam elevations and path length calculations.

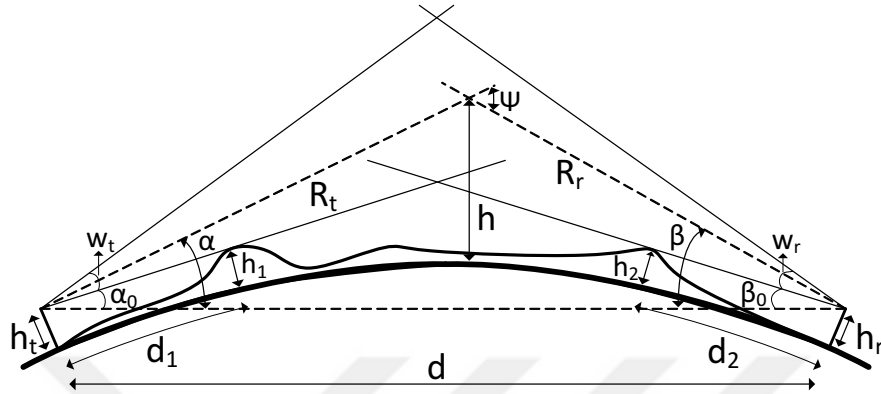


Figure 3.1: The geometry of the troposcatter link.

3.3.1 Beam Elevations Above Horizon

We define elevation angles with respect to horizontal line between transmitter and receiver (denoted as horizon). Instead of antenna elevation angles, we use *beam elevation* that is the elevation of the antenna beam with respect to horizon (α_0, β_0) . The proposed model divides the 3 dB beam-width of the antennas to small parts and calculates path lengths separately for each ray.

Beam elevations of the antennas are critical in the troposcatter systems. Even 1° increase in the elevation angle can result in 9 dB decrease in the received power due to both increase in the path lengths and increase in the height of the common volume [7]. As the beam elevations increase, the path lengths becomes longer and this effect results in higher transmission-loss. In the same way, increase in the beam elevations cause increase in the height of the common volume. Upper parts of atmosphere has less scatterers such as hydro-meteors. Therefore, the beam elevations should be as low as possible in order to achieve higher received powers.

In Figure 3.1, beam elevations also depend on the obstacles in the path and the lowest angle which cannot be blocked by any obstacle is called as *radio horizon*.

Table 3.1: The parameters in troposcatter geometry.

d	<i>Distance between TX/RX</i>
$h_{t,r}$	<i>Height of TX/RX</i>
$h_{1,2}$	<i>Height of obstacles</i>
$d_{1,2}$	<i>Distance between obstacle to TX/RX</i>
$R_{t,r}$	<i>Distance between scatter point and TX/RX</i>
α_0, β_0	<i>Beam elevation above horizontal line at TX/RX</i>
α, β	<i>Instantaneous beam elevation of the ray above horizon at TX/RX</i>
Ψ	<i>Scatter angle</i>
$w_{t,r}$	<i>Beam-width of the TX/RX</i>

Elevation angle of the antennas with respect to radio horizon is given as [4]

$$\theta_1 = \frac{h_1 - h_t}{d_1} - \frac{d_1}{2R_0}, \quad (3.1)$$

$$\theta_2 = \frac{h_2 - h_r}{d_2} - \frac{d_2}{2R_0}, \quad (3.2)$$

where R_0 represents the effective radius of the earth. Since low elevation angles are used in troposcatter systems, the curvature of the earth becomes important and the effective radius of the earth is given as

$$R_0 = k_e R_e \text{ km}, \quad (3.3)$$

where R_e is the earth radius (≈ 6370 km) and k_e represents the effective earth radius factor. Instead of using a median value for k as in [29], we calculate this parameter for real world conditions. The effective earth radius factor is given as [15]

$$k_e = \frac{157}{157 - \Delta N}, \quad (3.4)$$

where ΔN (N-units/km) is the average radio-refractive index lapse-rate through the lowest 1 km of the atmosphere and this value is derived by using real world data

sets [35] in Section 3.4.2. The resulting ΔN is calculated as 46 N-units/km on the average for a day in August [35]. By considering the monthly average value of 50 N-units/km that is provided by ITU-R for the same site at the same month [36], we can conclude that our calculations are consistent with the empirical results.

The beam elevations with respect to horizon for the minimum path lengths are found as [4]

$$\alpha_0 = \theta_1 + \frac{d}{dR_0} + \frac{h_1 - h_2}{d}, \quad (3.5)$$

$$\beta_0 = \theta_2 + \frac{d}{dR_0} - \frac{h_1 - h_2}{d}. \quad (3.6)$$

The resulting instantaneous beam elevation angles above horizon α, β (dashed lines in Figure 3.1) can be found by $\alpha = \alpha_0 + \Delta w$ where $\alpha_0 < \alpha < \alpha_0 + w_t$ and $\beta = \beta_0 + \Delta w$ where $\beta_0 < \beta < \beta_0 + w_r$. In our model, $1/\Delta w$ number of rays launch from the antennas to estimate PDS of the channel.

3.3.2 Path Lengths

The distance between the scattering point and transmitter/receiver ($R_{t,r}$) is calculated with the following relationship

$$\frac{d}{\sin(\Psi)} = \frac{R_t}{\sin(\beta)} = \frac{R_r}{\sin(\alpha)}, \quad (3.7)$$

where the scattering angle is given by $\Psi = \alpha + \beta$. By using (3.7), the total path length is given by $L = R_t + R_r$.

In order to find the delay spread of each ray, the difference between L and a reference path is required. Since troposcatter communication has no line-of-sight (LOS) component, the shortest path which is associated with the lowest possible beam elevation angles (α_0 and β_0) is assumed as the reference path. Therefore, the path length difference between the reference path and instantaneous path is used for delay calculations and formulated as [4]

$$\Delta d = L - d_0 = \frac{d}{2}(\alpha\beta - \alpha_0\beta_0) \text{ km}. \quad (3.8)$$

The delay spread of the reference path is assumed as the excess delay of the channel.

3.4 Troposcatter Path Loss

We introduce the differential transmission-loss calculations to estimate the PDS, coherence bandwidth and RMS delay spread of the troposcatter channel by using real world measurements. This section provides the scattering cross-section model and the differential radar range equation calculations for the troposcatter channel.

3.4.1 Scattering Cross-Section

Scattering cross-section represents the area in which the propagating waves can be scattered and depends on the wavelength of the rays, permittivity and most importantly the size and distribution of the scattering particles. Therefore, the variations in the air turbulence are modeled via scattering cross-section in the proposed approach. We utilize the existing scattering cross-section studies [12, 32] to derive differential scattering cross-section. In addition, we combine the scattering studies with the meteorological structure constant of the index of refraction studies [33, 34] to utilize real world measurements [35] for the first time in the literature.

The sizes of the scattering particles with respect to the wavelengths are critical because the type of the dominant scattering depends on this relationship. We model the clear air turbulence within this study. Therefore, the sizes of the scatterers in the troposcatter links are generally in the order of 1-10 μm whereas high frequency troposcatter systems is expected to use carrier frequencies between 1-10 GHz (wavelengths between 30-300 mm). Therefore, the wavelengths of the signals are much larger compared to the scattering particles. In such scenarios, the Rayleigh scattering approximation is valid for the scattering and it assumes that the electrical fields inside the scattering particles form constant vectors [12]. However, in case of rain drops scattering and higher frequency ranges, the validity of the Rayleigh scattering approximation becomes questionable [15] because the sizes of the scatterers and the wavelength becomes more comparable in such cases.

By using the Rayleigh scattering approximation, the differential scattering cross-

section is found in Appendix 13-A as

$$\sigma_V(\alpha, \beta, n) = 2\pi k^4 \cos(\Psi)^2 \Phi(\alpha, \beta, n) \text{ m}^2, \quad (3.9)$$

where $k = 2\pi/\lambda$ is the wavenumber and the fourth power dependence of the wavenumber comes from Rayleigh scattering approximation. Φ is the turbulence spectrum which is modeled with Kolmogorov theory [12]. Dependencies of parameters are included in formulas. α and β show dependency on the geometry, and n represents any dependency on the refractivity index and its gradient.

In the Kolmogorov spectrum technique, the clear air atmospheric turbulence is modeled via the turbulence eddies: the outer scale (Eddy size $> L_0$), and the inner scale turbulences ($L_0 > \text{Eddy size} > l_0$). The outer scale turbulence is anisotropic according to the climate conditions, and it is caused by the wind shear or the temperature gradient. The values of the L_0 may vary 10 – 100 m [37]. For the inner scale turbulences, the turbulence is isotropic and the energy of the eddies dominates the spectrum [12]. Therefore, the spectrum is proportional with $\Phi(k_s) \approx k_s^{(-11/3)}$, where $k_s = 2k \sin(\Psi/2)$. The microwave propagation is mostly affected by the inner scale turbulence [37], and the resulting Kolmogorov spectrum is formulated as [12]

$$\Phi(\alpha, \beta, n) = 0.033 C_n(\alpha, \beta, n)^2 (2k \sin(\Psi/2))^{-11/3} \text{ m}^3, \quad (3.10)$$

where C_n represents the structure constant of the index of refraction which is given as [33]

$$C_n(\alpha, \beta, n)^2 = 2.8 \left(\frac{\partial n}{\partial z} \right)^2 L_0(z)^{4/3}, \quad (3.11)$$

where n represents the index of refraction, and z represents the height. L_0 is the outer scale of the turbulence that represents the distance in which the atmosphere can be assumed as homogeneous. Experimentally validated values are used for this parameter as provided in [34]

$$L_0(z) = \frac{5}{1 + \left(\frac{z - 7500}{2000} \right)^2}. \quad (3.12)$$

3.4.2 Refractivity Calculations

We calculate C_n^2 by using both real world water vapor mixing ratio measurements that are provided by NASA Langley Research Center [35] and the ITU-R refractivity model which estimates the average refractivity profile of the troposphere [36]. We utilize the real world measurements to verify our model for a specific climate region, and the ITU-R's model is also utilized to show that the proposed method is able to predict the transmission-losses in other climate types as well.

With Real World Data set

One of the main contributions of our approach is the utilization of the real world water vapor mixing ratio measurements [35] because the proposed technique can take real world measurements as input instead of the statistical methods [9, 10]. Since C_n^2 depends on the index of refraction, we first find the vertical refractivity profile based on water vapor mixing ratio measurements. The fluctuations in the index of refraction is calculated using the simple relationship between them [36]

$$\begin{aligned} N &= (n - 1) \times 10^6, \\ \frac{\partial N}{\partial z} &= \frac{\partial n}{\partial z} \times 10^6 \text{ N-units/m.} \end{aligned} \quad (3.13)$$

For microwave frequencies, the refractivity of the air is formulated as [36]

$$N = \frac{77.6}{T} (P + 4810 \times e/T) \text{ N-units,} \quad (3.14)$$

where P represents total atmospheric pressure in mb, T represents the temperature of the air in Kelvin and, e represents vapor pressure that can be represented as

$$e = \frac{P}{\varepsilon} \omega \text{ mb,} \quad (3.15)$$

where ω represents the water vapor mixing ratio of the air that is the amount of water in air in g/kg. ω is directly taken from height dependent measurements. The molecular mass ratio of water vapor to air (ε) is taken as 0.622 [31].

By using (3.11), (3.12), (3.14) and (3.15), the C_n^2 is calculated as $3.08 \times 10^{-12} \text{ m}^{-2/3}$ on the average for the utilized data set [35]. The experimentally measured results for C_n^2 in microwave frequency range vary between $1 \times 10^{-12} - 6 \times 10^{-12} \text{ m}^{-2/3}$. Thus, the calculated structure constant of the index of refraction results are consistent with the real world measurements.

In this study, we consider the clear air atmospheric turbulence under standard atmospheric conditions. Therefore, we do not consider the effects of the atmospheric ducts. We also analyze the generated vertical refractivity profiles to detect any kind of duct formations. According to analysis, we do not observe any duct formation for the used data set [35]. However, the presence of atmospheric ducts may cause power loss due to its trapping effects [22].

With ITU-R Model

With the ITU-R refractivity model [36], the reference refractivity of the atmosphere can be calculated as

$$N = N_0 \exp(-z/z_0), \quad (3.16)$$

where z is the height in m, N_0 is the average value of the atmospheric refractivity at sea level in N-units, and z_0 is the scale height in m. [36] provides the average values for the N_0 for 5-year data set, but only one value is specified for z_0 as 9.5 km. Therefore, the result of the ITU-R refractivity model [36] will have less accuracy to model the atmospheric turbulence.

Figure 3.2 presents sample refractivity results that are calculated with the real world data set and ITU-R model. As noticed, the ITU-R model is just the reference curve for the refractivity. However, both curves becomes closer after 3 km. For this reason, we utilize the ITU-R refractivity model to generate the transmission-loss results for high ranges in Section 3.5.2 because the high ranges form high troposcatter common volumes where the ITU-R model can give reasonable results for the refractivity of the troposphere.

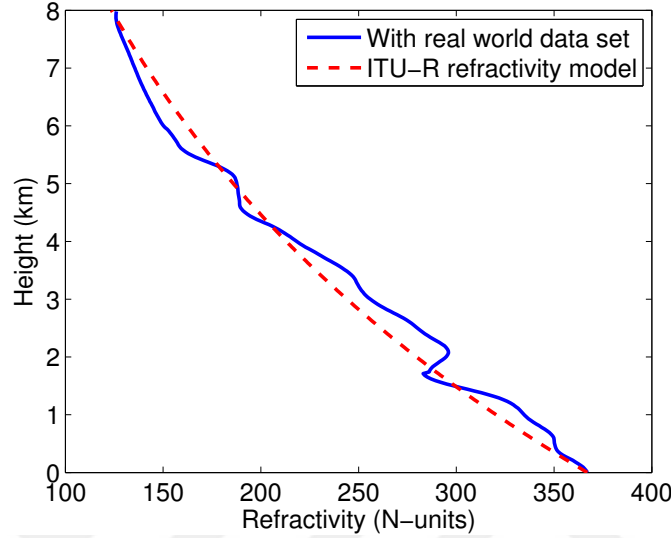


Figure 3.2: The geometry of the troposcatter link.

3.4.3 Troposcatter Antennas

Since troposcatter channel suffers from high path loss due to the long path lengths and scattering, low beam-width reflector antennas having 2.4 to 3 m diameters are used in troposcatter communication. As a result, the antennas can provide high antenna gains thanks to high apparatus areas. In addition, low beam-widths provide signals to spread less, so that the resulting delay spread and path loss are lower.

Since troposcatter antennas use low beam-widths around 3 to 10 mrad according to the path and atmospheric conditions, the type of the antenna has low effect in the simulations [5]. Therefore, the troposcatter antennas are modelled as isotropic antennas and the antenna gains can be found by [24]

$$G_{t,r} = 10 \log \left[\eta_{ant} \left(\frac{\pi D_{ant}^2}{\lambda} \right) \right] \text{ dBi}, \quad (3.17)$$

where η_{ant} is antenna efficiency, D_{ant} is antenna reflector diameter. (3.17) yields antenna gain of 41.5141 dBi for the carrier frequency of 4.7 GHz, antenna efficiency of 0.65 and antenna diameter of 3 m that are used in this work.

3.4.4 Polarization of The Scattered Waves

Polarization has substantial influence in the scattering process because randomness of the troposcatter common volume generally makes the scattered signal to have elliptically polarized notwithstanding its previous polarization [12]. Therefore, the polarization of the scattered rays is nearly random due to the randomness of the air turbulence.

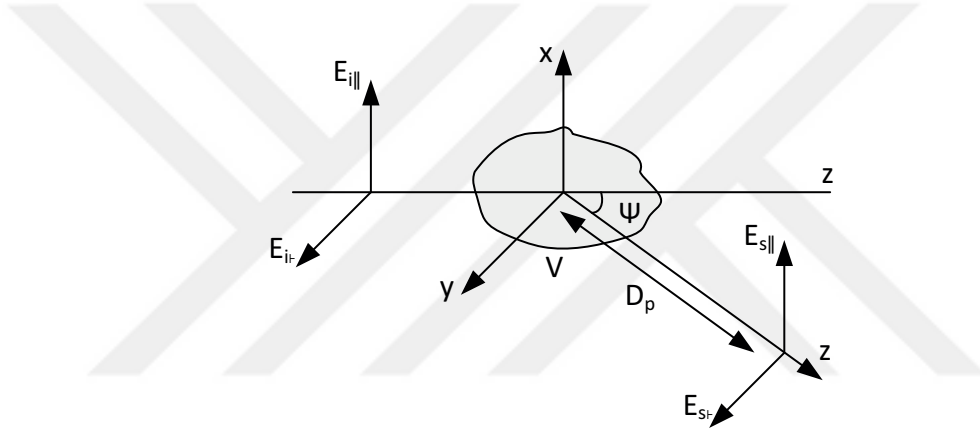


Figure 3.3: Geometry of the scattering

Since the scattering particles are much smaller in size with respect to the wavelength, the shape of the scattering particles are assumed to have spherical shapes. Therefore by using Rayleigh scattering approximation for 2D troposcatter links, the polarization of scattered waves at some distance from the scattering point can be determined by Stoke's matrix that can be represented as [12]

$$\begin{pmatrix} E_{s\perp} \\ E_{s\parallel} \end{pmatrix} = \begin{pmatrix} e^{ikr} \\ R \end{pmatrix} \begin{pmatrix} \cos(\Psi) & 0 \\ 0 & 1 \end{pmatrix} \begin{pmatrix} E_{i\perp} \\ E_{i\parallel} \end{pmatrix} \quad (3.18)$$

where E_s represents the polarization of the signal at a distance R from the scattering point and E_i represents the polarization of the wave at the scattering point as shown in Figure 3.3.

The polarization mismatch factor between the receiver antenna and the received ray can be determined by

$$\varrho = |\hat{e}_s \hat{e}_r|^2, \quad (3.19)$$

where $\hat{e}_r$ is the polarization vector of the receiver antenna and $\hat{e}_s$ is the polarization vector of the scattered wave that can be determined by (3.18).

3.4.5 The Radar Range Equation

Since the transmission-loss of each ray is calculated separately with the proposed ray-based approach, we define differential bistatic radar range equation as

$$P_r(\alpha, \beta, n) = \int_{dV_c} \frac{P_t G_t G_r \sigma_V(\alpha, \beta, n) \lambda^2 \varrho}{(4\pi)^3 R_t(\alpha, \beta)^2 R_r(\alpha, \beta)^2} dV, \quad (3.20)$$

where P_t is transmit power, G_t and G_r are antenna gains for transmitter and receiver, respectively. ϱ represents polarization mismatch factor which is used directly as introduced in [29]. Since the parameters such as range, obstacle height and frequency are fixed in the differential calculations, P_r is shown only as a function of geometry and refractive index. dV_c is differential scattering volume which is given as

$$dV_c(\alpha, \beta) = \frac{R_r^2 R_t^2 (dw)^2}{\sqrt{R_r^2 + R_t^2} \sin(\Psi)} \text{ m}^3, \quad (3.21)$$

where dw is the differential beam-width of the ray.

The Kolmogorov spectrum assumes that the atmospheric turbulence as locally locally homogeneous. Therefore, the scattering cross-section is assumed constant for the scattering volume dV_c , and the differential bistatic radar range equation is simplified as

$$P_r(\alpha, \beta, n) = \frac{P_t G_t G_r \sigma_V(\alpha, \beta, n) dV_c(\alpha, \beta) \lambda^2 \varrho}{(4\pi)^3 R_t(\alpha, \beta)^2 R_r(\alpha, \beta)^2} \text{ W}. \quad (3.22)$$

By combining (3.9), (3.10), (3.11), (3.21), and (3.22) we derive the bistatic equation as

$$P_r(\alpha, \beta, n) = \frac{P_t G_t G_r \pi^2 dw^2 \cos(\Psi)^2 \left(\frac{\partial n}{\partial z}\right)^2 L_0(z)^{4/3} \varrho}{21.64 \lambda^2 \sqrt{R_r^2 + R_t^2} \sin(\Psi) (2k \sin(\Psi/2))^{11/3}} \text{ W}, \quad (3.23)$$

where $\partial n / \partial z$ is directly calculated by using the real world measurements. $\sqrt{R_r^2 + R_t^2}$ term shows the linear relationship between the range and the transmission-loss. The same result also appears in the ITU-R's empirical model [5].

3.4.6 Simulation Results

PDS of the channel can be generated by using our ray-based approach. Figure 3.4 presents both PDS, and transmission-loss (P_r/P_t) distribution of multipath components for a measurement instant with the following channel parameters: $f = 4.7$ GHz, $d = 150$ km, $h_{t,r} = 27$ m, $h_{1,2} = 41$ m, $d_{1,2} = 21$ km and $G_{t,r} = 41.5$ dB. According to Figure 3.4(a), the distribution of the transmission-loss of the multipath components is close to Rayleigh distribution. This result is mentioned in [6,8–10], but it was never quantitatively shown. In this work, we derive this conclusion quantitatively for the real world conditions for the first time in the literature.

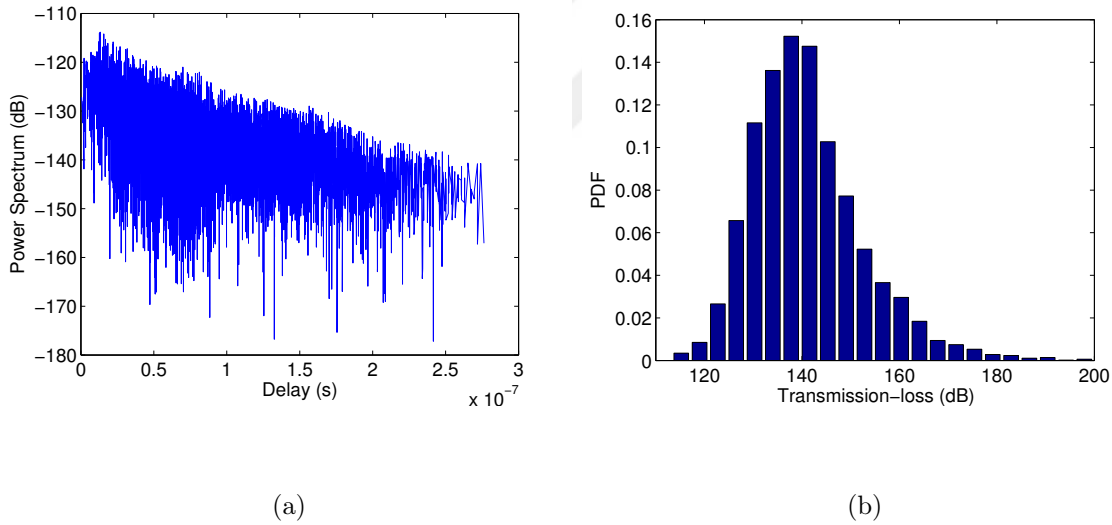


Figure 3.4: (a) PDS spectrum and, (b) histogram of multipath components for a 150 km troposcatter link for 4.7 GHz.

3.5 Path Loss and Link Budget Calculations

To verify our ray-based approach, we utilize ITU-R P.617 annual and average transmission-loss prediction method which is introduced in [5]. In [5], ITU-R provides the empirical model for predicting transmission-loss values for non-exceedance percentages of the

time. The equation and parameters of the ITU-R's model show changes according to climate zones such as dry, tropical, etc. According to the climate map in [5], the data set that we are using is in category 3 (dry climate). In addition, we provide link budget calculations for the troposcatter links.

3.5.1 ITU's Model for Troposcatter Path Loss

In [5], ITU provides the empirical model for predicting path loss values for non-exceedance percentages of the time. Non-exceedance percentage of the time q means that the path loss of the channel does not exceed the resulting empirical value for $q\%$ of the time. ITU's model is more accurate for 50% non-exceedance and above. The equation and parameters of the ITU's model show changes according to climate zones such as dry, tropical, etc. According to the climate map in [5], the data set that we are using is in category 3 (dry climate). For this reason, the presented formulas and used parameters are belong to the climate category 3.

The path loss dependence L_N on the height of the common volume is found as

$$L_N = 20 \log(5 + \gamma H) + 4.34\gamma h \text{ dB}, \quad (3.24)$$

where $H = 10^{-6}\theta d/4$ km and $h = 10^{-9}\theta^2 R_0/8$ km where γ is the atmospheric structure parameter which is used as 0.32 for climate 3 [5]. $\theta = d/R_0 + \theta_1 + \theta_2$ where θ_1, θ_2 can be found by (3.1) and (3.2).

Conversion factor in dB for $q = 90\%$ is given as

$$Y(90) = \begin{cases} -11.5, & d_s < 100 \\ -8.519 \times 10^{-8}d_s^3 + 7.444 \times 10^{-5}d_s^2 + 4.18 \times 10^{-4}d_s - 12.1, & 100 \leq d_s < 465 \\ -8.4, & \text{otherwise,} \end{cases} \quad (3.25)$$

where $d_s = \theta R_0/1000$ km.

The aperture-to-medium coupling loss L_c is found by

$$L_c = 0.07 \exp[0.055(G_t + G_r)] \text{ dB}, \quad (3.26)$$

where $G_{t,r}$ are the transmitter and receiver antenna gains in dB, respectively.

The average annual path loss not exceeded for $q\%$ of the time can be found as

$$\begin{aligned} L(q) = & M + 30 \log f_{MHz} + 10 \log d_{km} + \\ & L_N + L_c - G_t - G_r - Y(q) \text{ dB}, \end{aligned} \quad (3.27)$$

where f_{MHz} is the frequency in MHz and M is the meteorological parameter which is taken as 19.30 dB. $Y(q) = Y(90)C(q)$, where for 50% non-exceedance $C(50) = 0$, and for 90% non-exceedance $C(90) = 1$ [5].

ITU's model is most accurate for 50% non-exceedance and above. However, [5] also indicates that the path loss can be considered symmetrical to find lower percentages. The accuracy of lower percentages will not be high, however it can be used to give a loose lower bound for the average path loss. By using the formula provided in [5], average annual path loss not exceeded for 10% of the time can be given as

$$L(10\%) = L(50\%) - [L(90\%) - L(50\%)] \text{ dB}. \quad (3.28)$$

3.5.2 Comparison of Our Model and ITU's Model

To find the transmission-loss results with our proposed approach, the transmission-loss of the link is calculated for each measurement instant. There are 1000 measurement instants that are measured in August near Cape Verde Islands [35]. The average of these transmission-loss values for different times is denoted as the transmission-loss of the link for a given range. Since ITU-R's model is most accurate below 5 GHz, we compare the results only for 4.7 GHz link with respect to range. The following channel parameters are utilized in the simulations: $h_{t,r} = 27$ m, $h_{1,2} = 41$ m, $d_{1,2} = 21$ km and $G_{t,r} = 41.5$ dB. The maximum LoS range can be found with the generic formula $d_{LOS} < 4.12\sqrt{h_t} + 4.12\sqrt{h_r}$. With our channel parameters, LOS range can reach up to 42 km. Beyond this range is denoted as b-LoS range. Therefore, we focus on the ranges higher than 50 km in this subsection.

To verify our results, we compare the transmission-loss results of ITU-R's model and our proposed approach. We derive the PDS of the troposcatter channel by using

troposcatter geometry and transmission-loss calculations introduced in Section 3.3 and 3.4. In Figure 3.5(a), the average of 1000 instants for each range and ITU-R's model for different non-exceedance percentages are presented. Since the transmission-loss results that are generated with our model are the average of the time instants, it is expected that our proposed approach should be close to the 50% not exceeded of the time. In Figure 3.5(a), our model is very close to the 50% curve.

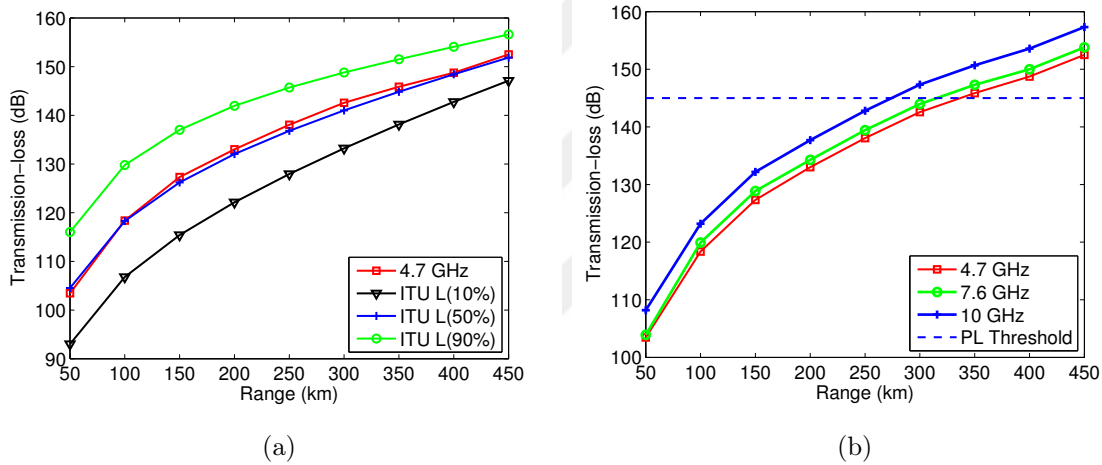


Figure 3.5: Transmission-loss vs. range (a) Comparison of our model and ITU-R's model for 4.7 GHz, (b) Transmission-loss for different frequencies.

3.5.3 Link Budget Calculations

To determine the maximum b-LoS communication range that can be provided by the troposcatter communication, we provide troposcatter link budget calculations by using our ray-based approach. The channel parameters that we utilize can be found in Table 3.2.

Received power in dBm is found as

$$P_{r\text{dBm}} = P_{t\text{dBm}} - L_{dB} - L_{q\text{dB}}, \quad (3.29)$$

where L_{dB} is the path loss.

Table 3.2: Link budget channel parameters.

Parameter	Value	Unit
<i>Transmit power</i> (P_t)	60	dBm
<i>Antenna diameter</i> (D_{ant})	3	m
<i>Antenna gains</i> ($G_{t,r}$)	41.5	dB
<i>Cable losses</i> (L_q)	5	dB

For modern troposcatter equipment, the minimum received power should be -90 dBm [24, 38]. To satisfy the minimum of -90 dBm received power, the maximum path loss is calculated as 145 dB by using (3.29) and channel parameters in Table 3.2. Figure 3.5(b) shows the path loss for different frequencies and the calculated 145 dB path loss threshold (PL threshold). As noticed, 4.7 GHz can provide communication link up to 300 km. However, this path losses are in the order of 50% not exceeded of the time as shown in Figure 3.5(a). If we look at the losses in the Figure 3.5(a), 4.7 GHz can achieve the communication 90% of the time up to 250 km. As a result, troposcatter communication can be used in b-LoS links up to 250 km with very high availability rates. Since high frequencies can achieve lower ranges, the utilization of lower frequencies is more preferable. Therefore, there is a trade-off between range and available bandwidth of the channel because higher frequencies can provide higher bandwidth.

3.6 Coherence Bandwidth and RMS Delay Spread

The fading characteristics of the channel is another important design constraint for the b-LoS troposcatter communications. We examine the coherence bandwidth and RMS delay spread of the channel in order to determine whether the selected bandwidth experiences flat fading or frequency selective fading according to the channel conditions.

The fading characteristics of the channel can be analyzed with the frequency domain representation of PDS which is given as [39]

$$\phi_c(\Delta f; \Delta t) = \int_{-\infty}^{\infty} \phi_c(\tau; \Delta t) e^{-j2\pi\Delta f\tau} d\tau, \quad (3.30)$$

and associated 3 dB point of $\phi_c(\Delta f; \Delta t)$ is denoted as coherence bandwidth.

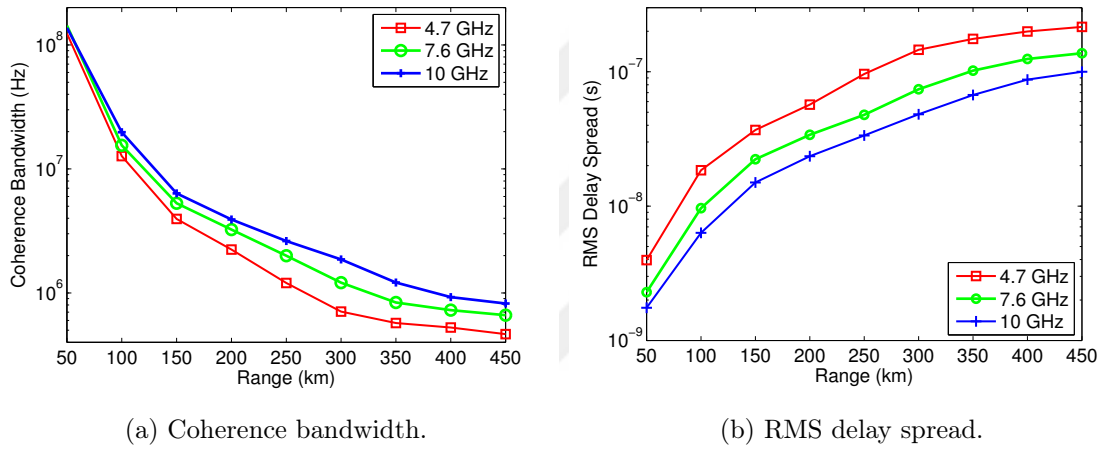


Figure 3.6: Coherence bandwidth and RMS delay spread of troposcatter link vs. range.

Figure 3.6(a) represents the coherence bandwidth vs. range for different carrier frequencies for the following channel parameters: $h_{t,r} = 27$ m, $h_{1,2} = 41$ m, $d_{1,2} = 21$ km, $G_{t,r} = 41.5$ dB dB and $P_t = 1000$ W. Here, there is a direct relationship between carrier frequency and coherence bandwidth of the channel because the beam-width of the parabolic reflector antennas depend on the carrier frequency, i.e., $w_t = w_r = 70\lambda/D_{ant}$, where D_{ant} is the antenna diameter. The same effect can be seen in RMS delay spread results that are presented in Figure 3.6(b). Since the decreasing beam-width of the antennas decreases the troposcatter common volume, the delay spread of the channel becomes less. According to these results, 150 km b-LoS troposcatter link has 4 MHz coherence bandwidth and 1.85×10^{-8} s RMS delay spread for 4.7 GHz. Therefore, 1 MHz bandwidth for 150 km link at 4.7 GHz experiences

flat fading. However, at higher ranges the channel may become frequency-selective at these bandwidth levels and the implementation of orthogonal-frequency-division-multiplexing systems would be preferable to avoid frequency-selective effects of the channel.

3.7 Coherence Distance/Time

With the recent advancement in the troposcatter equipments, there are transportable troposcatter equipments which can be mounted on the vehicles which have approximately 2.4 m parabolic reflector antennas. The current literature only considers the fixed troposcatter communications, but especially military applications may require b-LoS communication link on mobile. Therefore, the main aim of this section is to present the coherence distance of the channel to calculate the coherence time of the channel for different speeds.

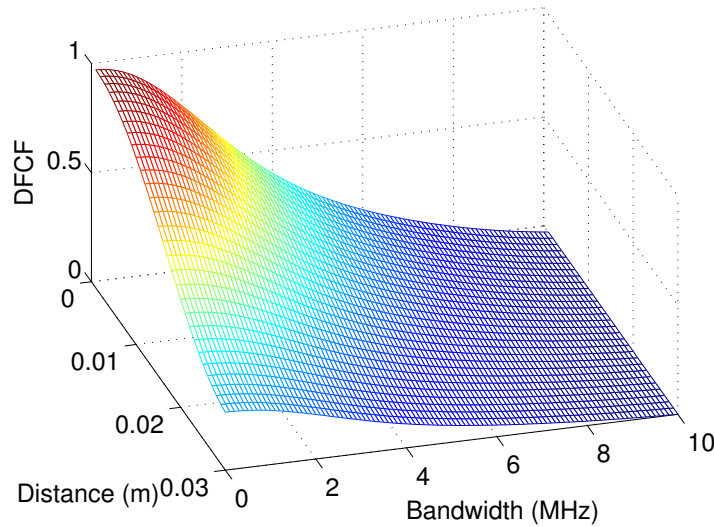


Figure 3.7: DFCF of the troposcatter channel.

We utilize ray-based methods to find the distance-frequency correlation function (DFCF) for troposcatter links. To this end, we utilize [29] to calculate DFCF. The DFCF is calculated by determining the correlation between PDSs of the troposcatter

links for varying distance and frequency. In [29], 2D troposcatter channel is modeled. Therefore, we assume that the movement will occur on the communication path, thus we slightly increase the range and calculate the correlation between PDSs to calculate the distance correlations. For frequency correlation, we slightly increase the carrier frequency and calculate the correlation between PDSs. At the end, we repeat these steps for both parameters to find the DFCF of the channel. The resulting DFCF of the troposcatter channel can be seen in Figure 3.7 for the following channel parameters: 4.7 GHz carrier frequency, 150 km range, 1.5° beam-widths at both transmitter/receiver and both of the antennas are pointed to horizon as discussed in [29].

The fluctuations in the troposcatter channel can be both caused by the fluctuations of the atmospheric turbulence or the movement of the receiver/transmitter. To this end, the Doppler spread of the fixed troposcatter link is measured in [28], and the Doppler spread for the troposcatter channel is reported as 15 Hz at maximum. Thus, the main source of the Doppler spread will be the speed of the vehicle for the troposcatter links because the Doppler spread due to mobility will dominate the atmospheric changes. For this reason, we assume steady atmospheric conditions in this chapter for simplicity.

Since the troposcatter communications utilize very narrow beams ($0.6^\circ - 1.5^\circ$), the Doppler spread due to mobility is expected to be low because the troposcatter communications is also utilize very low beam elevation angles. Doppler spread (D_s) is given as [27]

$$D_s = \frac{v}{\lambda} \cos(\theta), \quad (3.31)$$

where v is the speed of the vehicle, λ is the wavelength and θ is the angle between the received ray and horizon.

In [29], the beam elevation angles of 0.47° is utilized for over sea conditions. For 1.5° beam-width, θ will vary between 0.47° and 1.97° for the troposcatter link. Therefore, $v = 10$ m/s results in ≈ 157 Hz for this angle range by (3.31). As noticed from Figure 3.7, the coherence distance for the channel (D_c) can be determined as 0.03

m where the distance correlation function is low. The coherence time of the channel ($T_c = D_c/v$) for $v = 10$ m/s becomes 3 ms. The relationship between coherence time and Doppler spread can be found with the well-known relationship as

$$T_c = \alpha \frac{1}{D_s}. \quad (3.32)$$

From the ray-based simulation, we calculate the coherence time as 3 ms. The Doppler spread of the channel is calculated as 157 Hz. Therefore, the resulting α becomes 0.47. The coherence time of the channel for different beam-widths and speeds can be directly calculated by (3.32) without any time consuming ray-based simulations.

In this analysis, the coherence time of the channel is calculated as 3 ms which is much lower compared to the signal times at the MHz bandwidth levels. Therefore, the troposcatter channel will have slow fading at these mobility levels, and the troposcatter communication can be performed under mobility conditions. However, the received signal power will decrease as the range increase and the adaptive orientation of the antennas may be required to provide a reliable link. For these reasons, the mobility in the troposcatter links is an important open research area.

3.8 Vertical Diversity

Modern b-LoS communication requires reliability and high data rates. However, troposcatter introduces high losses due to the changes in the atmospheric conditions. Therefore, troposcatter b-LoS systems require the employment of MIMO systems to increase the reliability and the maximum data rate that can be provided. In this section, we investigate the utilization of vertically placed antennas and their correlations as a function of vertical distance between them.

2×2 vertical diversity provides four different common volumes as in Figure 3.8 because increasing antenna heights decreases beam elevation of the antennas. However, the separation between antennas should be kept small for practical reasons. In this case, different common volumes will intersect with each other and this effect introduces correlation between vertically spaced antennas.

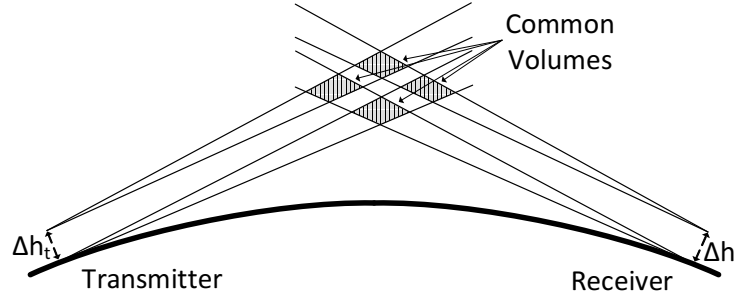


Figure 3.8: The geometry of the vertical diversity in troposcatter.

The employment of more receiver antennas than transmitters is more advantageous and common in modern systems. For this reason, we determine the correlation between receiver antennas by modifying our ray-based algorithm. We assume that transmitters are uncorrelated with enough spacing between them. To find the correlation between receivers, we exclude the effects of atmospheric profiles. Each ray is assumed to reach the receiver with the same power as in [40] because our main aim is to find the correlation between antennas which can be generalized for all climate conditions. Thus, ray powers between transmitter i and receiver j is calculated as

$$P_{ri,j}(\alpha, \beta) = \exp \left(-j \frac{2\pi}{\lambda} (R_{ti,j}(\alpha, \beta) + R_{ri,j}(\alpha, \beta)) \right), \quad (3.33)$$

where $R_{ti,j}$ and $R_{ri,j}$ are the distances between the scattering point with transmitter i and receiver j , respectively. The resulting correlation $\rho_{k,l}$ between receivers k, l is found by the correlation between PDSs $\phi_{i,k}(\Delta t)$ and $\phi_{i,l}(\Delta t)$ which are calculated by using (3.33).

Figure 3.9 presents correlation vs. vertical height difference over wavelength $(\Delta h/\lambda)$ in vertical diversity troposcatter systems. The increase in the carrier frequency increases the correlation at the same ratio. However, since the wavelength of the higher frequencies are lower, increase in the carrier frequency does not increase the required spacing for a given correlation value. To verify our results, we compare our results with the empirical recommendations provided by ITU [5]. The safe vertical

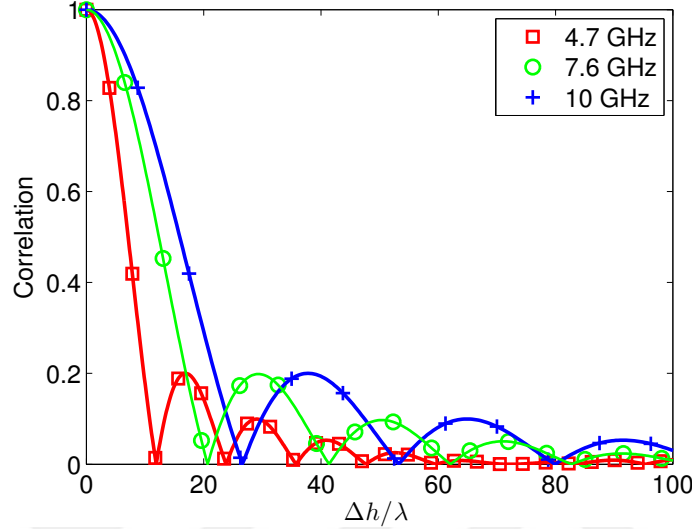


Figure 3.9: Correlation vs. vertical height difference for receiver.

spacing for antennas to be uncorrelated is given as [5]

$$\Delta h_v = 0.36(D_{ant}^2 + 4I_v^2)^{1/2}, \quad (3.34)$$

where D_{ant} is antenna diameter and I_v is the empirical scale length in vertical direction which can be taken as 15 m [5]. According to (3.34), the resulting spacing between antennas should be 10.85 m which is $170\Delta h/\lambda$ for 4.7 GHz. According to our results, the correlation is almost zero for $80\Delta h/\lambda$ and above for 4.7 GHz. ITU's empirical results give safe spacing between antennas to be uncorrelated, and our results are consistent with this empirical result. Therefore, this consistency verifies the correlation results of our channel modeling approach.

3.9 Capacity Results for Troposcatter Vertical Diversity

We investigate the performance of b-LoS troposcatter communication in terms of capacity to show that b-LoS troposcatter communication is suitable for high data rate applications. For this reason, this section first includes capacity formulations by using our approach and simulation results for MIMO troposcatter systems.

3.9.1 Capacity Formulation

The system use 1 MHz bandwidth for 4.7, 7.6 and 10 GHz carrier frequencies. For this bandwidth, the channel experiences flat-fading as described in Section 3.6. Flat fading channel is modeled as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}, \quad (3.35)$$

where $\mathbf{H}$ is the channel gain matrix $N_R \times N_T$, where N_T and N_R are the number of transmitter and receiver antennas, respectively. $\mathbf{x}$ and $\mathbf{y}$ are transmitting and receiving vectors with dimensions $N_T \times 1$ and $N_R \times 1$.

In our approach, the changes with time is modeled with real world measurements. We use channel matrix normalization in order to model the effect of large-scale path loss that changes with time due to atmospheric effects. The large-scale path loss is derived by using real world measurements as described in Section 3.4. Small-scale fading is assumed as Rayleigh fading as described in Section 3.4.6. Normalized channel gain matrix $\tilde{\mathbf{H}}$ is found as [41]

$$\tilde{\mathbf{H}} = \frac{\mathbf{H}}{\sqrt{\frac{1}{N_T N_R} \|\mathbf{H}\|_F^2}}, \quad (3.36)$$

where $\|\cdot\|_F$ is the Frobenius norm.

By using the channel matrix normalization, the average received power and the average Signal-to-Noise Ratio (SNR) on the channel is found as [41, 42]

$$P_{rav} = \frac{P_t}{N_T N_R} \|\mathbf{H}\|_F^2, \quad (3.37)$$

$$SNR_{av} = \frac{P_t}{\sigma_n^2 N_T N_R} \|\mathbf{H}\|_F^2, \quad (3.38)$$

where P_t is the transmit power, σ_n^2 is the noise power. P_{rav} is the average receive power that is derived with our ray-based approach with real world measurements. There are 1000 measurements in our data sets that are measured in August near Cape Verde Islands [35]. The noise power at the receiver is modeled as thermal noise $\sigma_n^2 = kT_a B$ where k is the Boltzman's constant, T_a is the antenna noise temperature.

We assume that there is no correlation between transmitters. However, receiver antennas are placed closer to each other and there is correlation between them as

analyzed in Section 3.8. With correlated receiver antennas, normalized channel gain matrix is modeled as [43]

$$\tilde{\mathbf{H}} = (\mathbf{R}_H^{RX})^{1/2} \mathbf{G}, \quad (3.39)$$

where $\mathbf{G}$ is the complex Gaussian iid channel gain matrix. $(\mathbf{R}_H^{RX})^{1/2}$ is the root of the receiver antenna correlation matrix $(\mathbf{R}_H^{RX})^{1/2}[(\mathbf{R}_H^{RX})^{1/2}]^H = \mathbf{R}_H^{RX}$. The root of a matrix can be found with Cholesky factorization.

In the channel capacity calculations, we utilize uniform power allocation to the all eigenvalues by assuming the power level of the channel is high enough. The resulting channel capacity matrix can be represented as [27]

$$C = \log_2 \det \left(I + \frac{SNR_{av}}{N_T} \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H \right), \quad (3.40)$$

$$C = \sum_{i=1}^{\min(N_T, N_R)} \log_2 \left(1 + \frac{\rho}{N_T} \lambda_i \right), \quad (3.41)$$

where $\gamma = SNR_{av}/N_T$, and λ_i are the nonzero eigenvalues of the Wishart matrix [44]

$$W = \begin{cases} \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H, & N_R \leq N_T \\ \tilde{\mathbf{H}}^H \tilde{\mathbf{H}}, & N_R > N_T \end{cases} \quad (3.42)$$

3.9.2 Simulation Results

The channel parameters that are used in our simulations can be found in Table 3.3. 1000 atmospheric measurements are used to estimate SNR_{av} in (3.40) by using proposed ray-based approach, and 1000 realization for complex Gaussian channel matrix $\mathbf{G}$ in (3.39) are performed for each measurement instant. Therefore, the total of 1 million realization is performed for each capacity results presented by using MATLAB simulations.

Cumulative distribution function (CDF) of capacity in 2×2 MIMO troposcatter link as a function of carrier frequency presented in Figure 3.10(a). Due to higher transmission-loss in higher frequencies, capacity decrease with increasing carrier frequency as expected. In addition, CDF of capacity for single-input-single-output

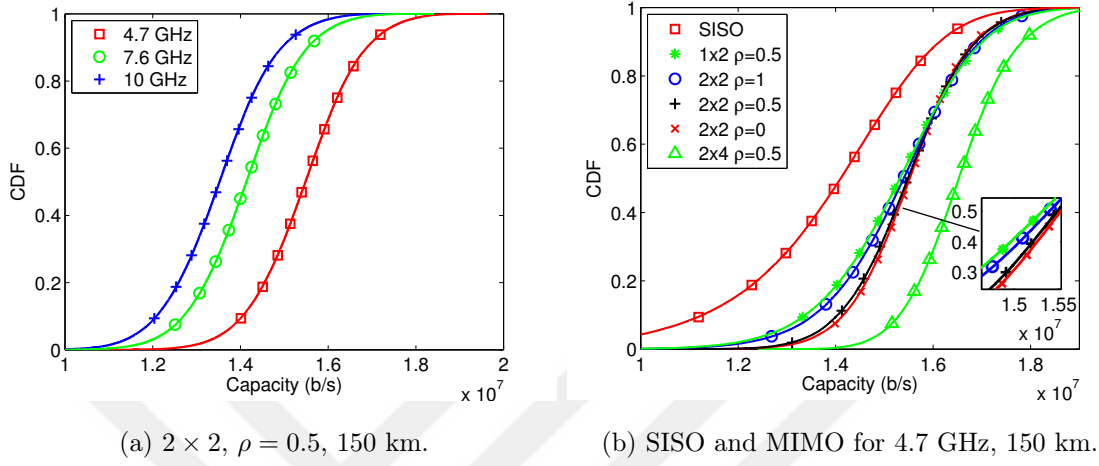


Figure 3.10: Capacity of CDF for different frequencies and antennas sizes for troposcatter communication

(SISO), single-input-multiple-output (SIMO) and MIMO employments for 4.7 GHz and 150 km presented in Figure 3.10(b). Both reliability and capacity of the link are increased with MIMO employments. As the correlation coefficient increase, the capacity gain is reduced. Increase in the number of receivers without changing the number of transmitters can increase achievable capacity. Therefore, the employment of more receiver antennas than transmitters is more preferable for systems requiring high data rates. Lastly, MIMO systems also decreases the span of achievable data rates, so that the link becomes more reliable.

Despite of the high losses that are introduced in troposcatter communications, SISO b-LoS troposcatter communication can provide high data rates more than 10 Mbps at 90% of the time for 1 MHz bandwidth. With the partially correlated ($\rho = 0.5$) MIMO employments the link can provide more than 10 Mbps at all times as in Figure 3.10(b). Therefore, our analysis show that troposcatter communication can be utilized for high data rate b-LoS applications with high availability.

Table 3.3: Troposcatter communication parameters.

Parameter	Value	Unit
<i>Horizontal distance (d)</i>	150	km
<i>Transmitter height (h_t)</i>	27	m
<i>Receiver height (h_r)</i>	27	m
<i>Distance between obstacle to TX/RX ($d_{1,2}$)</i>	21	km
<i>Height of obstacles ($h_{1,2}$)</i>	41	m
<i>Transmit power (P_t)</i>	1	kW
<i>Antenna gains ($G_{t,r}$)</i>	41.5	dBi
<i>Antenna noise temperature (T_a)</i>	290	K
<i>Antenna diameters (D_{ant})</i>	3	m

Chapter 4

FADING CORRELATION MODEL FOR MIMO-OFDM TROPOSCATTER COMMUNICATIONS

4.1 Introduction

Troposcatter medium is a lossy wave-guide due to high path lengths and scattering. Thus, the implementation of diversity techniques is required to provide reliable and high data rate b-LoS troposcatter systems. The main diversity techniques for troposcatter communications are space, frequency and angle diversity. In addition, the combination of diversity techniques may be utilized to increase the spectral efficiency with less cost. For example, the space-frequency diversity employs horizontally placed antennas utilizing frequency diversity. Since the frequency diversity receivers can be mounted on the same parabolic reflector, quad diversity can be achieved with only 2 antennas in each side. Therefore, the main objective of this chapter is to analyze the correlation between antennas in these diversity techniques to determine the proper spacings to achieve the desired gains for the first time in the literature.

The contribution of this chapter is twofold: Firstly, we develop an analytical model for the fading correlation of the troposcatter antennas utilizing space, frequency and angle and space-frequency diversity techniques for the first time. These diversity techniques are employed by the modern troposcatter systems [1], but there are no analytical fading correlation studies. Therefore, we develop the *Ring Scatter Model* (RSM) to analyze the fading correlation in troposcatter links and compare our results with the empirical values provided by the industry [1], ITU [5], and the experimental results [8,13,25,26,46,47]. In addition, one of the most important contributions of this work is the prediction of 50% coherence bandwidth based on the frequency separation

results. In this way, the fading behavior of the channel can be determined for a given channel parameters. Secondly, we compare the employment of different diversity techniques with the distribution of achievable data rates by extending our previously proposed ray-based channel model approach for troposcatter communications [29]. Our ray-based channel model is a ray-tracing method which calculates the power and delay of each ray to generate PDS. The novelty of our approach is to use the real world measurements for atmospheric modeling, thus each ray is subjected to different conditions according to its height (For more details, see [29]). In this work, we extend our approach to implement the diversity techniques and MIMO-OFDM (Orthogonal-Frequency-Division-Multiplexing).

The reminder of the chapter is organized as follows. Section 4.2 presents the related work. Section 4.3 provides the RSM fading model. In Section 4.4, the ray-based channel model is reviewed. Section 4.5 and 4.6 include the formulation and simulation results for the data rates for the diversity techniques, respectively.

4.2 Related Work

Although the fading correlation is well studied especially in mobile links [40, 48, 49], there is no theoretical study which considers the fading correlation in the troposcatter links. Therefore, our main aim is to analyze the fading correlation in different diversity schemes: space, frequency, angle, and space-frequency diversity techniques. On the other hand, there are empirical recommendations for these diversity techniques in order for the antennas to be uncorrelated [1, 5]. In addition, there are available experimental studies which considers the fading correlation in the troposcatter links. [8, 13] provides the fading correlation results for space diversity along with the fading statistic of the channel. According to their results, the channel shows Rayleigh fading under low integration times. [25, 46] focus on frequency correlation in the troposcatter links. The correlation and path-loss results for the angle diversity systems are provided in [26, 47]. We utilize these empirical suggestions and experimental studies to validate our results.

To model the fading correlation, there are several methods. For this purpose, the fading correlation can be determined by channel sounding experiments [49]. However, the cost of channel sounding experiments for troposcatter is too high due to the requirement for high power amplifiers and large antennas. Therefore, the correlation between antennas may be determined via analytical studies. This is the main reason for this chapter. In addition, one of the most popular fading correlation model is "One Ring" model, which is first introduced in [48]. In this model, the receiver is assumed to be surrounded by local scatterers and the correlation between the antennas are determined by using channel geometry calculations. However, the scatterers in troposcatter channel are located in the troposcatter common volume unlike the one ring model. Therefore, we develop the RSM for the fading correlation in the troposcatter system based on the methodology in [40, 48] by using the troposcatter channel geometry [4]. We extend RSM to analyze space, frequency, angle and space-frequency diversity fading correlation calculations.

4.3 Ring Scatter Model for Troposcatter Fading Correlations

To achieve the desired gains with MIMO, the antennas should have desired correlation values. Thus, we develop an analytical model, *ring scatter model* (RSM), to investigate the fading correlations in the main diversity techniques that are suitable for the b-LoS troposcatter communications: space, frequency and angle diversity. In addition, the combination of diversity techniques also has a potential to provide higher achievable data rates. Therefore, space-frequency diversity which is the combination of space and frequency diversity is introduced and investigated as well.

RSM fading correlation analysis are developed for single-user to single-user communication link. Troposcatter link has $M_{t,r}$ antennas at the transmitter and receiver sides, respectively. Therefore, the channel response in frequency domain ($\mathbf{H}_{M_r \times M_t}[k]$) can be represented as

$$\mathbf{r}[k] = \mathbf{H}[k]\mathbf{s}[k] + \mathbf{v}[k], \quad (4.1)$$

where $\mathbf{r}_{M_r \times 1}$ is the received signal vector, $\mathbf{s}_{M_t \times 1}$ represents the transmit vector, $\mathbf{v}_{M_r \times 1}$

is additive-white-Gaussian-noise (AWGN).

The remainder of the section includes the scattering model, and the RSM method for the diversity techniques.

4.3.1 Scattering Model

The troposcatter power depends on both the path geometry and the atmospheric turbulence. Therefore, both of these factors will have strong effects on the correlation analysis. The troposcatter is caused by the atmospheric scintillations due the the changes in the refractive index of the atmosphere. According to the turbulence characteristics, the scattering can be modeled as single or multiple scattering [12, 50]. The microwave propagation in the troposphere is related to the tenuous distribution of the particles, and it can be modeled with the first order multi-scattering approximation in the tenuous medium for a unit particle as [12, 50]

$$\frac{P_r}{P_t} = \frac{\lambda^2 G_t(\hat{i}) G_r(\hat{o})}{(4\pi)^3 R_t^2 R_r^2} \sigma_V(\hat{i}, \hat{o}, n) \exp(-\tau_t - \tau_r), \quad (4.2)$$

where λ is the wavelength, $P_{t,r}$ are the transmitter and receiver power. $G_{t,r}$ are the antenna gains which are modeled with Gaussian pattern. $R_{t,r}$ are the distances between scattering point to transmitter and receiver respectively. $\sigma_V(\hat{i}, \hat{o}, n)$ is the scattering cross-section, where $\hat{i}$ is the incoming ray vector, $\hat{o}$ is the scattered ray vector, and n is the index of refraction. $\tau_{t,r}$ are the optical distances for $R_{t,r}$ [12].

In correlation calculations, the constant terms can be omitted to simplify the model because they will have no effect: λ^2 and $(4\pi)^3$ in (4.2). In addition, the terms that have very slight changes can be eliminated. $R_t^2 R_r^2$ term in of (4.2) will also show very slight changes due to narrow beam-widths. For 250 km range and 1.5° beam-width, the ratio of the minimum path over maximum path (R_t^{Min}/R_t^{Max}) is found as 0.9992. Thus, this term can be eliminated to simplify the model. Furthermore, the $\tau_{t,r}$ are the path integral of the R_t and R_r over the index of refractions and and they represents the effect of the scattering particles between the antennas and the common volume. Since the expected antenna spacing in space diversity ($\approx 100\lambda \ll R$), and

no spacing for the frequency and angle diversity, the values for optical distances will be almost the same for each ray. Therefore, we eliminate the $\exp(-\tau_t - \tau_r)$ in (4.2) as well.

To model the atmospheric turbulence, we use Booker-Gordon model [12] instead of the Kolmogorov theory which is utilized in [29] because the correlation within the atmospheric turbulence can be directly modeled with Booker-Gordon model. In this model, the correlation function of the atmospheric scintillations is modeled with the exponential function, and it is given as [12]

$$B_n(l_x, l_y) \approx \exp(-l_x/l_x^c - l_y/l_y^c), \quad (4.3)$$

where $l_{x,y}$ are the horizontal and vertical distance to the reference point, and $l_{x,y}^c$ are the correlation distances of the turbulence for vertical and horizontal axis. For troposcatter, l_y^c is approximately ≈ 50 m and it varies 20 – 130 m according to the atmospheric conditions [12]. Since l_x^c is much greater than l_y^c as in [12], we assume that $l_x = 4 \times l_y$ [12].

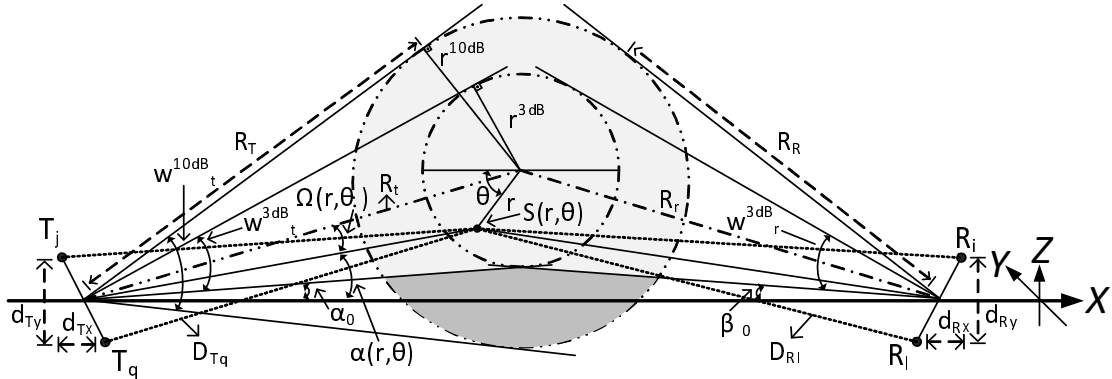


Figure 4.1: Geometry of RSM method.

The geometry dependency of the scattering cross-section can be determined with $\sin(\Psi)^{-11/3}$ as experimentally validated in [47]. With the proposed simplifications for the correlation analysis, the resulting power relationship in the troposcatter channel is found as

$$P_r/P_t \approx \sin(\Psi)^{-11/3} B_n(l_x, l_y) G_t(\hat{i}) G_r(\hat{o}), \quad (4.4)$$

where D_{T_j} is the distance between scatterer ($S(r, \theta)$) and transmitter j , and D_{R_i} is the distance between scatterer $S(r, \theta)$ and receiver i as in Figure 4.1. Since the antenna gains will change dramatically for 10 dB beam-widths, we include the antenna gains in the model as well.

The scattering model which is considered in this chapter is developed for the clear air atmospheric turbulence between 1 – 10 GHz because the atmospheric correlation model is valid for the clear air conditions and the scattering power relationship depends on the Rayleigh scattering which is more accurate for 1 – 10 GHz [12]. In addition, we assume that the transmitter and receiver have the same vertical orientation for simplicity. Therefore, we consider only the symmetrical troposcatter paths in this chapter. However, our model can be extended for non-symmetrical cases with geometrical approximations.

4.3.2 RSM Model

In troposcatter, only the scattered rays inside the troposcatter common volume can be received due to path geometry. The scattered rays outside of this region will reach the receiver with either lower or higher angles than the 10 dB beam-width of the receiver. Although the scattering particles are located through the troposphere, we only consider the scatterers that are located in a ring within the intersection of the transmitter and receiver 10 dB beam-widths as in Figure 4.1. The boundaries for the scatterers is given as θ in $[-\pi, \pi)$ and r in $[0, r^{10dB}]$. The radius of the rings are given by

$$\begin{aligned} r^{3dB} &= R_t \sin(w_t^{3dB}/2), \\ r^{10dB} &= R_t \sin(w_t^{10dB}/2), \end{aligned} \quad (4.5)$$

where R_t is the path between the center of the rings and transmitter. w_t^{3dB} and w_t^{10dB} are the transmitter 3 dB and 10 dB beam-widths, respectively.

To maximize the received power, the 3 dB beam-widths of the antennas are adjusted to the radio horizon as in Figure 4.1 [4, 29]. Therefore, the lower part of the

ring (the darker region in Figure 4.1) will be blocked by the path geometry due to the curvature of the earth. Since RSM method utilizes the 10 dB beam-widths, the lower part of the ring will be eliminated from the correlation calculations as described in Appendix 13-B.

Due to the troposcatter geometry, the path-loss observed by the both end of the communication is symmetrical [1]. The troposcatter links generally utilize the same type of antennas at the both ends. For these reasons, we assume that the troposcatter path is symmetrical. Thus, we have $w_t^{3dB} = w_r^{3dB}$ and $w_t^{10dB} = w_r^{10dB}$ that imply $R_T = R_R$ as [29]

$$R_{T,R} = R \sin(\beta_0) / \sin(\alpha_0 + \beta_0), \quad (4.6)$$

where α_0 and β_0 are the elevations of the antenna beams with respect to horizon for transmitter and receiver, respectively [29]. R is the horizontal distance between transmitter and receiver.

Since the b-LoS troposcatter communication does not have any line-of-sight component, only the scattered rays are considered in RSM and the power of the scattered rays are determined by (4.4). Suppose that there are K effective scatterers in each ring and L rings within the common volume. Thus, the normalized path gain $\mathbf{H}_{i,j}$ is represented as [40]

$$\begin{aligned} \mathbf{H}_{i,j} = & \int_0^{r^{10dB}} \int_0^{2\pi} \sum_{n=1}^L \sum_{k=1}^K \delta(\theta - \theta_k) \delta(r - r_n) \zeta(r, \theta) \\ & G_t(r, \theta) G_r(r, \theta) \sin(\Psi(r, \theta))^{-11/3} B_n(l_x, l_y) \\ & \exp(-j2\pi/\lambda(D_{T_j} + D_{R_i}) + j\phi(r, \theta)) d\theta dr, \end{aligned} \quad (4.7)$$

$\zeta(r, \theta)$ is defined to exclude the lower part of the ring as described in Appendix 13-D. $G_{t,r}(r, \theta)$ are the antenna gains, as defined in Appendix 13-C. In Section 4.3.1, the multi-scatter nature of the troposcatter is approximated by the single scattering [12]. Therefore, the phase ($\phi(r, \theta)$) is assumed as uniformly distributed $[-\pi, \pi)$ and iid. Thus, phase term is canceled in the correlation analysis as in [40].

For the fading correlation analysis, $\mathbf{H}_{M_r \times M_t}$ matrix is reshaped into $\hat{\mathbf{h}}_{M_r M_t \times 1}$ vector. In this case, the fading correlation between different antenna pairs becomes the

covariance between $\widehat{\mathbf{h}}_{i,j}$ and $\widehat{\mathbf{h}}_{l,q}$ that can be represented as $E[\widehat{\mathbf{h}}_{i,j}\widehat{\mathbf{h}}_{l,q}^*]$ where $(\cdot)^*$ is the conjugate operation [40].

4.3.3 Space Diversity

We assume that there are infinitely many scatterers inside the scattering common volume, thus L and K in (4.7) go to infinity. By using (4.7), the spatial fading correlation between different troposcatter antenna pairs is represented as

$$\begin{aligned} E[\widehat{\mathbf{h}}_{i,j}\widehat{\mathbf{h}}_{l,q}^*] &= \int_0^{r^{10dB}} \int_0^{2\pi} B_n(|r \sin(\theta)|, |r \cos(\theta)|)^2 \\ &\quad G_t(r, \theta) G_r(r, \theta) \zeta(r, \theta) \xi(r, \theta) \\ &\quad \exp(-j2\pi/\lambda(\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}})) d\theta dr, \end{aligned} \quad (4.8)$$

where $\Delta D_{T_{j,q}} = D_{T_j} - D_{T_q}$ is the path length difference between local scatterer to transmitters j and q , and $\Delta D_{R_{i,l}} = D_{R_i} - D_{R_l}$ is the path length difference between local scatterer to receivers i and l . $\xi(r, \theta)$ function represents the $\sin(\Psi_1)^{-11/3} \times \sin(\Psi_2)^{-11/3}$ term as a function of r and θ as defined in Appendix 13-D. B_n is the correlation function (4.3). Since the reference point is assumed as the center of the ring, the vertical and horizontal distance from the reference can be simply find by $|r \sin(\theta)|$ and $|r \cos(\theta)|$, respectively.

Since troposcatter paths use low beam-widths and elevation angles, (4.8) can be simplified by using geometrical approximations. For low beam-widths, $\Delta D_{T_{j,q}} \approx d_{T_x} \cos(\Omega(r, \theta)) + d_{T_y} \sin(\Omega(r, \theta))$, where $\sin(\Omega(r, \theta)) \approx r/R_t \sin(\theta)$, and $\cos(\Omega(r, \theta)) \approx (1 - \frac{1}{4}(r/R_t)^2 + \frac{1}{4}(r/R_t)^2 \cos(2\theta)) = \xi(r, \theta)$ where $d_{T_{x,y}}$ are the vertical and horizontal spacing of the transmitters. R_t is the distance between transmitter and the center of the troposcatter common volume as in Figure 4.1. Since the troposcatter path is assumed as symmetrical, $R_t = R_r$ and given as

$$R_{t,r} = R \frac{\sin(\beta_0 + w_t^{3dB}/2)}{\sin(\alpha_0 + \beta_0 + w_t^{3dB})}. \quad (4.9)$$

By using the geometrical approximations, $\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}}$ term in (4.8) is repre-

sented as

$$\begin{aligned}\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}} &= d_{T_y} r / R_t \sin(\theta) + d_{T_x} \xi(r, \theta) + \\ &+ d_{R_y} r / R_r \sin(\theta) + d_{R_x} \xi(r, \theta).\end{aligned}\quad (4.10)$$

We only consider horizontal space diversity because it is the most common type of spatial diversity for troposcatter. Therefore, x axis component of the receive and transmit antennas becomes $d_{T_x} = d_{R_x} = 0$. Since the troposcatter channel is symmetrical for transmit and receive sides, the correlation calculations that are developed for one side will be valid for the other side as well. For this reason, we only focus on the correlation between two receivers which are communicating with the same transmitter. This condition implies that $d_{T_y} = 0$. By using (4.8), (4.10) and geometrical approximations, the correlation for horizontally spaced receivers is given as

$$\begin{aligned}E[\widehat{\mathbf{h}}_{i,j} \widehat{\mathbf{h}}_{i,l}^*] &= \int_0^{r_{10dB}} \int_0^{2\pi} \xi(r, \theta) B_n(|r \sin(\theta)|, |r \cos(\theta)|)^2 \\ &\quad \zeta(r, \theta) G_t(r, \theta) G_r(r, \theta) \exp(-j2\pi/\lambda d_{R_y} r / R_t \sin(\theta)) d\theta dr.\end{aligned}\quad (4.11)$$

(4.11) can be solved with numerical integration on MATLAB. The calculated correlation results are normalized based on the zero spacing value which has the correlation of 1.

Figure 4.2(a) shows the correlation vs. antenna spacing to wavelength ratio (d/λ) by (4.11) for 250 km range and $l_y^c = 50$ m. As noticed, the increase in the beam-width decreases the correlation at the same ratio. The increase in the beam-width increases r_{10dB} by virtue of (4.5), and this condition also increases the troposcatter common volume. As the common volume increase, the separations between scatterers have higher distances and less correlation. Figure 4.2(b) shows the spatial correlation results for different vertical correlation distances for 1.5° 3 dB beam-widths. As noticed, there is a inverse relationship between the correlation distance and correlation.

According to [1], the antennas should be placed at least 100λ or higher distance apart to achieve low correlation values. Our results in Figure 4.2(a) also show that 100λ spacing is required for 1° and 1.5° beam-widths to achieve low correlation values.

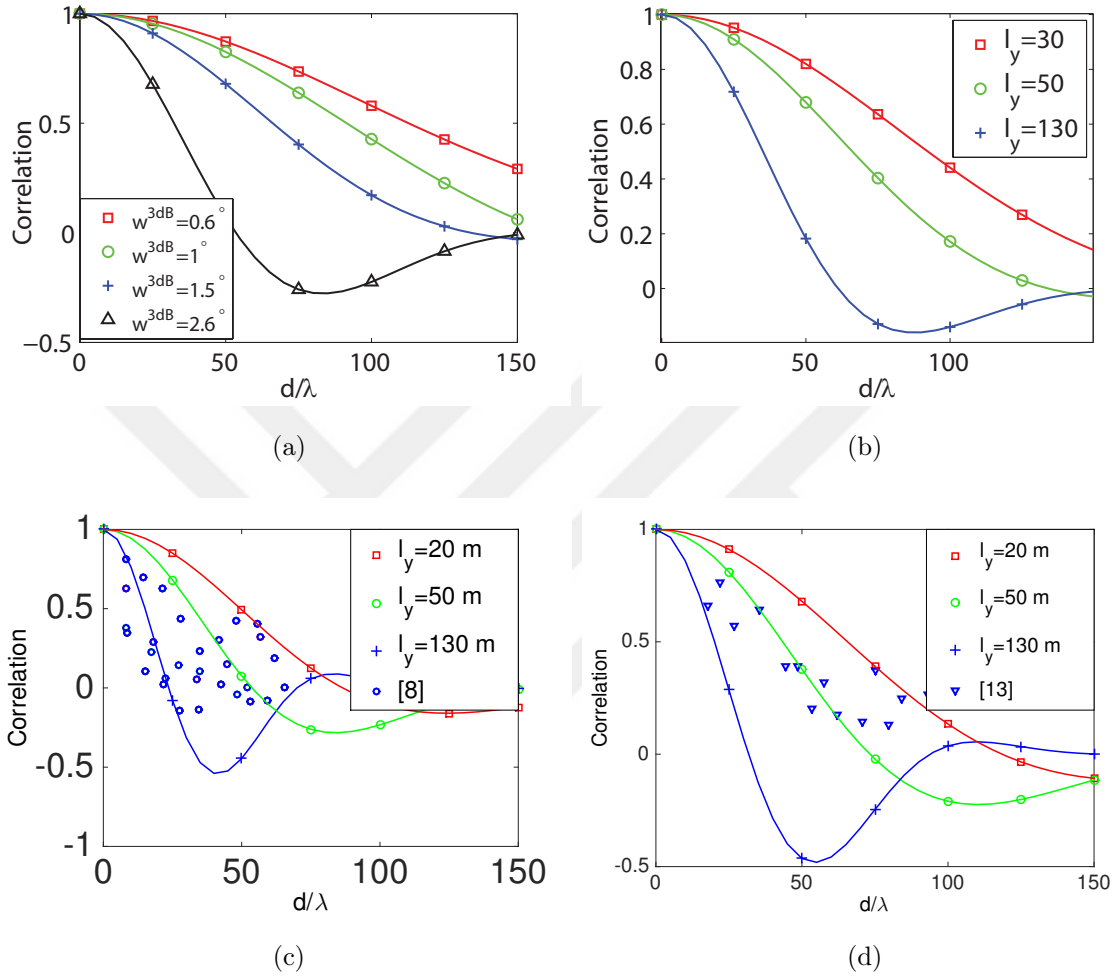


Figure 4.2: Fading correlation vs. spatial separation (a) for different beam-widths, and (b) for different correlation distance, (c) and (d) for the comparison with the experimental measurements.

Since 4.7 GHz troposcatter system with 3 m antenna diameter has $\approx 1.5^\circ$ beam-width ($w = \frac{70\lambda}{D_{ant}}$ where D_{ant} is the antenna diameter [24]), we can conclude that the our analytical results are consistent with the empirical results.

In addition, Figure 4.2(c) and 4.2(d) provides the comparison of the RSM results with the experimental results for 258 km and 2.6° 3 dB beam-width (Figure 3 in [8]), and for 321 km and 2.13° 3 dB beam-width (Figure 7 in [13]). As noticed,

our correlation results with the same channel parameters in Figure 4.2(c) and 4.2(d) are consistent with the experimental results presented in [8, 13]. However, some deviations between the theoretical and experimental results can be observed in these figures. Since the RSM considers the clear air propagation, the real world atmospheric conditions may cause changes in the correlation values. On the other hand, the general trend of the correlations can be well-estimated with RSM.

4.3.4 Frequency Diversity

For frequency diversity, the effect of the frequency separation is reflected in the system with the small change in wavelength (λ) in (4.7). In this way, the wavelength for antenna pairs will slightly differ. Therefore, the resulting covariance of (4.7) for antennas having frequency separation is represented as

$$E[\hat{\mathbf{h}}_{i,j} \hat{\mathbf{h}}_{t,q}^*] = \int_0^{r^{10dB}} \int_0^{2\pi} G_t(r, \theta) G_r(r, \theta) \xi(r, \theta) \zeta(r, \theta) B_n(|r \sin(\theta)|, |r \cos(\theta)|)^2 \exp \left\{ -j2\pi \left(\frac{D_{T_j} + D_{R_i}}{\lambda_j} - \frac{D_{T_q} + D_{R_l}}{\lambda_q} \right) \right\} d\theta dr, \quad (4.12)$$

where $\lambda_j = c/(f + \Delta f)$, $\lambda_q = c/f$, and Δf is the frequency separation. Therefore, the path calculations in (4.12) becomes

$$\begin{aligned} & (D_{T_j} + D_{R_i})(f + \Delta f)/c - (D_{T_q} + D_{R_l})f/c = \\ & (D_{T_j} + D_{R_i})\Delta f/c + (\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}})f/c. \end{aligned} \quad (4.13)$$

Firstly, we assume there is no spatial separation between receivers to investigate the effects of frequency diversity. Therefore, $(\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}})f/c$ term in (4.13) disappears. D_{T_j} and D_{R_i} are found by using their relationship with R_r . Since the distance between antennas and scattering common volume is much higher than the radius of common volume ($R_{t,r} \gg r^{10dB}$), we can use far distance approximation in which R_t and D_{T_j} are assumed to be parallel to each other. In the same way, R_r and D_{R_i} also become parallel as shown in Figure 4.3. By performing the required

geometrical calculations, the D_{T_j} and D_{R_i} is represented as

$$D_{T_j} = R_t - r \cos(\theta - \alpha), \quad (4.14)$$

$$D_{R_i} = R_r + r \cos(\theta + \beta). \quad (4.15)$$

The proof of these equations are left to the reader. Since the troposcatter path is assumed as symmetrical, $R_t = R_r$. Thus, $D_{T_j} + D_{R_i}$ term becomes

$$D_{T_j} + D_{R_i} = 2R_r - r \cos(\theta - \alpha) + r \cos(\theta + \beta). \quad (4.16)$$

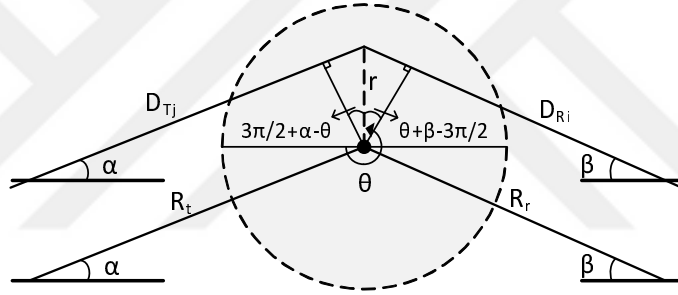


Figure 4.3: Channel geometry for the frequency diversity.

By using (4.12), (4.13) and (4.16), the covariance expression for frequency diversity is given as

$$\begin{aligned} E[\hat{\mathbf{h}}_{i,j} \hat{\mathbf{h}}_{l,q}^*] &= \int_0^{r^{10dB}} \int_0^{2\pi} \xi(r, \theta) \zeta(r, \theta) G_t(r, \theta) G_r(r, \theta) \\ &\quad B_n(|r \sin(\theta)|, |r \cos(\theta)|)^2 \exp \left\{ -j2\pi (\Delta f / c \right. \\ &\quad \left. (2R_r - r \cos(\theta - \alpha) + r \cos(\theta + \beta))) \right\} d\theta dr. \end{aligned} \quad (4.17)$$

Figure 4.4(a) presents the fading correlation vs. the frequency separation for different beam-widths in a 250 km and correlation distance 50 m troposcatter link. As in the space diversity, the increase in the beam-width decreases the correlation due to the increase in the common volume. According to the empirical recommendations provided by ITU [5], the frequency spacing between antennas (Δf) should be $\Delta f = 1.44f(D_{ant}^2 + 15^2)^{1/2}/\theta R$ Mhz, where f is the carrier frequency in Hz and R is the

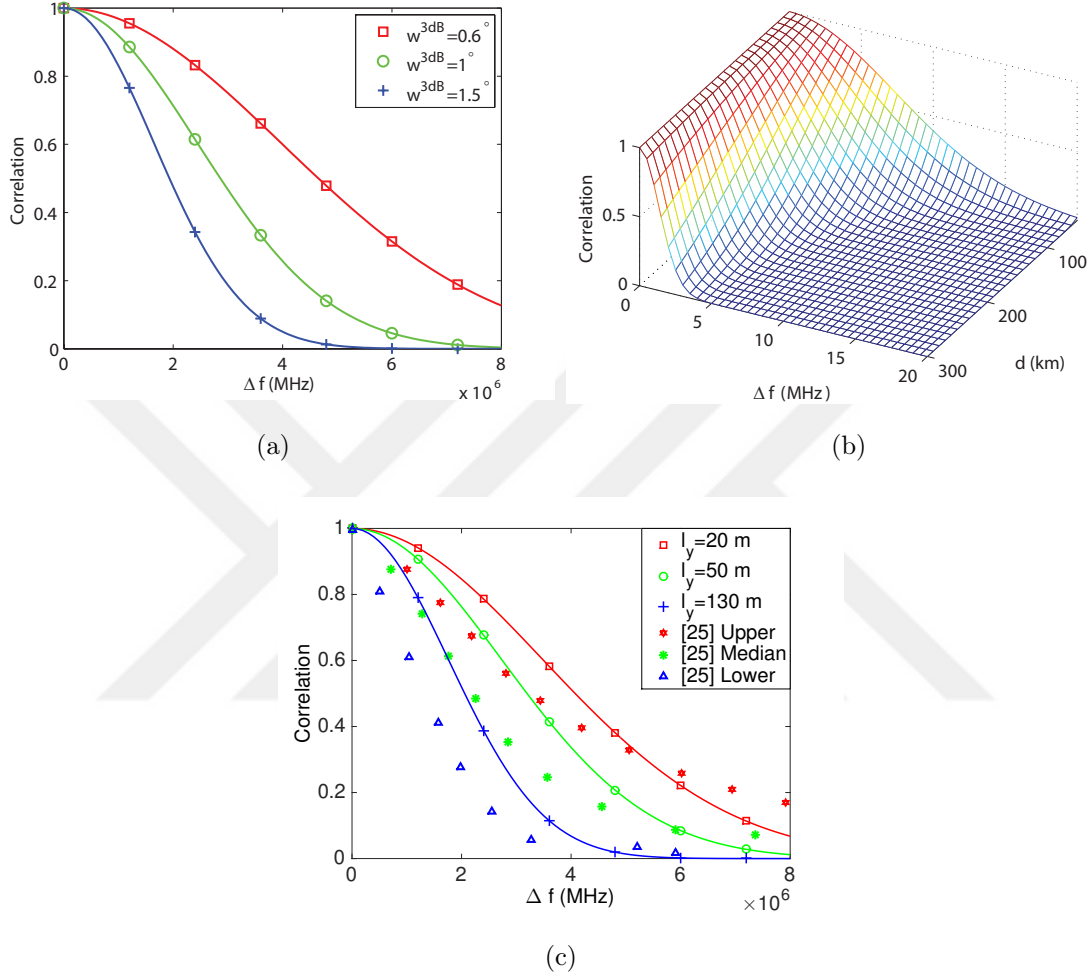


Figure 4.4: The fading correlation results for (a) frequency separation, (b) range and frequency separation.

range in m. For 4.7 GHz carrier frequency, 3 m antenna diameter and 250 km range, the recommended frequency spacing is 13 MHz. For the same antenna diameter (1.5° curve in Figure 4.4(a)), 10 MHz and greater frequency separations allow the antennas to be uncorrelated in the frequency diversity systems.

In addition, [25, 46] provides experimental frequency correlation results for the troposcatter links. To this end, Figure 4.4(c) provides the experimental frequency correlation results (Figure 9 in [25]) along with the RSM results by using the exper-

imental channel parameters: 192 km and 1.17° 3 dB beam-width at 7.6 GHz. As noticed, the trend of the experimental and RSM results are similar. However, the results of the RSM model is higher than the experimental results. This situation can be caused by the atmospheric conditions of the experiment site. However, the behavior of the frequency correlation can be well-captured with the RSM method.

The frequency separation formula (4.17) depends on the range by (4.9) as in the ITU's formula [5]. Figure 4.4(b) presents the fading correlation vs. range and frequency separation with the RSM for 1.5° 3 dB beam-width and 50 m vertical correlation distance. According to our results, the increasing range decreases the fading correlation at the same frequency separation value because the increasing range also cause increase in the troposcatter common volume. Also, the ITU's frequency separation formula is inversely proportional to range.

Most importantly, the analytical fading calculations for the frequency diversity can be utilized to derive the coherence bandwidth of the channel for 50% correlation. According to our results, 250 km troposcatter link at 4.7 GHz and 3 m antenna diameter have $B_c(50) \approx 2$ MHz. Therefore, the channel will have frequency-selective fading for bandwidths greater than 2 MHz. In such cases, the channel requires the implementation of orthogonal-frequency-division-multiplexing (OFDM) to mitigate the effects of frequency-selective fading. In this chapter, we focus on the high data rate employments of troposcatter communications. Thus, we use higher bandwidths than the coherence bandwidth and introduce OFDM capacity calculations for the troposcatter systems by using our previously developed ray-based method [29] in Section 4.5.

4.3.5 Space-frequency Diversity

Troposcatter systems can also use the combination of diversity techniques. Space-frequency diversity is a promising technique which can provide higher data rates with low cost. Space-frequency diversity systems utilize horizontally placed antennas with frequency diversity. Since the frequency diversity antennas can be mounted on

the same parabolic reflector antenna, higher gains can be achieved with the space-frequency diversity by using the same number of parabolic reflector antennas as in space diversity. Therefore, we derive analytical expressions for the correlation between antennas as a function of spatial and frequency separation.

By using (4.12) and (4.13), the fading correlation for space-frequency diversity can be found directly using the same methodology as in the frequency diversity part. For the 2 receivers that are communicating with the same transmitter, $\Delta D_{T,j,q} = 0$. Therefore, by combining (4.12), (4.13) and (4.16) for horizontally placed antennas ($d_{R_x} = 0$), the covariance expression for the space-frequency diversity is found as

$$E[\hat{\mathbf{h}}_{i,j} \hat{\mathbf{h}}_{l,q}^*] = \int_0^{r^{10dB}} \int_0^{2\pi} \zeta(r, \theta) B_n(|r \sin(\theta)|, |r \cos(\theta)|)^2 \exp \left\{ -j2\pi \left(\frac{\Delta f}{c} (2R_r - r \cos(\theta - \alpha) + r \cos(\pi - \theta - \beta)) \right) \right\} \xi(r, \theta) G_t(r, \theta) G_r(r, \theta) \exp(-j2\pi/\lambda d_{R_y} r/R_t \sin(\theta)) d\theta dr. \quad (4.18)$$

(4.18) is calculated with numerical integration on MATLAB.

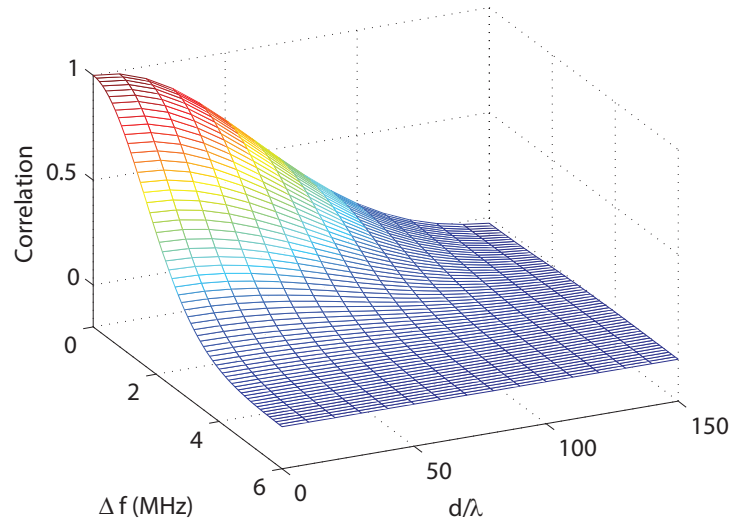


Figure 4.5: Fading correlation vs. spatial and frequency separations.

Figure 4.5 shows the fading correlation for space-frequency diversity for 250 km range, 50 m vertical correlation distance and 1.5° 3 dB beam-width. As noticed,

the required frequency separation for low correlation decreases with the increasing spatial separation. Similarly, the required spatial separation can be decreased with the frequency separation. For this reason, this employment is especially promising for systems that require placement of multiple antennas in close distances by using both parameters to achieve higher gains in MIMO.

4.3.6 Angle Diversity

Vertical angle diversity is also a promising method for the troposcatter communications because 2×2 vertical angle diversity system can form four different troposcatter common volumes. In addition, the angle diversity receivers can be mounted on the same parabolic reflector. Therefore, the cost of additional antennas will be low compared to space diversity. When the angle spacing of the antennas are higher than the beam-width of the antennas, the troposcatter common volumes will not intersect, and there will be low correlation between the troposcatter common volumes. However, in practical employments, troposcatter systems use very low angular separations because higher angular separation causes increase in the path lengths, and results in more path loss. In angle diversity, the source of the correlation between antennas is the intersection of two different troposcatter common volumes as in Figure 4.6.

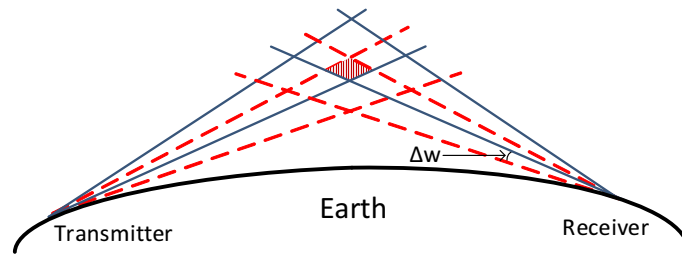


Figure 4.6: Troposcatter common volumes in the angle diversity system.

In RSM method, scatterers are assumed to be located just within the ring to make simple theoretical analysis. However, angle diversity system can be analyzed without this assumption with slight modification of the proposed model. For the

angle diversity, the shape of the common volume is modeled as parallelogram as in Figure 4.6. The only source of correlation is the intersection of the lower and upper common volumes. The rays that are associated with the intersection point have the same path lengths because the receivers are mounted to the same parabolic reflector, so that $(\Delta D_{T_{j,q}} + \Delta D_{R_{i,l}}) = 0$. Therefore, the correlation for the angle diversity is given as

$$\begin{aligned} E[\widehat{\mathbf{h}}_{i,j} \widehat{\mathbf{h}}_{l,q}^*] &= \int_{\alpha_0}^{\alpha_0 + w_t^{10dB}} \int_{\beta_0}^{\beta_0 + w_r^{10dB} + \Delta w} G_t^A(\alpha - \alpha_0 - w_t^{3dB}) G_r^A(\beta - \beta_0 - w_r^{3dB} - \Delta w) \\ &\quad B_n(|R_t \cos(\alpha) - R/2|, |h - h_C|)^2 d\alpha d\beta, \end{aligned} \quad (4.19)$$

where Δw is the angular spacing, $G_{t,r}^A$ are the Gaussian antenna gain functions which is given as $G_{t,r}^A(\Theta) = \exp(-\Theta^2 / (0.6 \times w^{3dB})^2)$ [6]. The middle of the scattering volume is assumed as the reference point for the correlation function B_n . The scattering height h and the center height of the troposcatter common volume h_C is given as [4]

$$h = R \frac{\sin(\beta) \sin(\alpha)}{\sin(\alpha + \beta)}, \quad (4.20)$$

$$h_C = R \frac{\sin(\beta_0 + w_t^{3dB}/2) \sin(\alpha_0 + w_t^{3dB}/2)}{\sin(\alpha_0 + \beta_0 + w_t^{3dB}/2 + w_r^{3dB}/2)}. \quad (4.21)$$

Figure 4.7 shows the angular fading correlation for different beam-widths for 250 km and 50 m vertical correlation values. As noticed, the fading correlation becomes very low when the angular separation is higher than the 3 dB beam-widths, as also suggested in [1]. Since the intersecting region will be higher for higher beam-widths, the correlation for the higher beam-widths are higher in the same separation values. The similar results can be found in [26] for lower 3 dB beam-widths.

Since increase in the angle causes increase in the path lengths and scattering angle, the upper beam will have considerably higher path-loss. This path-loss difference between lower and upper common volumes is called as squint-loss. The results for the squint-loss are presented in Section 4.6.1.

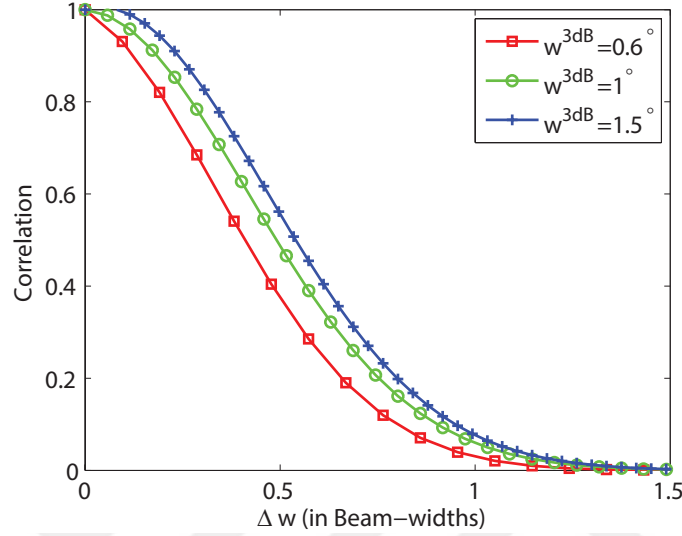


Figure 4.7: Fading correlation vs. angular separations.

4.4 Ray-based Troposcatter Channel Modeling Approach

In this work, we utilize our previously developed ray-based method [29] to compare different diversity techniques in terms of data rate. In addition, we present simulation results for the squint-loss in the angle diversity case. In [29], the beam-width of the antennas are divided into small parts, thus the delay and power calculations for each ray are performed separately. In this way, each ray is subjected to different atmospheric conditions according to its path.

The ray powers are calculated with differential bistatic radar equation

$$P_r = \frac{P_t G_t G_r \sigma_V \lambda^2 \varrho}{(4\pi)^3 R_t^2 R_r^2} W, \quad (4.22)$$

where P_t is the transmit power in Watt (W), $G_{t,r}$ are the antenna gains for transmitter/receiver. ϱ is the polarization mismatch factor. σ_V is differential scattering cross-section.

The atmospheric profiles are included via the scattering cross-section. We define the differential scattering cross-section by using Rayleigh scattering and Kolmogorov spectrum [12]. In this way, our model can take the real world water-vapor mixing

ratio measurements [35] as input to model the non-homogeneities and fluctuations in the air turbulence. The differential scattering cross section is represented as [29]

$$\sigma_V = 2\pi k^4 \cos(\Psi)^2 dV_c \Phi(k_s) \text{ m}^2, \quad (4.23)$$

where k is the wave-number, Ψ is the scatter angle, dV_c is the differential scattering volume. In addition, $\Phi(k_s)$ is the Kolmogorov spectrum which is formulated as [12]

$$\Phi(k_s) = 0.33\pi^3 C_n^2 (2k \sin(\Psi/2))^{-11/3} \text{ m}^3, \quad (4.24)$$

where $k_s = 2k \sin(\Psi/2)$. C_n is the structure constant of the refractive-index which is directly generated by using the water-vapor mixing ratio measurements [35] (For more details [29]).

4.5 Maximum Data Rate

In this section, we present the calculations for the achievable data rates that can be provided by the b-LoS troposcatter communication under the employment of different diversity techniques. To this end, we improve our ray-based channel modeling approach [29] to implement the diversity techniques in the troposcatter channel with MIMO-OFDM.

4.5.1 System Model for Maximum Data Rate

The block diagram of MIMO-OFDM troposcatter system is in Figure 4.8. In this work, only single-user to single-user link is considered to show the performance of the troposcatter channel, and the channel is assumed to have slow fading. Full channel state information (CSI) is assumed at both sides of the communication. Therefore, the maximum ratio transmission technique [51] is utilized. In this technique, the transmitter beam-forming is implemented with maximum-ratio-combining (MRC). Transmitter beam-forming with MRC reception is also called as MIMO-MRC.

In frequency domain, the $M_r \times 1$ received signal vector for k^{th} OFDM tone ($\mathbf{y}[k]$) is given as

$$\mathbf{y}[k] = \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t s + \mathbf{n}, \quad (4.25)$$

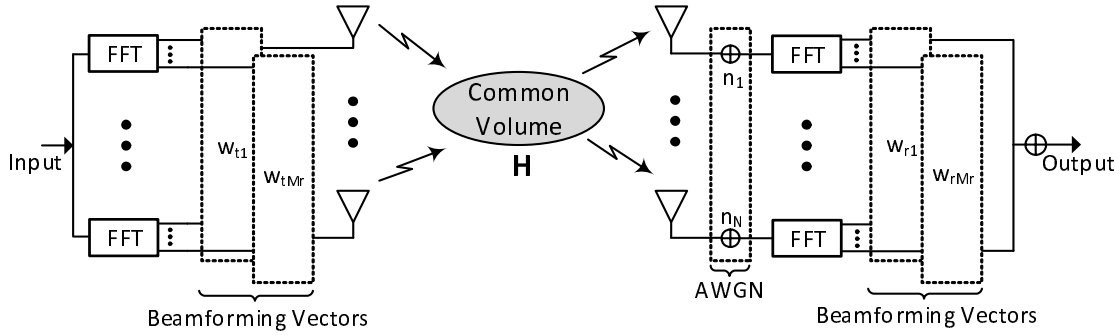


Figure 4.8: OFDM antenna array block diagram.

where s is the transmitted symbol with average power P_s , $\mathbf{w}_t$ is the transmitter beam-forming vector, $\mathbf{n}$ represents AWGN vector, and N is the number of OFDM tones.

The estimated output signal ($\tilde{\mathbf{y}}[k]$) is given by

$$\tilde{\mathbf{y}}[k] = \mathbf{w}_r^H \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t s + \mathbf{w}_r^H \mathbf{n}, \quad (4.26)$$

where $(\cdot)^H$ represents the conjugate transpose operation, $\mathbf{w}_r$ is the receiver beam-forming vector.

By using the following property $\|\mathbf{w}_r\|^2 = 1$, the SNR for k^{th} tone can be found with the following expression [52]

$$\begin{aligned} \gamma[k] &= \bar{\gamma} \frac{\mathbf{w}_t^H \mathbf{H} (e^{j2\pi k/N})^H \mathbf{w}_r \mathbf{w}_r^H \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t}{\mathbf{w}_r^H \mathbf{w}_r}, \\ &= \bar{\gamma} \mathbf{w}_t^H \mathbf{H} (e^{j2\pi k/N})^H \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t, \end{aligned} \quad (4.27)$$

where $\bar{\gamma} = P_s/N_0$ is the average SNR per receive antenna.

Since we have the $\|\mathbf{w}_t\|^2 = 1$ constraint for the transmit beam-forming vector, $\mathbf{w}_t^H \mathbf{H} (e^{j2\pi k/N})^H \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t$ is on the unit sphere. This expression is maximized by the $\lambda_{\max}$, the largest eigenvalue of $\mathbf{H} (e^{j2\pi k/N})^H \mathbf{H} (e^{j2\pi k/N})$ (For more detail, see [51–53]). Therefore, the maximum achievable data rate, *capacity*, is given by

$$C = \max_{P_k: \sum P_k = P_t} \sum_{k=0}^{N-1} B_N \log \left(1 + \frac{\lambda_{\max}^k P_k}{\sigma_n^2 B_N} \right), \quad (4.28)$$

where P_t is the total power constraint, B_N is the bandwidth of the each sub-channel (B/N), λ_{max}^k is the largest eigenvalue for $\mathbf{H}(e^{j2\pi k/N})^H \mathbf{H}(e^{j2\pi k/N})$. σ_n^2 is the noise power in W/Hz which is modeled as thermal noise. In addition, P_k is the allocated power for k^{th} OFDM sub-channel and they are calculated by the well-known water-filling algorithm [27]

$$\frac{P_k}{P_t} = \begin{cases} 1/\gamma_0 - 1/\gamma_k, & \gamma_k \geq \gamma_0 \\ 0, & \gamma_k < \gamma_0 \end{cases} \quad (4.29)$$

where γ_0 is the cutoff value and $\gamma_k = \lambda_{max}^k P_t / \sigma_n^2 B_N$. By using (4.28) and (4.29), the capacity is simplified into [27]

$$C = \sum_{k: \gamma_k \geq \gamma_0} B_N \log(\gamma_k / \gamma_0). \quad (4.30)$$

Up to now, the framework for the capacity calculations is formed. The remaining of the section includes how to generate $\mathbf{H}(e^{j2\pi k/N})$ matrix with our previously proposed ray-based method and RSM for fading correlation.

4.5.2 Channel Gain Matrix for Diversity Techniques

To generate $\mathbf{H}(e^{j2\pi k/N})$, we first use the channel matrix normalization to exclude the effects of large-scale path loss which is calculated by the ray-based technique [29]. The large-scale path loss is derived by using real world measurements as described in Section 4.4. The small-scale fading is assumed as Rayleigh fading as suggested in [5, 6, 8–10]. Normalized channel gain matrix $\tilde{\mathbf{H}}(e^{j2\pi k/N})$ is found as [41]

$$\tilde{\mathbf{H}}(e^{j2\pi k/N}) = \frac{\mathbf{H}(e^{j2\pi k/N})}{\sqrt{\frac{1}{M_t M_r} \|\mathbf{H}(e^{j2\pi k/N})\|_F^2}}, \quad (4.31)$$

where $\|\cdot\|_F$ is the Frobenius norm. Thus, the average SNR on the channel is found as [41, 42]

$$SNR_{av} = P_t / (\sigma_n^2 M_t M_r) \|\mathbf{H}(e^{j2\pi k/N})\|_F^2, \quad (4.32)$$

where SNR_{av} is the average receive SNR that is calculated with the ray-based method. As described in Section 4.4, the water-vapor mixing ratio measurements are utilized

for atmospheric modeling, and the data set includes 1000 measurements that are measured in August near Cape Verde Islands [35]. Therefore, the average SNR is calculated for each measurement instant to model the atmospheric fluctuations.

With the channel normalization, (4.27) can be represented as

$$\begin{aligned} \mathbf{w}_t^H \mathbf{H} (e^{j2\pi k/N})^H \mathbf{H} (e^{j2\pi k/N}) \mathbf{w}_t = \\ SNR_{av} \mathbf{w}_t^H \tilde{\mathbf{H}} (e^{j2\pi k/N})^H \tilde{\mathbf{H}} (e^{j2\pi k/N}) \mathbf{w}_t. \end{aligned} \quad (4.33)$$

This expression is maximized by the largest eigenvalue of $\tilde{\mathbf{H}} (e^{j2\pi k/N})^H \tilde{\mathbf{H}} (e^{j2\pi k/N})$, and $\tilde{\mathbf{H}}$ is generated for each diversity technique differently.

Space diversity

With the correlation results that are generated in Section 4.3.3, the normalized channel gain matrix is modeled as [43]

$$\tilde{\mathbf{H}} (e^{j2\pi k/N}) = (\mathbf{R}_H^{RX})^{1/2} \mathbf{G} (\mathbf{R}_H^{TX})^{1/2}, \quad (4.34)$$

where $\mathbf{G}$ is the complex Gaussian iid matrix. $(\mathbf{R}_H^{RX})^{1/2}$ is the root of the receiver antenna correlation matrix $(\mathbf{R}_H^{RX})^{1/2} [(\mathbf{R}_H^{RX})^{1/2}]^H = \mathbf{R}_H^{RX}$ which is found by Cholesky factorization. $(\mathbf{R}_H^{TX})$ is the transmitter correlation matrix.

Frequency diversity

Since the frequency separation between antennas nulls the cross components, there are $M_t = M_r$ parallel channels in frequency diversity. Therefore, $\mathbf{G}$ in (4.34) is generated as diagonal matrix for frequency diversity.

Space-frequency diversity

Since the frequency diversity is provided with only one parabolic reflector, we can use horizontally placed antennas with frequency diversity, and achieve quad-diversity. Suppose that N_{SFD} represents the number of parabolic reflectors in one side and there

are equal number of them in each side, $\mathbf{G}$ matrix in (4.34) for the space-frequency diversity becomes

$$\mathbf{G}_{M_F N_{SFD} \times M_F N_{SFD}} = \text{diag}(\mathbf{G}_1, \dots, \mathbf{G}_{N_{SFD}}). \quad (4.35)$$

where $\mathbf{G}_{1, \dots, N_{SFD}}$ represents the complex Gaussian iid channel matrix with $M_F \times M_F$, where M_F is the number of frequency used for the frequency diversity. The function diag forms a block diagonal matrix from its input arguments.

Angle diversity

In angle diversity, the increase in the elevation angles cause significant decrease in the average SNR. Instead of channel matrix normalization, the average SNR for each antenna pair is calculated separately for $M_t = M_r$ case. (4.27) is represented as (4.36) for $M_t = M_r$ angle diversity system where $\circ$ is Hadamard product that is element by element matrix multiplication. Therefore, (4.36) is maximized with the largest eigenvalue of $[SNR_{avi,j}]_{\substack{i=1, \dots, M_r \\ j=1, \dots, M_t}} \circ \tilde{\mathbf{H}}(e^{j2\pi k/N})^H \tilde{\mathbf{H}}(e^{j2\pi k/N})$. (4.36) is calculated with the ray-based technique for each antenna pair for the angle diversity. $\tilde{\mathbf{H}}(e^{j2\pi k/N})$ term in (4.36) is also modeled as in (4.34).

$$\begin{aligned} \mathbf{w}_t^H \mathbf{H}(e^{j2\pi k/N})^H \mathbf{H}(e^{j2\pi k/N}) \mathbf{w}_t &= \frac{1}{M_t} \mathbf{w}_t^H \begin{bmatrix} SNR_{av1,1} & \cdots & SNR_{av1,M_t} \\ \vdots & \ddots & \vdots \\ SNR_{avM_r,1} & \cdots & SNR_{avM_r,M_t} \end{bmatrix} \\ &\circ \left(\tilde{\mathbf{H}}(e^{j2\pi k/N})^H \tilde{\mathbf{H}}(e^{j2\pi k/N}) \right) \mathbf{w}_t. \end{aligned} \quad (4.36)$$

4.6 Performance Evaluation

In this section, we compare the troposcatter diversity techniques by using maximum data rate simulations. In addition, we present the squint-loss results for the angle diversity.

4.6.1 Squint-Loss Results

For space and frequency diversity cases, the antennas utilize the same vertical orientation. Therefore, different antenna pairs in space and frequency diversity systems can maintain the same average received power levels. On the other hand, the angle diversity utilize angular separation and upper beam is subjected to significantly higher path-loss because both the path lengths and the scattering angle increase. For this reason, we analyze the additional path-loss caused by the angular separation which is also known as squint-loss.

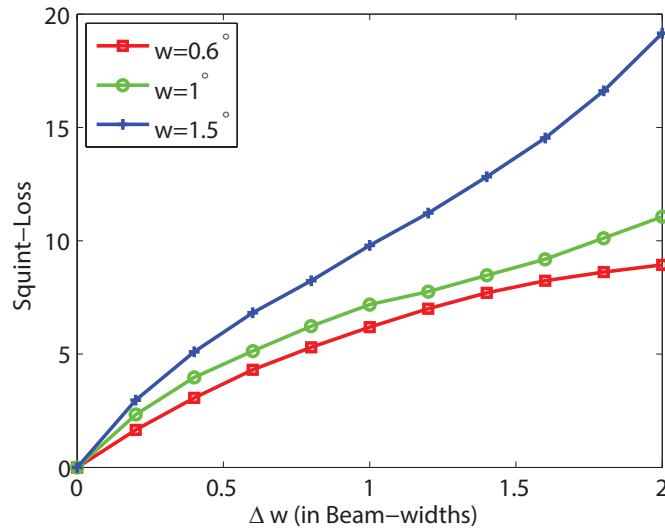


Figure 4.9: Squint-loss results for angle diversity.

We utilize the ray-based channel model [29] as in Section 4.4 for the following channel parameters: 250 km, 4.7 GHz carrier frequency, 41.5 dB antenna gains in each side. Figure 4.9 includes the squint-loss results for different beam-widths. As noticed, the squint loss increases with increasing beam-width because the separations are given in beam-widths. Similar results for the squint-loss is presented in [47] for the angle diversity. According to the experimental results, the upper and lower beams show 5 – 10 dB squint-loss for 0.6° 3 dB beam-width and 1° beam separation ($\Delta w \approx 1.6$ in beam-width). As noticed in Figure 4.9, the average squint-loss calculated with [29] is

also in the experimentally measured range.

4.6.2 Maximum Data Rate Results

We present the achievable data rate simulation results by using the RSM to estimate the required separations for the partial correlation between antennas. The required separations for partial correlations (for $\rho = 0.5$) can be found in Table 4.1.

Table 4.1: Troposcatter diversity separations.

Diversity technique	Separations	Correlation (ρ)
<i>Space diversity</i> (2×2)	$d = 65\lambda$ m	0.5
<i>Frequency diversity</i> (2×2)	$\Delta f = 2$ MHz	0.5
<i>Angle diversity</i> (2×2)	$\Delta w = 14$ mrad	0.5
Squint-loss= 6 dB		
<i>Space-frequency diversity</i> (4×4)	$d = 30\lambda$ m	0.5
	$\Delta f = 1.7$ MHz	

The troposcatter simulation parameters can be found in Table 4.2. In addition, there are 1000 atmospheric measurements in the used data sets [35]. We perform 1000 realization for the random $\mathbf{G}$ matrix as in Section 4.5.2. Therefore, the maximum data rate results are generated with the total of 1 million realization by using the ray-based model [29] in MATLAB.

Figure 4.10 presents the CDF of the data rate for different diversity techniques. As expected, 2×2 space diversity provides significantly higher data rates compared to frequency and angle diversity. In addition, the span of achievable data rates decreases with the space diversity, thus the additional antenna makes the channel more reliable. However, 2×2 frequency diversity provides the lowest data rate gains because the frequency separation between antennas nulls the cross antenna signal powers. On the other hand, 2×2 angle diversity, which can be performed with only 2 parabolic

Table 4.2: Troposcatter communication parameters.

Parameter	Value	Unit
<i>Horizontal distance (d)</i>	250	km
<i>Carrier frequency (f)</i>	4.7	GHz
<i>Transmit power (P_t)</i>	1000	W
<i>Bandwidth (B)</i>	10	MHz
<i>Average Transmission-loss (P_r/P_t)</i>	-138	dB
<i># of OFDM symbols (N)</i>	64	-
<i>Beam-widths ($w_{t,r}$)</i>	26	mrاد
<i>Beam-elevations (α_0, β_0)</i>	14.7	mrاد
<i>Antenna gains ($G_{t,r}$)</i>	41.5	dB _i
<i>Antenna diameters (D_{ant})</i>	3	m

reflectors, has significantly higher performance compared to the frequency diversity. Since the increase in the elevation angle decreases the received power of the additional troposcatter common volumes, the space diversity outperforms the angle diversity.

Although the frequency diversity has the worst achievable data rate performance, the space-frequency diversity can provide quad diversity by using 2 parabolic reflectors in each side. In this case, there are 2 horizontally placed antennas, and each antenna mounted with 2×2 frequency diversity receivers. As seen in Figure 4.10, the achievable data rate performance of the space-frequency diversity is higher than the space diversity, and its implementation is much more economical compared to adding additional parabolic reflector antennas.

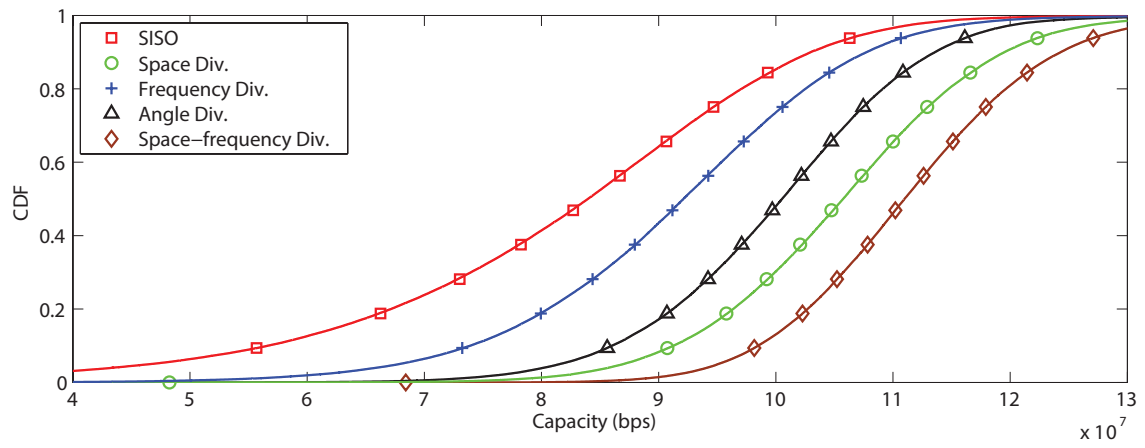


Figure 4.10: The distribution of the achievable data rates for different diversity techniques.

Chapter 5

RAIN ATTENUATION MODEL FOR TROPOSCATTER COMMUNICATION LINKS

5.1 Introduction

Troposcatter communications utilize scattering effect of lower troposphere to provide direct communication at BLOS ranges up to 300 km [24, 29, 54]. Troposcatter communication has high losses due to high path lengths and scattering. Therefore, path lengths should be minimized by making elevation angles of antennas as small as possible. In this way, troposcatter common volume, which is the intersection of transmit and receive antenna beams formed at the lower troposphere, and scattered rays are exposed to the effects of hydro-meteors. Since modern troposcatter equipments utilize frequency range between 4.7 – 15 GHz to provide more bandwidth, the most important hydro-meteor for troposcatter communications becomes rain because the effects of cloud and fog become considerable for carrier frequencies higher than 50 GHz [14].

There are many rain attenuation models in the literature [15–18]. Many of them consider rain attenuation in SATCOM [17, 18]. Since the geometry of SATCOM is different compared to troposcatter, rain attenuation models for SATCOM cannot be used for troposcatter communications. In addition, the most relevant study for our work is provided by International Telecommunication Union (ITU) [15] in which rain attenuation in troposcatter links is modeled with a rectangular rain cell having a uniform rainfall distribution and located in the middle of communication path. However, distribution of rainfall rate in horizontal plane shows significant changes in the real world conditions [55]. In addition, troposcatter paths cover 50 – 300 km

ranges, thus it is very high likely that the link is exposed to multiple rain cells.

The main contribution of this chapter is to develop a rain attenuation model for troposcatter communications which considers distribution of rainfall rates and multiple rain cells for the first time. Therefore, we develop rain path length expressions for nonuniform rainfall distribution and calculate rain loss due to each rain cell. We utilize available rain models [55–57] to generate synthetic rain distribution. We simulate the proposed rain attenuation model by modifying our previously introduced ray-based troposcatter channel modeling approach [29] to predict rain loss and maximum data rates in the presence of rain for the first time.

5.2 Ray-based Approach

In the ray-based approach, path length and power of each ray is calculated separately to form PDS [29]. In this way, each ray is subjected to different conditions according to its own path. In this approach, differential bistatic radar range equation is given as

$$P_r = \frac{P_t G_t G_r \sigma_V \lambda^2 \varrho a_{rain}}{(4\pi)^3 R_t^2 R_r^2} \text{ W}, \quad (5.1)$$

where P_t is the transmitter power, λ is the wavelength, $G_{t,r}$ are the TX/RX antenna gains, $R_{t,r}$ is the distance between the scatterer to TX/RX [4]. ϱ is the polarization mismatch factor. σ_V is the differential scattering cross-section which is modeled with the Rayleigh scattering approximation [12, 29]. a_{rain} is the rain attenuation which is introduced in this chapter. With the ray-based approach, rain attenuation associated with each ray is calculated separately according to distribution of rainfall rates. In this chapter, rain attenuation is mostly presented in dB as A_{rain} .

5.3 Rain Cell Model

We utilize HYCELL model [55, 56] to generate rain cells and their spatial distribution. However, we slightly modify HYCELL to integrate with our ray-based method and rain attenuation model. In [55], HYCELL utilizes both Gaussian and exponential

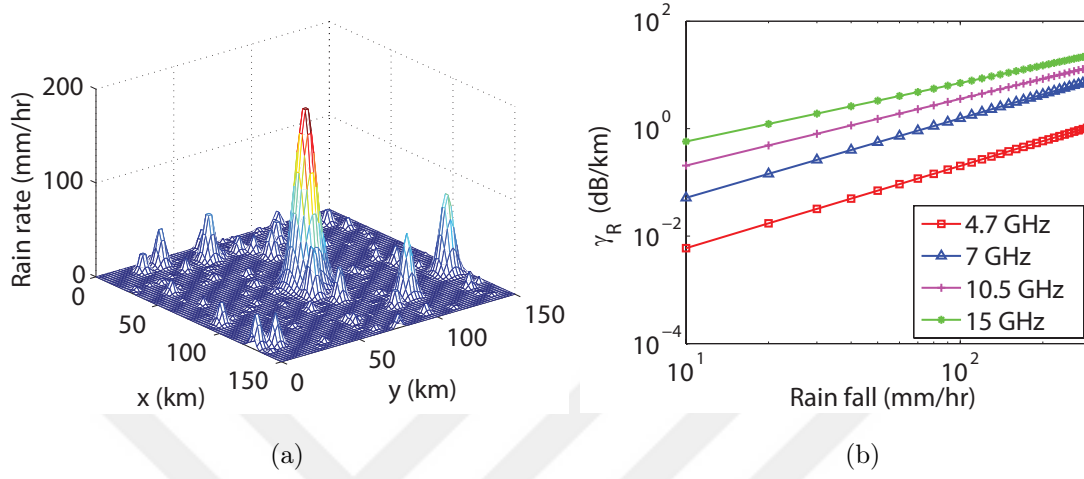


Figure 5.1: (a) Spatial distribution of rain cells and, (b) rain specific attenuation.

function to characterize rainfall distribution within a rain cell. Since rainfall rates that are associated with exponential function is too low, we model the rain cells only with Gaussian component. Therefore, rainfall rate distribution in a single cell is found as

$$R(x, y) = \begin{cases} R_G \exp \left[- \left(x^2/a_G^2 + y^2/b_G^2 \right) \right], & \text{if } R > R_1 \\ 0, & \text{otherwise,} \end{cases} \quad (5.2)$$

where R_G represents the peak rainfall rate. a_G, b_G are the distance where the rainfall rate decreases by a factor of $1/e$ in x and y axis. We assume symmetrical rain cells $a_G = b_G$. R_1 is the minimum rain rate, and it is assumed as 1 mm/hr as in [55]. Spatial distributions of rain cells are determined based on local spatial probability of rain, and we utilize ITU-R P. 837 – 6 [58] for local rain statistics. With HYCELL method [55, 56], we derive peak rain rates and diameters of rain cells for Istanbul, Turkey [58] as in Figure 5.1(a) for $150 \times 150 \text{ km}^2$ region.

In HYCELL model, spatial distribution of rain cells is studied, but rain height is not discussed. In this chapter, maximum rain height is critical to calculate rain attenuation in troposcatter links. As in Figure 5.2, rectangular rain cells are modeled in propagation path and only portion of rays which passes through a rain cell is

affected by rain. Therefore, height of a rain cell (h_R) is determined by [16] $h_R = h_0 + 0.36$ km, where h_0 is the mean 0°C isotherm height above mean sea level and average values for this parameter is used as provided in [16].

5.4 Rain Attenuation Model

Scattering and absorbing effects of rain droplets degrade performance of the troposcatter communications. Therefore, we propose a rain attenuation model for troposcatter communication. Unlike scattering particles ($1 - 10 \mu\text{m}$), sizes of rain droplets can vary $1 \mu\text{m}$ up to 1 mm . Since size of rain droplets and wavelength of signals become comparable especially for high carrier frequencies between $10 - 15 \text{ GHz}$, the effect of rain droplets cannot be modeled by Rayleigh scattering approximation as in [29]. To this end, in the presence of rain, path loss calculations require a correction factor (S) for the Rayleigh scattering approximation when carrier frequency is higher than 10 GHz and, this correction factor in dB is given as [15]

$$S = \begin{cases} R_G^{0.4} 10^{-3} [4(f - 10)^{1.6} ((1 + \cos(\Psi))/2) + \\ + 5(f - 10)^{1.7} ((1 + \cos(\Psi))/2)], & f > 10 \text{ GHz} \\ 0, & f \leq 10 \text{ GHz} \end{cases} \quad (5.3)$$

where R_G is the peak rainfall rate (mm/hr), f is the frequency (GHz), and Ψ is the scatter angle.

5.4.1 Rain Specific Attenuation

Rain specific attenuation (γ_R), that is amount of path loss in dB per km due to rain, is [59]

$$\gamma_R = k R^\gamma \text{ dB/km}, \quad (5.4)$$

where k and γ are empirical coefficients that show dependence on carrier frequency, path elevation angle and polarization. Empirical values for these coefficients can be found in [59]. Figure 5.1(b) shows rain specific attenuation as a function of rainfall rate for different carrier frequencies by using (5.4) and the parameters given by [59].

As noticed, high frequencies show more dependence on rain attenuation because effects of rain droplets increases as sizes of rain droplets and wavelength become more comparable.

5.4.2 Rain Path Calculations

Rain attenuation depends on rain cell geometry as in Figure 5.2 and 5.3. Therefore, rain path calculations should be performed separately for each path to include the effects of rain cell geometry. For troposcatter channel, rain path calculations are proposed in [15], but this model assumes a single rain cell at the middle of a troposcatter link with uniform rainfall rate distribution. As a result, this model is unrealistic for the real world conditions. Therefore, we develop rain path calculations for rain cells which can be anywhere between transmitter and receiver.

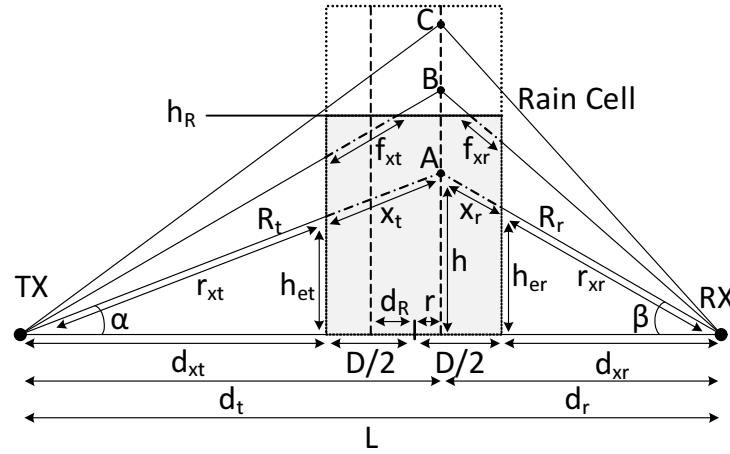


Figure 5.2: Geometry of the rain cell for Category 1.

As in Figure 5.2 and 5.3, troposcatter rays can be affected from rain conditions in different ways according to place of rain cell and scattering point. As a result, we divide rays into two main category according to location of rain cells. In the first category, distance between center of the rain cell and center of the communication path (d_R) is smaller than rain width $D/2$ as in Figure 5.2. In the second category

($d_R > D/2$), only one side of path is affected by rain cell as shown in Figure 5.3. To include multiple rain cells, rain attenuation calculations should be performed for each rain cell separately. At the end, total rain attenuation is given as

$$A_{rain}^{Total} = \sum_{i=1}^{N_R} A_{rain}^i \text{ dB}, \quad (5.5)$$

where N_R is the number of rain cell, A_{rain}^i is the rain attenuation associated with rain cell i .

Category 1 ($d_R < D/2$)

As in Figure 5.2, path calculations for this category follow as:

Lower Paths A: Lower paths (path **A** in Figure 5.2) have scattering point within the rain cell. Rain cell path lengths for transmitter side (x_t) and receiver side (x_r) are found as

$$\begin{aligned} d_{t,r} &= R_{t,r} \cos(\alpha, \beta), r = d_t - L/2 - d_R, \\ d_{x_{t,r}} &= d_{t,r} - D/2 \mp r, r_{x_{t,r}} = d_{x_{t,r}} / \cos(\alpha, \beta), \\ x_{t,r} &= \begin{cases} R_{t,r} - r_{x_{t,r}} & d_{t,r} > D/2 \\ R_{t,r} & d_{t,r} < D/2. \end{cases} \end{aligned} \quad (5.6)$$

Rain attenuation for lower paths at transmitter (A_{rainT}) is found by taking the integral of (5.4) for nonuniform rainfall distribution in (5.2) over the rain path calculated in (5.6) as

$$\begin{aligned} A_{rainT} &= \int_{-\frac{D}{2}}^{\frac{x_t}{\cos(\alpha)} - \frac{D}{2}} kR(x)^\gamma \frac{1}{\cos(\alpha)} dx \text{ dB}, \\ &= \int_{-\frac{D}{2}}^{\frac{x_t}{\cos(\alpha)} - \frac{D}{2}} \frac{R_G^\gamma}{\cos(\alpha)} \exp\left(-\frac{x^2}{a_G^2}\right)^\gamma dx \text{ dB}. \end{aligned} \quad (5.7)$$

With the same method, rain attenuation for receiver side is found as

$$A_{rainR} = \int_{\frac{D}{2} - \frac{x_r}{\cos(\beta)}}^{\frac{D}{2}} \frac{R_G^\gamma}{\cos(\beta)} \exp\left(-x^2/a_G^2\right)^\gamma dx \text{ dB}, \quad (5.8)$$

Total rain attenuation in each ray is the summation of (5.3), (5.7) and (5.8) as

$$A_{rain} = S + A_{rainT} + A_{rainR} \text{ dB}.$$

Middle Paths B: Scattering point of middle paths (associated with **B** in Figure 5.2) are outside of the rain cell and also some portion of the rays is still inside. Therefore, path lengths for middle paths is given as

$$f_{x_{t,r}} = \begin{cases} (h_R - h_{e_{t,r}}) / \sin(\alpha, \beta), & h_{e_{t,r}} < h_R \\ 0. & \text{otherwise.} \end{cases} \quad (5.9)$$

where $h_{e_{t,r}} = d_{x_{t,r}} \tan(\alpha, \beta)$

In addition to rain attenuation, scattering cross-section decreases as a function of height above rain height with -6.5 dB/km [15]. Therefore, total rain attenuation for middle paths is given as

$$\begin{aligned} A_{rain} = & 6.5(h - h_r) + \int_{-\frac{D}{2}}^{\frac{f_{x_t}}{\cos \alpha} - \frac{D}{2}} \frac{R_G^\gamma}{\cos(\alpha)} \exp\left(-\frac{x^2}{a_G^2}\right)^\gamma dx \\ & + \int_{\frac{D}{2} - \frac{f_{x_r}}{\cos \beta}}^{\frac{D}{2}} \frac{R_G^\gamma}{\cos(\beta)} \exp\left(-\frac{x^2}{a_G^2}\right)^\gamma dx + S \text{ dB.} \end{aligned} \quad (5.10)$$

Upper Paths C: Upper paths, that are associated with **C** in Figure 5.2, have no parts in rain cell. However, since scattering cross-section decreases above rain height as in middle paths, rain loss in upper paths is determined as

$$A_{rain} = S + \exp(-k_r(6.5(h - h_r))) \text{ dB.} \quad (5.11)$$

Category 2 ($d_R > D/2$)

Category 2 rays are associated with rain cells on only one side as in Figure 5.3. Path calculations for this category are given as:

Lower Paths D: In category 2, lower paths are associated with **D** ($h_{e_{t,r}} < h_R$) in Figure 5.3. Slant paths inside rain cells are found as

$$\begin{aligned} x_t &= \begin{cases} D / \cos(\alpha), & h_{e_r} > h_{e_t} \\ 0, & \text{otherwise,} \end{cases} \\ x_r &= \begin{cases} D / \cos(\beta), & h_{e_r} < h_{e_t} \\ 0, & \text{otherwise.} \end{cases} \end{aligned} \quad (5.12)$$

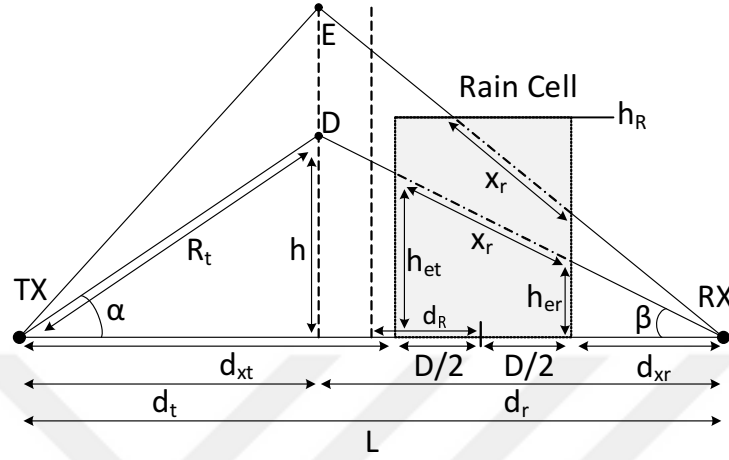


Figure 5.3: Geometry of the rain cell for Category 2.

Middle Paths E: In category 2, middle paths are associated with **E** ($h_{et} > h_R$ or $h_{er} > h_R$) in Figure 5.3. Rain cell path lengths for path **E** can be found as

$$\begin{aligned}
 x_t &= \begin{cases} (h_R - h_{et}) / \sin(\alpha), & h_{er} > h_{et} \\ 0, & \text{otherwise,} \end{cases} \\
 x_r &= \begin{cases} (h_R - h_{er}) / \sin(\beta), & h_{er} < h_{et} \\ 0, & \text{otherwise.} \end{cases}
 \end{aligned} \tag{5.13}$$

For paths **D** and **E**, resulting rain attenuation expressions are given as

$$\begin{aligned}
 A_{rain} &= 6.5(h - h_r) + \int_{-\frac{D}{2}}^{\frac{x_t}{\cos \alpha} - \frac{D}{2}} \frac{R_G^\gamma}{\cos(\alpha)} \exp\left(-\frac{x^2}{a_G^2}\right)^\gamma dx \\
 &+ \int_{\frac{D}{2} - \frac{x_r}{\cos \beta}}^{\frac{D}{2}} \frac{R_G^\gamma}{\cos(\beta)} \exp\left(-\frac{x^2}{a_G^2}\right)^\gamma dx + S \text{ dB.}
 \end{aligned} \tag{5.14}$$

Upper paths for category 2 is the same as category 1. Thus, (5.11) is also valid for category 2.

5.5 Maximum Data Rate

Maximum data rate in single-input-single-output troposcatter links can be calculated as

$$C(t) = B \log_2(1 + P_r(t)/N) \text{ bps,} \tag{5.15}$$

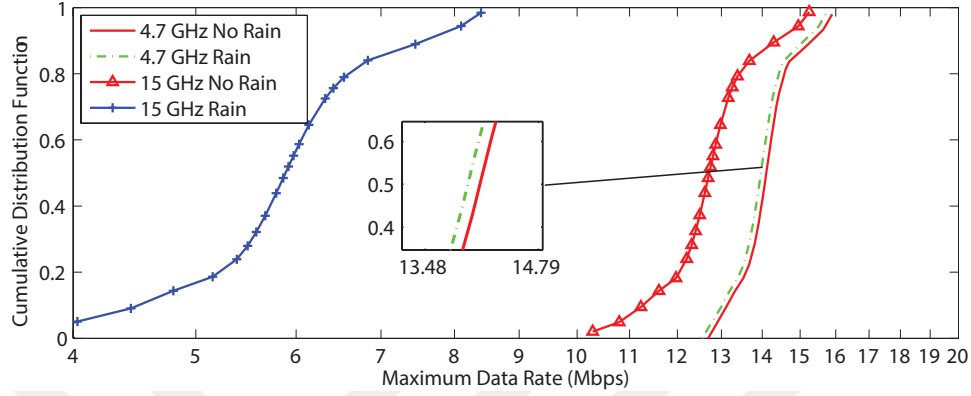
where P_r is the received power. N is the noise power (modeled as thermal noise $N = kBT$), T is the temperature in K, B is the bandwidth and k is the Boltzmann constant.

Our ray-based method [29] utilizes the real-world water vapor mixing ratio measurements [35]. Therefore, this approach can model fluctuations in the channel. In this work, we also utilize the same data sets [35] to model the atmospheric variations in the presence of rain to derive data rates and rain loss results. This approach [29] is also utilized in [54] to derive data rates for diversity techniques. In this way, PDS of the channel is calculated for each measurement instant with (5.1) and the path length calculations in [4, 29]. Average power of PDS for each time instant is denoted as $P_r(t)$ in (5.15). The effect of rain attenuation is integrated to the ray-based approach with a_{rain} in (5.15) to calculate data rates and rain loss in the presence of rain.

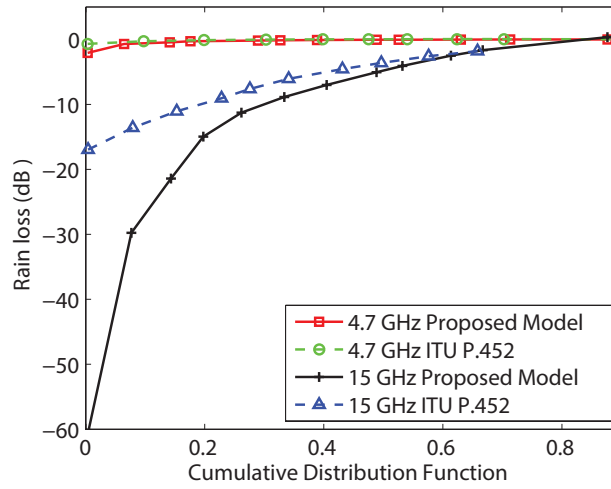
5.6 Simulation Results

In the ray-based approach, rain attenuation for each ray is calculated with the troposcatter path length calculations for the rainfall distribution as in Figure 5.1(a). The rain simulations are performed for each cross-section of Figure 5.1(a) for each 500 m to find the CDF of rain loss for the modeled region on MATLAB. The following channel parameters are utilized: range $L = 150$ km, beam elevation of antennas $\alpha_0 = \beta_0 = 8.3$ mrad, bandwidth $B = 1$ MHz. For frequency dependent terms, we utilize commercial troposcatter equipment data-sheets [3]: $G_t = G_r = 38.3$ dB, antenna beam-widths 2.08° , transmit power $P_t = 1000$ W for 4.7 GHz and $G_t = G_r = 49.7$ dB, antenna beam-widths 0.54° , transmit power $P_t = 250$ W for 15 GHz.

Figure 5.4(a) shows the distributions of the maximum data rates for different frequencies with and without rain by using the real world data sets as discussed in Section 5.5. The amount of rain loss increases with carrier frequency because the wavelengths and the sizes of the rain droplets becomes more comparable with the increasing carrier frequency. As in Figure 5.4(a), only extreme rain conditions can affect the performance of the 4.7 GHz curves. However, 15 GHz show very high dependence



(a)



(b)

Figure 5.4: (a) Distribution of data rates under rain for different carrier frequencies, (b) CDF of rain loss.

on rain. The same result can be seen in Figure 5.4(b) which presents distribution of rain loss with the proposed rain model and the single cell model in ITU P.452 [15]. As expected, the rain loss estimates with the single cell model are considerably lower especially at high frequencies because the single cell model cannot model the effects of multiple rain cells. In addition, 4.7 GHz shows slight dependence on rain attenuation, but there may be 3 dB power loss in the link. Therefore, we can conclude that lower

carrier frequencies are more promising for troposcatter communication links thanks to their resistance to the rain conditions.



Chapter 6

B-LOS COMMUNICATIONS WITH DUCTING LAYER

6.1 Introduction

Ducting layer is a promising communication medium for b-LoS applications because atmospheric ducts can provide a direct communication channel between navy-navy and navy-low flying units. To this end, this chapter includes the overview of the characteristics of the ducting layer b-LoS communications and the possible utilization of the ducting layer in the b-LoS naval communication networks. To this end, we review the channel modeling techniques and the software tools that can be used for the ducting layer. We present a preliminary analysis to show that ducting layer can achieve lower path-loss up to 20 dB at 500 km compared to the free space. In addition, we make a linear regression fitting for the path loss curve to find the path-loss exponent of the channel. To the best of our knowledge, there is no study which considers the path-loss exponent for the ducting layer. Therefore, this study provides both the characteristics of the ducting layer and the possible application areas by outlining the open research issues in the field.

The remaining of the chapter is organized as follows. Section 6.2 provides background information about the refractivity and different types of atmospheric ducts. In addition, the effects of ducts on wireless communication are discussed at the end of the section. In Section 6.3, the channel modeling approaches and the software tools are reviewed, and a preliminary analysis which investigates the path-loss exponent of the ducting layer is presented. Furthermore in Section 6.3, the open research issues on the ducting layer channel modeling are also pointed out. Lastly, the possible application areas of ducting layer b-LoS communications can be found in Section 6.4.

6.2 Properties of Atmospheric Ducts

The formation and the effects of atmospheric ducts entirely depend on the refractivity of the lower atmosphere. This section provides background for the refractivity, formation of ducts and effects of ducts on wireless communications.

6.2.1 Refractive Index, Refractivity and Modified Refractivity

Tropospheric radio refractive index (n) relies on weather conditions: wind, temperature, pressure and humidity at the lower atmosphere. Refractive index changes slightly with height. It varies between 1.000250 and 1.000400. Therefore, instead of the tropospheric radio refractive index, refractivity, which is the scaled version of the tropospheric radio refractive index, will be used and the refractivity (N) is defined as [62]

$$N = (n - 1) \times 10^6 \text{ N-units}, \quad (6.1)$$

where n is the atmospheric refractive index.

(6.1) is an idealized refractivity expression in which the earth surface is assumed to be flat, but to be more realistic the earth's curvature and the variation of height are considered in the modified refractivity (M) [62]

$$\begin{aligned} M &= N + \left(\frac{h}{a}\right) \times 10^6, \\ &= N + 157 \times h \text{ M-units}, \end{aligned} \quad (6.2)$$

where h is the height above sea level in km and a is the radius of the earth in km.

Tropospheric wave propagation depends on the rate of change in the modified refractivity that is associated with the refractivity gradient ($\partial M / \partial h$). There are four different refractive conditions according to the refractivity gradient as in Fig. 6.1. In sub-refraction condition, the signal is bent away from the surface. Standard refractivity condition is associated with the wave propagation in the standard atmosphere. Super-refraction condition causes the signal to propagate downward at a rate which is larger than the standard condition and smaller than the curvature of the earth.

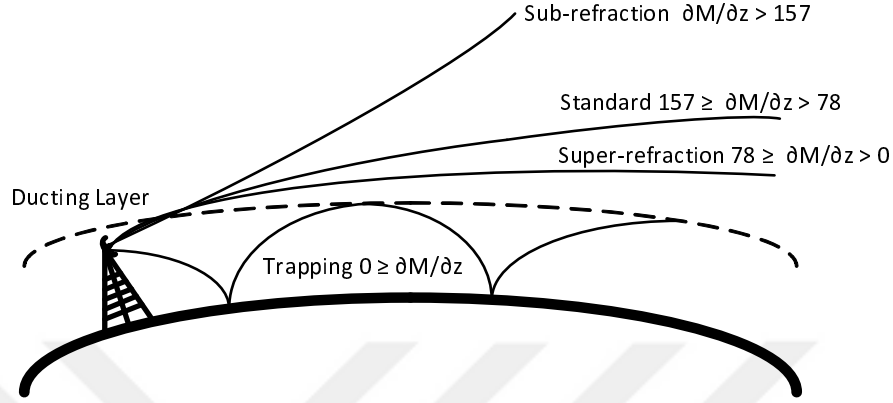


Figure 6.1: Rays under various refractive conditions.

The last and most important one is the trapping condition in which the signal is refracted back to the earth's surface. In this way, the refracted rays are trapped between surface or sea and ducting layer. The trapping condition is associated with a negative modified refractivity gradient and duct height is given by the height where the gradient of the modified refractivity changes from negative to positive, see Fig. 6.2.

The refractivity gradient and the duct height are the most important parameters for modeling atmospheric ducts. The characteristics of atmospheric ducts can be determined via the atmospheric parameters by using radiosonde or bulk atmospheric measurements. In addition, indirect methods are also available such as refractivity from clutter techniques which estimates the refractivity of the lower atmosphere from the clutter signal returns [20].

6.2.2 Formation and Characteristics of Ducts

The duct formation is an atmospheric phenomenon. Humidity, air-sea temperature, pressure and wind play significant roles in the duct formation. Among all the atmospheric conditions, humidity is the key and essential factor in the process. As a result, duct formation is more probable in regions with high evaporation rates such as equatorial and tropical regions.

The rapid humidity decrease in the few meters of the lower troposphere causes the duct formation. In addition, air-sea temperature is the most important factor which contributes to the process. When the air is warmer than the earth or sea surface, the temperature inversion occurs and it enhances the duct formation. Temperature inversion is caused by advection of dry and warm air mass or cooling of land when the air is still warm at nights. In addition, negative air-sea temperature difference can prevent the duct formation. This weather condition is named as unstable condition. Factors such as rough terrain, high winds, cold, stormy, rainy, and cloudy conditions also have destructive effects on the duct formation [19].

According to their physical formation process, there are three main types of atmospheric ducts:

Evaporation duct

Humidity difference in the air-sea boundary causes the formation of evaporation ducts. High evaporation rates are essential for evaporation ducts. Evaporation duct is the most common type of duct formation and it presents up to 90% of the time in equatorial and tropical regions. The M profile of the evaporation duct is modeled with the logarithmic curve in Fig. 6.2(a) [63]. Evaporation duct height is mostly between 10-20 m and it rarely reaches up to 40 m.

Evaporation ducts are the most promising duct formation which have potential to be utilized in the wireless communication applications thanks to their high availability rates. Since evaporation ducts are nearly permanent at coastal and maritime environments, b-LoS communications with the evaporation ducts can be employed especially for naval communications.

Surface-based duct

Surface-based ducts are mostly caused by advection of warm and dry air over the ocean or land which creates humidity and temperature inversion. Surface-based ducts can be modeled with trilinear curve in which the negative slopped part represents the

trapping layer as in Fig. 6.2(b) [20] and the duct height can reach hundreds of meters. Surface-based ducts occur up to 40% of the time depending on the season and location and these statistics can be found in [64].

Elevated duct

The presence of marine layers results in the formation of elevated ducts. The physical mechanism that creates elevated ducts is similar with surface-based ducts. However, in the elevated ducts the lower boundary of the trapping layer is above the surface. Elevated ducts occur at relatively higher heights between 600 m up to 3000 m and can also be modeled with trilinear curve in Fig. 6.2(c) [20]. Especially, air-to-ground links are prone to the trapping effect of the elevated ducts because trapped signals follow longer paths than expected and can reach to the receiver with higher delay spreads and different angles than expected. Therefore, elevated ducts have potential to create unexpected interference in the air-to-ground links. The occurrence rate of the elevated ducts can be up to 50% according to season and location, and these statistics can be also found in [64].

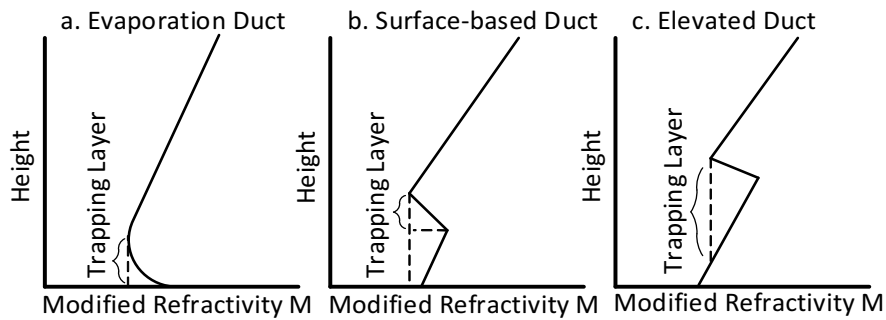


Figure 6.2: Modified refractivity models for duct types.

6.2.3 Effects of Ducts on Wave Propagation

Besides the path loss enhancements, atmospheric ducts also introduce some challenges in the b-LoS wireless communications. Main challenges introduced by the atmospheric

ducts and the techniques to overcome these challenges can be outlined as follows:

Inter-Symbol-Interference (ISI)

Atmospheric ducts trap the propagating signals at the lower atmosphere and causes signals to follow different paths. The trapped signals reach the receiver with different delay spreads and phase components. Therefore, these trapped signals can create destructive/constructive interference at the receiver. The receiver should optimally combine these components by using adaptive antenna techniques to prevent interference. To this end, optimal transmit and receive beam patterns should be designed according to the characteristics of the atmospheric ducts and these techniques provides both diversity and array gain along with interference suppression as recommended in [61].

Unexpected interference

Up to now, we consider both transmitter and receiver are located within the duct. On the other hand, the effects of ducts are prominent even if one of the antennas stay within the duct such as satellite communications and air-to-ground links in maritime and coastal environments. When satellite signals have low elevation angles, these signals can be trapped by atmospheric ducts and the trapped signals may be received by the receiver with different delay spreads and AoA. As a result, the receiver may misinterpret AoA of the incoming signal which is refracted by the ducting layer, and the surveillance system can be adversely affected by this effect.

The trapped signals in ducts follow longer paths and reaches to the receiver with higher delays as experimentally measured in [65]. Therefore, even in LOS ranges, this effect can create multipath fading and change the pattern of the interference in naval systems. As a result, (OFDM) can be promising for the ducting layer if the channel creates high delay spreads that makes the coherence bandwidth much lower than the bandwidth of the signal, orthogonal-frequency-division-multiplexing

Sea surface roughness

The trapped signals in the ducting layer are in contact with the sea surface. Especially under strong winds, open sea surfaces form rough surface boundaries. As a result, the reflection from rough sea surface makes signals follow different paths and even some of the transmitted signals are backscattered. Therefore, the sea surface roughness causes decrease in the received power strength and modeling of rough surface boundary is an important challenge for the ducting layer propagation modeling. There are some experimental studies that investigate the effects of rough sea surfaces on the wave propagation within ducts [66], but these experimental works do not provide enough data to model extreme surface conditions and there are hardly a few theoretical models for the sea surface roughness at the presence of atmospheric ducts. To this end, this topic requires both experimental and theoretical research attention.

Aerosols and weather conditions

The presence of aerosols such as bacteria, pollen and dust are highly probable in the lower atmosphere. The presence of such aerosols in the communication layer may create scattering and reflection which cause radar echoes and path-loss increase at the receiver side. In addition, hydro-meteors in the lower atmosphere, especially rain, can create significant path-loss due to its reflective effects between 10-20 GHz. To this end, rain and aerosol attenuation in the ducting layer is a promising open research field as well.

6.3 Channel Modeling Techniques for Atmospheric Ducts

Ducting layer wave propagation is complex due to the trapping effect because the trapped signals are in contact with both rough sea surface and duct layer. In addition to these factors, the ducting layer wave propagation depends on multiple channel parameters such as, antenna heights, duct height and type, carrier frequency and polarization. In this section, we are going to review the channel model approaches

to model the ducting layer communication by outlining open research fields. More importantly, we present preliminary simulation results for analyzing the path-loss exponent to indicate the received signal level improvements:

6.3.1 Parabolic Equation Methods

PE method utilizes parabolic approximation to Helmholtz wave equation. This technique is the dominant technique to model the ducting layer wave propagation with the capabilities of including both complex boundary conditions and refractivity variations of the lower atmosphere [67]. The existing numerical methods can provide fast and accurate results for PE. For these reasons, PE methods can be used especially for estimating the large scale path-loss in the ducting layer under different refractivity condition, polarization and surface types.

The remaining of the section includes the numerical methods to solve PE, available software tools and preliminary simulation results to show the signal level enhancement due to the atmospheric duct formation.

Numerical Methods for PE

PE can be solved by numerical operations and there are three popular numerical methods for solving PE: Split Step Fourier (SSF), Finite Difference (FD) and Finite Element (FE) [68]. The choice between these methods strongly depends on the scenario and conditions. The SSF method takes advantage of fast fourier transform (FFT) algorithms. Therefore, the implementation of PE with SSF is computationally efficient compared to other methods and, the technique can yield accurate and stable solutions. Secondly, FD provides highest resolution in modeling the boundary conditions by utilizing Crank-Nicholson finite difference scheme. Lastly, FE provides better modeling of the fast variations in the atmospheric conditions and provides flexibility to model the complex boundary conditions. A good review of these techniques can be found in [68].

Wave Propagation Tools

There are available wave propagation tools that can be used to model the ducting layer. The most important and widely used one is AREPS that is developed by the Atmospheric Propagation Branch at the Space and Naval Warfare Systems Center, San Diego [69]. AREPS is a hybrid wave propagation tool that combines ray optics and PE with SSF methods. It is capable of taking refractivity index as input from various sources like satellite, meteorological and bulk measurements. In addition, PETOOL, that is a free and on-line available tool, can model both forward and backward scattered waves under different terrain and refractivity condition [70]. Since PETOOL also considers the backward scatter waves unlike AREPS, it can be used to analyze regions with terrain block between the transmitter and the receiver such as islands. PETOOL is calibrated with using ray based methods and AREPS as described in [70]. Since, PETOOL can give reliable results like AREPS and it is a free on-line available tool, researchers may prefer PETOOL to model the wave propagation in the ducting layer. Furthermore, TEMPER, that is developed by the John Hopkins University Applied Physics Laboratory, is especially preferred for the R&D applications such as high fidelity radar design and 3D wave modeling applications. TEMPER is solely based on PE method [71]. Lastly, Wireless InSite, which is developed by REMCOM, is a hybrid model and it utilizes FD time domain method in order to solve PE. Although, this tool is not widely used in the ducting layer wave propagation modeling, it is a promising modeling tool for the projects that require modeling of complex boundary conditions.

Path-loss Analysis for Evaporation Duct

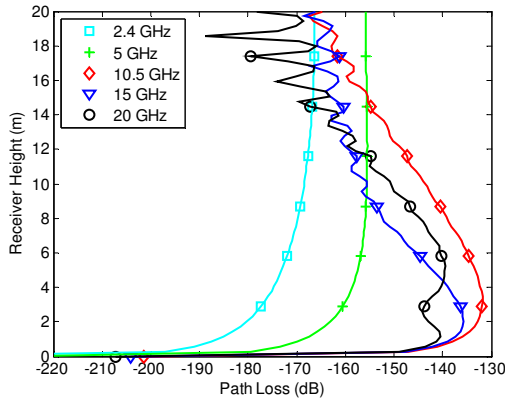
In this section, we present our preliminary analysis to show the received power enhancement by the atmospheric ducts. Since evaporation ducts are more probable compared to other types of ducts, we analyze an ideal 13 m evaporation duct which is modeled with a logarithmic curve as in [63]. We utilize PETOOL to solve PE for our preliminary large scale path-loss analysis.

Large scale path-loss in wireless communication can be represented with the generic formula as:

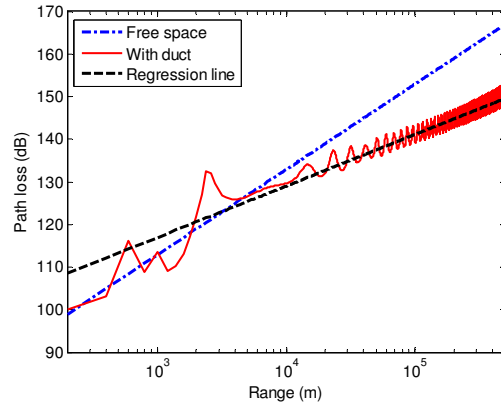
$$PL = A + 10\gamma \log d/d_0 + X_s \quad (6.3)$$

where γ is the path-loss exponent, d_0 is the b-LoS range, A is the path-loss at d_0 and X_s is the shadow fading.

Ducting conditions show significant dependence on the carrier frequency because the trapping condition becomes dominant above a certain frequency [62]. To this end, we simulate the wave propagation within a 13 m evaporation duct for different carrier frequencies (Transmitter is 4 m above sea level). Fig. 6.3(a) represents the receiver height vs. path-loss at 80 km distance. As noticed from this figure, 10.5 GHz can reach the lowest path-loss values and similar experimentally validated results can be found in [21]. Therefore, the signal level enhancements becomes highest in this high frequency range.



(a) Receiver height vs. path-loss.



(b) Path-loss vs. range.

Figure 6.3: Path-loss variations with respect to receiver height and range in 13 m ideal evaporation duct.

To show the signal level enhancement with respect to the free space, we present path-loss vs. range at 10.5 GHz for the transmitter and the receiver that are located

at 4 m above sea level. As in Fig. 6.3(b), path-loss in the ducting layer is significantly lower than the free space. Therefore, we can expect the path-loss exponent to be smaller than 2 (free space) because the signal energy is not spread through the atmosphere. Instead, it is concentrated within the ducting layer. The simple linear regression analysis with (6.3) yields the path-loss exponent of 1.2087 for ideal duct and flat sea surface (Black line in Fig. 6.3(b)). We expect that the path-loss exponent will be higher for the real world conditions with range dependent refractivity conditions and rough sea surface.

Although several works are available to model the amount of path-loss within the atmospheric ducts, there are no studies that provides the statistical modeling of the path-loss exponent and the shadow fading with respect to the atmospheric conditions, sea surface roughness, duct height, antenna heights and carrier frequency in the ducting layer communication. In addition, the effects of the polarization and the correlation between antennas are the open research fields which can be solved by PE simulations.

6.3.2 Ray Optics Methods

Ducting layer wave propagation modeling is dominated by PE because the ray analysis introduce significant errors at high ranges and they are much slower. However, the ray optics methods are still used in the hybrid models in order to provide fast and reliable estimates of the behavior of the wave-front and propagation angles [67].

The ray analysis utilize the solution for the Eikonal equation under nonstandard refractivity conditions and they can give the approximate behavior of the delay spreads and AoA of the multipath components. Since PE is inefficient to model these parameters which are essential to the channel modeling, the ray optics methods should be used in the ducting layer studies. There are some experimental studies that focus on AoA at the presence of the atmospheric ducts [72]. In [72], they also estimate AoA for the ducting layer by using both ray optics and PE methods. However, the distribution of delay spreads and AoA in the ducting layer are not well studied fields,

and they require significant experimental and theoretical research efforts. In addition, angle-of-departures (AoD) should be also studied for the ducting layer. Since only signals having low grazing angles can be trapped by the ducting layer, most of the transmitted power can spread through the atmosphere instead of the ducting layer if the beam-width of the antenna exceeds a certain value which depends on the antenna placement and the atmospheric conditions. To prevent signal spreading through the atmosphere instead of the ducting layer, the maximum beam-width that can be trapped by the atmospheric ducts should be derived with respect to transmitter antenna height, duct height and carrier frequency by utilizing the ray optics methods.

6.4 Applications Areas of the Ducting Layer *b-LoS* Communication

This section provides answers for “How the usage of the atmospheric ducts as a propagation medium can improve current state-of-art naval communication links?”. To this end, we review the possible application areas of the ducting layer *b-LoS* communications by especially focusing on the military naval applications.

6.4.1 Military applications

In current naval *b-LoS* communications, the transmitter and the receiver can use HF radio systems, or satellites or air units as a relay node to deliver data packages. However, HF radio systems are band-limited in their nature and they are inadequate to provide high data rates that modern military communications require. The other highly used option is the satellite communication which is expensive and introduces high communication delays. In addition, the satellite communication has security problems in the modern warfare because it can be highly affected by hostile jamming. Furthermore, using an air unit as a relay node is dangerous because the unit can be exposed to hostile attacks and the replacement of a relay node can waste considerable amount of mission time [61]. To this end, the ducting layer *b-LoS* communications is a promising alternative because the communication at the lower troposphere is less

prone to be affected by hostile jamming and also, the utilization of the ducting layer eliminates the need for an air unit as relay node. The ducting layer is expected to employ near 10 GHz as in Section 6.3, and with the signal level enhancements, it will be able to provide high data rates that modern military applications require.

In NCO, the connectivity between units are critical to empower strategical and tactical decisions. With the employment of the atmospheric ducts, there will be a direct communication channel between navy-navy and navy-low flying units (especially UAVs). Therefore, the employment of ad hoc networking with the b-LoS ducting layer can provide end-to-end connectivity between naval warfare. For example, a naval ship can relay an information that is coming from a LOS source to a b-LoS naval ship. To this end, the ad hoc b-LoS networks with the atmospheric ducts should be designed and, the distribution of data rates and relay delays should be derived for the ducting layer b-LoS systems. Since there are hardly a few works which uses the atmospheric ducts as a communication medium, the ad hoc networking with the ducting layer is a promising open research field to improve the naval NCO.

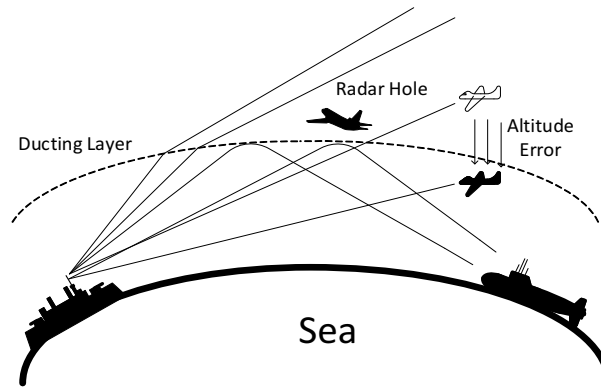


Figure 6.4: Wave Propagation with ducting layer.

6.4.2 Radar applications

Radar coverage is strongly affected by ducts in coastal and maritime environments [20]. Without considering the effects of ducts, the coverage can be miscalculated

and miscalculations impose unwanted problems like radar holes and altitude errors as shown in Fig. 6.4. Since the military radar applications require precise and accurate estimations, the effects of ducts should be taken into account in the coverage estimations. Therefore, the ducting layer studies also contribute to the radar coverage analysis as well.

6.4.3 Wireless Sensor Networks

Ducting layer b-LoS communications can be used to monitor civil or military sensor networks in maritime and coastal areas as in [21] because the implementation of satellite or cellular communication are expensive in large maritime areas. As a result, the ducting layer can provide low cost high data rate monitoring for the maritime wireless sensor networks.

Chapter 7

CHANNEL MODEL FOR SURFACE DUCTS: PATH-LOSS, DELAY SPREAD, AND AOA

7.1 Introduction

Atmospheric ducts are caused by non-standard atmospheric conditions, and the wave propagation in the presence of the atmospheric ducts is called as anomalous wave propagation [20, 60]. The atmospheric ducts can trap the propagating signals at the ducting layer that is the area between the ground and the atmospheric duct. In this way, most of the signal energy propagates in the ducting layer unlike the standard atmosphere. Therefore, the trapped signals can propagate through b-LoS distances with lower path-loss values.

The wave propagation and refractivity profile modeling of the atmospheric ducts are well-studied topics [22]. However, there is no work which considers the channel model for the atmospheric ducts. Therefore, we aim to develop a large-scale path-loss model for the surface ducts and analyze the delay spread and AoA of the surface ducts for the first time in the literature.

The contributions of this chapter is tri-fold. Firstly, we develop a statistical large-scale path-loss model for the surface ducts based on the PE simulations. The developed large-scale path-loss model is utilized to predict the distribution of the small-scale fading, and the communication ranges that can be provided by the surface ducts via link budget analysis. In addition, we develop a ray-optics (RO) method to calculate the ray trajectories with the surface ducts. By using the developed method, we analyze the delay spreads to determine the fading behavior of the channel. Available studies do not provide a realistic analysis for the delay spread of the ducting channels.

Lastly, we develop an analytical formula to determine the effective trapping beam-width, which represents the range of beam-widths that can be trapped by the surface ducts according to the channel conditions. The signals cannot be trapped by the surface ducts beyond the effective trapping beam-width, and their power propagates through the troposphere. The energy of these signals is wasted in the b-LoS systems. For this reason, we develop an analytical formula which can help both academia and system designers while designing efficient b-LoS communication links with the atmospheric ducts.

The remaining of the chapter is organized as follows. Section 7.2 provides the related works. Section 7.3 includes the refractivity model of the atmospheric ducts. In Section 7.4, we review both PE methods and RO methods to model the surface ducts. In addition, this section includes the derivation of the RO methods for the surface ducts and the effective trapping beam-width. Section 7.5 provides the derivation of the large-scale path-loss model for the surface ducts. In Section 7.6, the simulation results are presented for the delay spread, AoA and path-loss of the channel under various cases.

7.2 Related Work

The ducting channel studies can be divided into two main parts: refractivity estimation and wave propagation studies. The refractivity estimation techniques aim to predict the vertical refractivity gradient which is essential to model the wave propagation at the lower atmosphere. There are various methods for predicting the refractivity of the lower atmosphere: the direct measurement techniques like microwave refractometers and radiosonde balloons [73], LIDAR (Light Detection and Ranging) techniques [74], in situ bulk measurements and meteorological methods [75], GPS (Global Positioning System) measurements [76] and refractivity from clutter techniques [77, 78]. In this study, we utilize the refractivity measurement results presented in [79] which is based on the atmospheric measurements in Istanbul, Turkey and, [80] which is based on the microwave refractometer measurements with a heli-

copter in Wallops Island, Virginia, USA. We determine the parameters regarding to characterize the surface ducts as suggested in [79, 80].

Since the RO based techniques [81] or the normal-modes based techniques [81, 82] are unreliable or too resource consuming, the ducting channel propagation studies mostly focus on the PE methods which utilize the paraxial approximation to the wave equation. PE can be solved with numerical methods: SSF, FD and FE. Especially with the development of SSF methods [83], PE has become the most popular method to model the anomalous wave propagation [20, 21, 84, 85] because SSF methods provide fast and reliable results. In addition, PE methods can take the refractivity conditions as input and they can effectively model the complex boundary conditions. Good review of the ducting channel wave propagation modeling and SSF based PE method can be found in [22] and [68], respectively.

To model the wave propagation in the ducting channel, there are available wave propagation tools. The most widely used one is AREPS that is developed by the Atmospheric Propagation Branch at the Space and Naval Warfare Systems Center, San Diego [69]. The combination of RO and PE methods is also promising as suggested in [86–89]. Therefore, AREPS utilizes both the RO and SSF based PE methods. In addition, PETOOL [70], which is also based on SSF, is capable of modeling both forward and backward scattered waves under different terrain and refractivity conditions. Therefore, PETOOL is promising especially for the regions with irregular terrain conditions. Since PETOOL is a free and on-line available tool and it is calibrated with RO methods and AREPS as described in [70], we utilize PETOOL to model the surface ducts. In addition, we make RO analysis to the surface ducts to estimate the delay and AoA of the channel. Therefore, the characteristics of the channel can be estimated with hybrid methods as suggested in [86, 87].

There are experimental studies about the ducting channel wave propagation. [21] provides a high data rate employment of the atmospheric ducts for b-LoS sensor networks. In [21], the authors observed that the wave propagation analysis performed in AREPS has strong consistency with the experimental results. In addition, [90–94]

compare the simulation results with the PE methods and the experimental measurements for an 83 km link at 4.7, 10 and 15 GHz. According to their results, the region had strong evaporation ducts, and the experimental path-loss measurements were consistent with the PE simulations. Furthermore, [20] provides one of the most extensive review of the ducting channel with the comparison of the PE based results and experimental results for the surface-based ducts. The results of this work show that the surface-based ducts can provide 10 – 20 dB less path-loss compared to free space at b-LoS distances. Similar experimental results to verify PE simulations is presented in [95] as well. To sum up, the results of the PE methods are validated through a number of different experimental works for different type of ducts. As a result, the theoretical results that are generated with the PE based tools can be considered as reliable based on previous experimental validations.

7.3 Refractivity

The atmospheric ducts are formed by the non-standard atmospheric conditions. The formation and characteristic of the atmospheric ducts depend on the refractivity of the lower atmosphere. Therefore, we review the refractivity and modified refractivity to characterize the atmospheric ducts.

Tropospheric radio refractive index (n) changes slightly with the atmospheric parameters: wind, temperature, pressure and humidity at the lower atmosphere. n varies in the order of 10^{-3} . Therefore, refractivity is utilized instead of the refractive index and refractivity (N) is given as [62]

$$N = (n - 1) \times 10^6 \text{ N-units}, \quad (7.1)$$

where n is the atmospheric refractive index.

The trapped signals in the ducting layer has low grazing angles. Therefore, the refractivity model should consider the curvature of the earth. Thus, the modified

refractivity (M) is defined as [62]

$$\begin{aligned} M &= N + \left(\frac{h}{R_0} \right) \times 10^6, \\ &= N + 157 \times h \text{ M-units}, \end{aligned} \quad (7.2)$$

where h is the height above the surface level in km and R_0 is the earth radius in km.

According to the modified refractivity gradient, there are four different refractive conditions. In the sub-refraction condition ($\partial M/\partial z > 157$ M-units/km), the signal rays are refracted away from the ground through the atmosphere. Standard refractivity condition is associated with ($157 \text{ M-units/km} \geq \partial M/\partial z > 78 \text{ M-units/km}$) and this condition represents the standard wave propagation. The signals propagate downward in the super-refraction condition ($78 \text{ M-units/km} \geq \partial M/\partial z > 0 \text{ M-units/km}$). Most importantly, the trapping condition, which is also known as the *ducting condition*, is associated with $0 \geq \partial M/\partial z$, and the signal rays are trapped between the surface and the ducting layer. This effect can make the trapped signals to propagate over-the-horizon with low path-loss. For the formation of the atmospheric ducts, the humidity is an essential factor [22]. Thus, the ducting channel is a promising candidate for the b-LoS communications especially in the coastal and maritime environments as reviewed in [22].

There are four types of atmospheric ducts according to their formation process and refractivity profiles: evaporation ducts, surface-based ducts, elevated ducts, and surface ducts. The modified refractivity of the evaporation ducts can be modeled with a logarithmic function [63] as in Figure 7.1(a). Surface-based and elevated ducts are modeled with a tri-linear curve as in Figure 7.1. Surface ducts are modeled with a bi-linear curve as in Figure 7.1(c). Since the evaporation ducts are modeled with the logarithmic function, the evaporation ducts are complex to be analyzed with the RO based tools. As a result, we only consider the surface ducts in this chapter. However, an evaporation duct and a surface duct having the same duct height is expected to have similar maximum delay spread and AoA because their refractivity profiles are closer as in Figure 7.1. Thus, the results presented for the surface ducts can be utilized

to predict the the characteristics of the evaporation ducts as well.

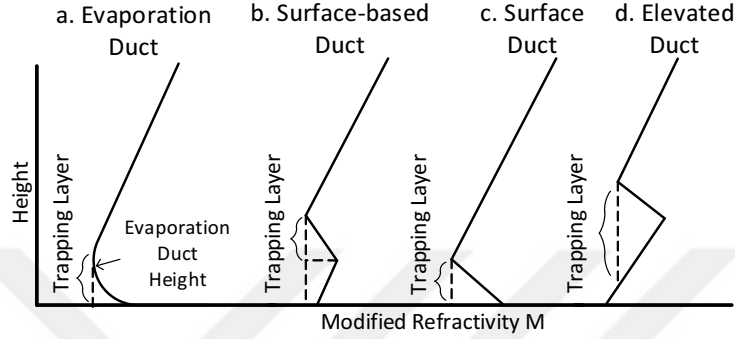


Figure 7.1: Modified refractivity profiles of atmospheric ducts.

In this chapter, the surface ducts are characterized with two atmospheric parameters: duct height (δ) and duct strength (ΔM). Duct height is the height where the gradient of the modified refractivity changes its sign as in Figure 7.1(c). Duct strength is the amount of change in the modified refractivity from the bottom to the top of the surface duct. The modified refractivity above the surface duct is assumed as the standard condition and decreases with 157 M-units/km. We utilize the experimental surface duct measurements [79, 80] to determine the duct height and duct strength parameters.

7.4 Channel Modeling Techniques for Surface Ducts

Ducting channel wave propagation requires the modeling of atmospheric parameters and complex boundary problems. For this reason, PE methods are preferred for estimating the path-loss in the ducting channel. In addition, we utilize RO methods to analyze the delay spread and AoA of the ducting channel with the surface ducts. In this section, we first review PE, and develop a RO based method to analyze the delay spread and AoA of the ducting channel.

7.4.1 Parabolic Equation Methods

PE methods use paraxial approximation to the Helmholtz wave equation, and it was originally developed in [96]. PE methods are capable of modeling the complex boundary conditions and the refractivity variations of the lower atmosphere [67]. Especially with the derivation of a simple solution to PE based on SSF method [83], PE methods became the dominant technique to model the wave propagation in the ducting channels. Although there are other numerical methods to solve PE: FD and FE, SSF method is more preferable with its accuracy and computational efficiency [20, 22, 68, 69, 97].

The formulation of the PE for the electromagnetic problems is summarized in [98]. The theory of PE is explained in detail in [20, 67, 99]. Interested reader may refer to these references. However, we do not discuss the theory of PE in this chapter. Instead, we review how the path-loss calculations are performed with the PE methods. For 2D narrow angle forward scatter waves, PE in scalar form is given as [68]

$$\frac{\partial u(x, z)}{\partial x} = \frac{i}{2k} \frac{\partial^2 u(x, z)}{\partial z^2} + \frac{ik}{2} (m^2(x, z) - 1) u(x, z), \quad (7.3)$$

where k is the wave number, $m = 1 + M10^{-6}$ is the modified refractive index, $u(x, z)$ is the reduced propagation direction, x represents the horizontal axis, and z represents the vertical axis (height). (7.3) can be used for both horizontal and vertical polarization.

We utilize PETOOL [70], which is a free on-line available tool based on the PE method with SSF, to solve (7.3). As discussed in Section 7.2, PETOOL is calibrated with AREPS which is an experimentally validated tool. With the PE methods, the path-loss (PL) can be found by using $u(x, z)$ as [67]

$$PL = 20 \log(4\pi) - 20 \log |u(x, z)| + 10 \log(r) - 30 \log(\lambda), \quad (7.4)$$

where, λ is the wavelength, and r is the range. Unlike available propagation studies, we utilize PE methods to develop a large-scale path-loss model for the first time in the literature in Section 7.5.

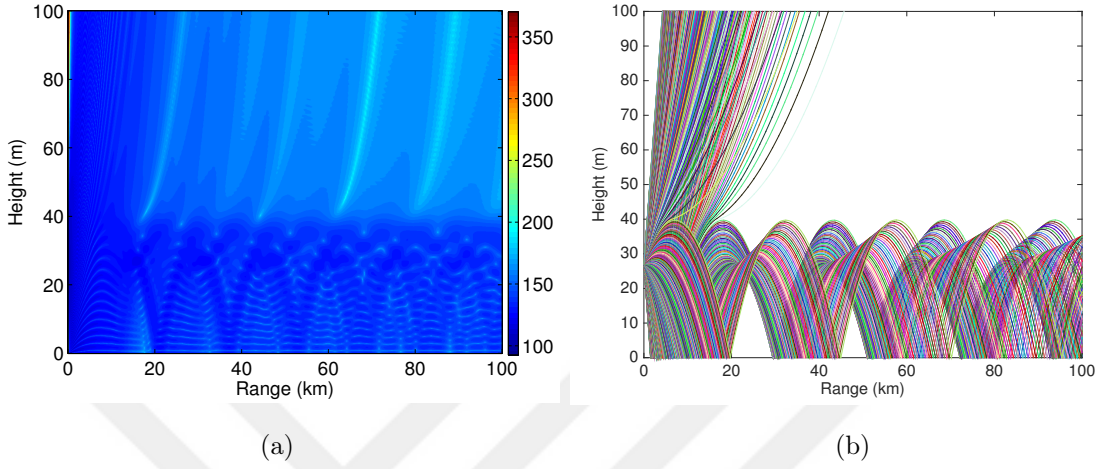


Figure 7.2: Wave propagation results with (a) PETOOL, (b) RO methods for 40 m surface duct, 10 GHz, 27 m transmitter height.

7.4.2 Ray-optics Methods

Although PE methods dominate the ducting channel wave propagation modeling, RO methods can be utilized to estimate delay spreads and propagation angles [17, 67] as in the hybrid models [69, 86]. In addition, [72] provides the comparison of the AoA predictions with the RO methods and the experimental observations. According to their results, the predicted AoA values were consistent with the observations. Since the PE methods are not capable of estimating the delay spreads and AoAs, RO methods are promising for the channel modeling in the ducting layer.

RO methods utilize the well-known Snell's equation. By solving the Snell's equation for the stratified atmosphere, the Eikonal equation is derived to calculate the trajectory of each ray. The trajectories of the rays are utilized to calculate the delay spreads and AoA of the channel. There are some studies in the literature which focus on the ray-tracing simulations to model the wave propagation in the atmospheric ducts. In [100, 112], the ray-tracing analysis are performed to estimate the delay in the elevated ducts, but their models assume that only three rays can reach to the receiver. According to these papers, an 80 km link with elevated ducts shows a few nanosecond

delay spread. In [101,102], they assume that odd number of rays can reach to the receiver and estimate AoA with the ray-tracing simulations for the surface-based ducts. Although the existing studies have significant results for the ducting channel, more realistic results can be generated with the RO methods. Therefore, we utilize RO methods to estimate the fading characteristics of the channel and the distribution of the AoA for the surface ducts. Lastly and more importantly, we derive an analytical expression for the maximum and minimum value of AoD which can be trapped by the surface ducts. With the derived expression, the transmitter can be designed to maximize the efficiency of the system by only propagate in the calculated beam-width span.

The remaining of the subsection includes the derivation of the ray trajectory formula and the derivation of the trapping beam-width:

Ray Trajectories

The ray trajectories in the atmospheric ducts can be estimated with the RO methods [17]. However, RO methods become more reliable above 3 GHz [62]. Since the evaporation ducts perform with the lowest path-loss near 10 GHz [21,22,84], we aim to focus 5 – 15 GHz frequency range in this chapter. The main drawback of RO methods is the inefficiency to include the effects of frequency. However, the results generated with RO methods show high consistency with PE results as in Figure 7.2 as suggested in [72].

In [103], the earth is assumed as flat for the ray-tracing simulations. However, the curvature of the earth becomes important at the b-LoS distances. Therefore, we utilize the earth radius transformation and the Snell's law to find the Eikonal equation for the surface ducts by assuming the atmosphere has infinitesimal stratified layers. In this way, the ray trajectories can be estimated with the Eikonal equation which is represented as

$$\frac{1}{n(z)} \frac{dn}{dz} + \frac{1}{R_0} = \frac{d^2 z}{dx^2}, \quad (7.5)$$

where R_0 is the radius of the earth which is used as 6370 km, and the derivation

of (7.5) can be found in Appendix 13-E. $n(z)$ shows very slight variations with the height which is in the order of 10^{-3} [19]. Therefore, (7.5) is simplified as

$$\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} = \frac{d^2 z}{dx^2}, \quad (7.6)$$

where $n(0)$ is used as 1.00035 [19].

Since the surface ducts have a bi-linear refractivity model as in Figure 7.1, (7.5) can be solved as described in the Appendix 13-F and the ray trajectories are calculated as

$$z = \frac{1}{2} \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) x^2 + \theta x + h_t, \quad (7.7)$$

where θ is the angle of the ray with respect to the ground at the transmitter. (7.7) can be solved step by step for each ray to calculate the trajectory of the trapped paths.

Figure 7.2(b) shows the ray trajectories for a 40 m surface duct ($\Delta M = 20$ M-unit), and Figure 7.2(a) presents the path-loss results for the same ducting conditions by using PETOOL with the following channel parameters: 27 m transmitter, horizontal polarization, and 10 GHz. As noticed, the regions with high ray densities are associated with the low path-loss values. In addition, the shadowed regions are located at the similar locations at both figures. Therefore, PE methods and RO methods give consistent results.

Trapping Beam-width

According to the channel geometry and the atmospheric conditions, only the rays that have certain angles can be trapped by the atmospheric ducts. The rays outside of this certain range cannot be trapped and propagate through the atmosphere. The energy of these rays are wasted in the b-LoS systems. To increase the efficiency, we develop an analytical model to estimate the effective trapping beam-width at the transmitter side for the first time in the literature.

By utilizing the proposed ray trajectory formula (7.7), we derive the span of the effective trapped beam-widths which depends on the change in the refractive index,

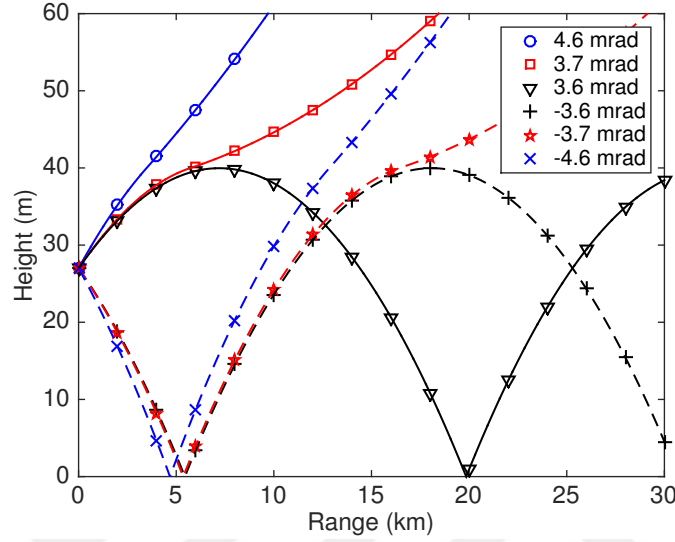


Figure 7.3: Ray trajectories with the ray-tracing method.

transmitter height and duct height. The derivation of the maximum and minimum trapping beam-widths ($\theta_{max,min}^T$) can be found in Appendix 13-G, and $\theta_{max,min}^T$ are found as

$$\theta_{max,min}^T = \pm \sqrt{2 \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) (h_t - \delta)}. \quad (7.8)$$

In the trapping layer, the refractive index decreases with the height, and this implies that $\frac{dn}{dz} < 0$. In addition, our model assumes that the transmitter is located within the surface duct ($h_t < \delta$). Therefore, the term inside the square-root in (7.8) is positive, and it has two real roots.

According to (7.8), a 40 m surface duct with $\Delta M = 20$ has $\theta_{max,min}^T = \pm 3.6 \text{ mrad} = \pm 0.207^\circ$ where the transmitter height is $h_t = 27 \text{ m}$. We also perform RO simulations based on (7.7) to verify our analytical trapping beam-width formula. Figure 7.3 shows the ray trajectories for the angles that are on the boundaries. As noticed, the analytically calculated trapping beam-width is consistent with the simulations. As a result, adaptive antenna techniques can be designed for the ducting channel systems to maximize the trapped power with this equation because the signal transmitted out of this beam-width region cannot be trapped by the surface ducts. In addition, for

the symmetrical ducting channels, where $h_t = h_r$ and the refractivity profile is range independent, the calculated trapping beam-widths form an upper and lower bound for the AoA in the ducting channel.

7.5 Large-scale Path-Loss Model for Surface Ducts

The ducting channel wave propagation depends on many factors: transmitter height, receiver height, duct height, duct strength, carrier frequency, polarization and surface conditions. PE based methods are capable of taking all of these parameters as input. Thus, we utilize PETOOL to solve PE to estimate the path-loss in the surface ducts for the given channel parameters. We aim to develop a path-loss model as a function the channel parameters: transmitter height, receiver height, duct height, duct strength and carrier frequency for the first time in the literature. Our model will be independent of the polarization because the effects of the polarization is not noticeable in the large-scale, and we use only the horizontal polarization for both of the antennas. In addition, the surface is assumed as smooth sea surface.

We aim to model the path-loss in the ducting channel for the b-LoS distances where the Line-of-Sight (LoS) transmission is impossible in the standard atmospheric conditions. For the b-LoS ducting channel, the large-scale path-loss equation is given as

$$PL = A + 10\gamma \log d/d_0 + X_s, \quad (7.9)$$

where γ is the path-loss exponent, d_0 is the b-LoS range, A is the path-loss at d_0 and X_s is the shadow fading. The b-LoS range is given by the well-known maximum LoS range equation for the standard refractivity conditions as

$$d_0 = 4.12\sqrt{h_t} + 4.12\sqrt{h_r}, \quad (7.10)$$

where h_t and h_r are the transmitter and receiver heights, respectively.

In this section, we develop a statistical large-scale path-loss model based on the PE simulations performed with PETOOL. In addition, the distribution of the shadow fading and the coefficients of the channel parameters are determined by using the

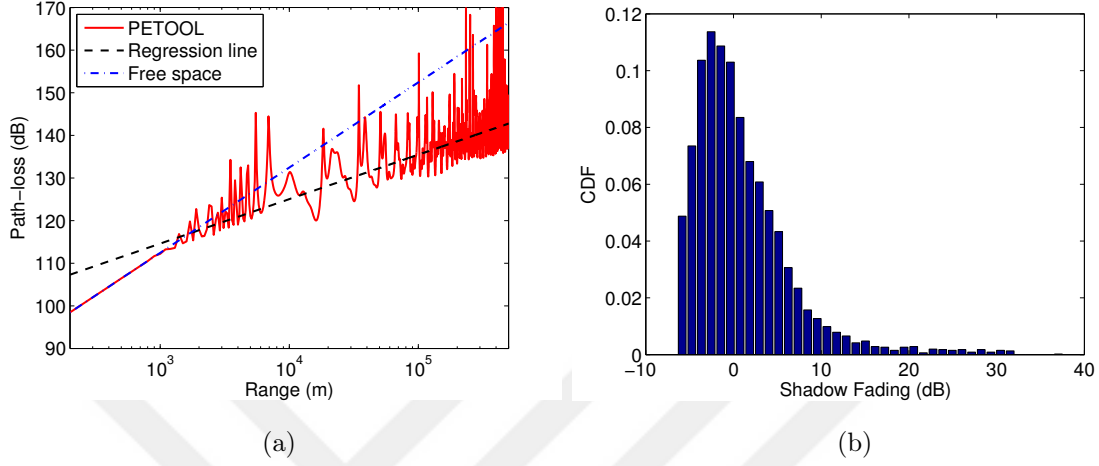


Figure 7.4: The path loss results for (a) the regression method and, (b) the distribution of the shadow fading.

multivariate linear regression method to (7.9). To this end, we perform PE simulations for the path-loss of the link for different channel conditions. Figure 7.4(a) contains the path-loss results generated with PETOOL, the free space path-loss and the fitted line to the PETOOL results for (7.9) with simulation parameters: 500 km range, 40 m surface duct, 27 m transmitter and receiver, 10 GHz carrier frequency, 20 M-unit duct strength and horizontal polarization. For the regression line fitting for (7.9), we utilize the least-square fitting. The deviations from the fitting line is treated as shadow fading, and the distribution of the shadow fading can be seen in Figure 7.4(b). The shadowed regions can be clearly seen in Figure 7.2.

Since the path-loss enhancements with the atmospheric ducts are most effective when both of the antennas are in the ducting layer [95], we assume that both the transmitter and the receiver are located within the surface duct. To develop the large-scale path-loss model, we perform extensive simulations by varying the channel parameters: carrier frequency ($f = 5 - 15$ GHz), duct height ($\delta = 10 - 60$ m), duct strength rate ($\Delta M/\delta = 0.1 - 0.3$ M-unit/m), transmitter height ($h_t = 0 - \delta$ m) and receiver height ($h_r = 0 - \delta$ m). The remaining of the section includes the derivation

of the coefficients and the statistical distribution of the large-scale path-loss model parameters: A , γ and X_s .

7.5.1 A Parameter

After performing a number of simulations by using PETOOL, we try to find the statistical distribution of the each parameter. Since the ducting channel that we are trying to model is symmetric, the effect of the transmitter and receiver antenna heights can be modeled with the difference between them $\Delta h = h_t - h_r$. We use the multivariate regression analysis in the least-square error sense to find the coefficients of the each channel parameter. A parameter is modeled as

$$\begin{bmatrix} 1 & \vdots & \vdots & \vdots & \vdots \\ \vdots & \Delta h_i & \delta_i & f_i & \Delta m \\ 1 & \vdots & \vdots & \vdots & \vdots \end{bmatrix} \begin{bmatrix} \alpha_A \\ \beta_A \\ \kappa_A \\ \xi_A \\ \varrho_A \end{bmatrix} = \begin{bmatrix} \vdots \\ A_i \\ \vdots \end{bmatrix}, \quad (7.11)$$

where Δh is the height difference between the transmitter and the receiver, δ is the duct height, f is the carrier frequency, Δm is the change in the modified refractivity index per m and A_i is the resulting value for each simulation case. By performing multivariate regression analysis in the least-square error sense, the resulting coefficients are found as $\alpha_A = 123.34$ dB, $\beta_A = 0.015$ dB/m, $\kappa_A = 0.13$ dB/m, $\xi_A = 8.66 \times 10^{-4}$ dB/MHz and $\varrho_A = -12.83$ dB×m. The variation (error) of A from the regression model is found as

$$\Delta A = \begin{bmatrix} \vdots \\ A_i \\ \vdots \end{bmatrix} - \begin{bmatrix} 1 & \vdots & \vdots & \vdots & \vdots \\ \vdots & \Delta h_i & \delta_i & f_i & \Delta n \\ 1 & \vdots & \vdots & \vdots & \vdots \end{bmatrix} \begin{bmatrix} \alpha_A \\ \beta_A \\ \kappa_A \\ \xi_A \\ \varrho_A \end{bmatrix}. \quad (7.12)$$

Figure 7.5(a) represents the deviation of A from our model and ΔA clearly shows a Gaussian distribution with the zero mean and standard deviation $\sigma_{\Delta A} = 2.39$ dB.

Therefore, A can be represented as

$$A \text{ dB} = \alpha_A + \beta_A \Delta h + \kappa_A \delta + \xi_A f + \varrho_A \Delta m + \sigma_{\Delta A} x, \quad (7.13)$$

where x is zero mean unit variance Gaussian random variable.

7.5.2 γ Parameter

By the same method, γ parameter is modeled as

$$\begin{bmatrix} 1 & \vdots & \vdots & \vdots & \vdots \\ \vdots & \Delta h_i & \delta_i & f_i^2 & \Delta m \\ 1 & \vdots & \vdots & \vdots & \vdots \end{bmatrix} \begin{bmatrix} \alpha_\gamma \\ \beta_\gamma \\ \kappa_\gamma \\ \xi_\gamma \\ \varrho_\gamma \end{bmatrix} = \begin{bmatrix} \vdots \\ \gamma_i \\ \vdots \end{bmatrix}. \quad (7.14)$$

According to the regression results, the resulting coefficients is found as $\alpha_\gamma = 1.3$, $\beta_\gamma = 2 \times 10^{-3} \text{ 1/m}$, $\kappa_\gamma = -2.9 \times 10^{-3} \text{ 1/m}$, $\xi_\gamma = -3.47 \times 10^{-10} \text{ 1/MHz}$, $\varrho_\gamma = -0.58 \text{ m}$. The coefficients β_γ and κ_γ result in too low values when they are multiplied with their parameters. Therefore, the effects of these factors can be neglected at the regression method. In (7.14), the frequency term is squared in order to make the error term as close as possible to the Gaussian distribution and as can be seen in Figure 7.5(b), $\Delta\gamma$ has a Gaussian distribution with zero mean and $\sigma_{\Delta\gamma} = 0.066$. In addition, the negative ϱ_γ indicates that increasing duct strength reduces the path-loss exponent. Therefore, γ can be modeled as

$$\gamma = \alpha_\gamma + \xi_\gamma f^2 + \varrho_\gamma \Delta m + \sigma_{\Delta\gamma} y, \quad (7.15)$$

where y is zero mean unit variance Gaussian random variable.

7.5.3 Shadowing Fading

As shown in Figure 7.4(a), there is a significant shadow fading in the link, and also the distribution of the shadow fading is shown in Figure 7.4(b). The best fitted distribution for the shadow fading in the channel can be found with the Kullback–Leibler

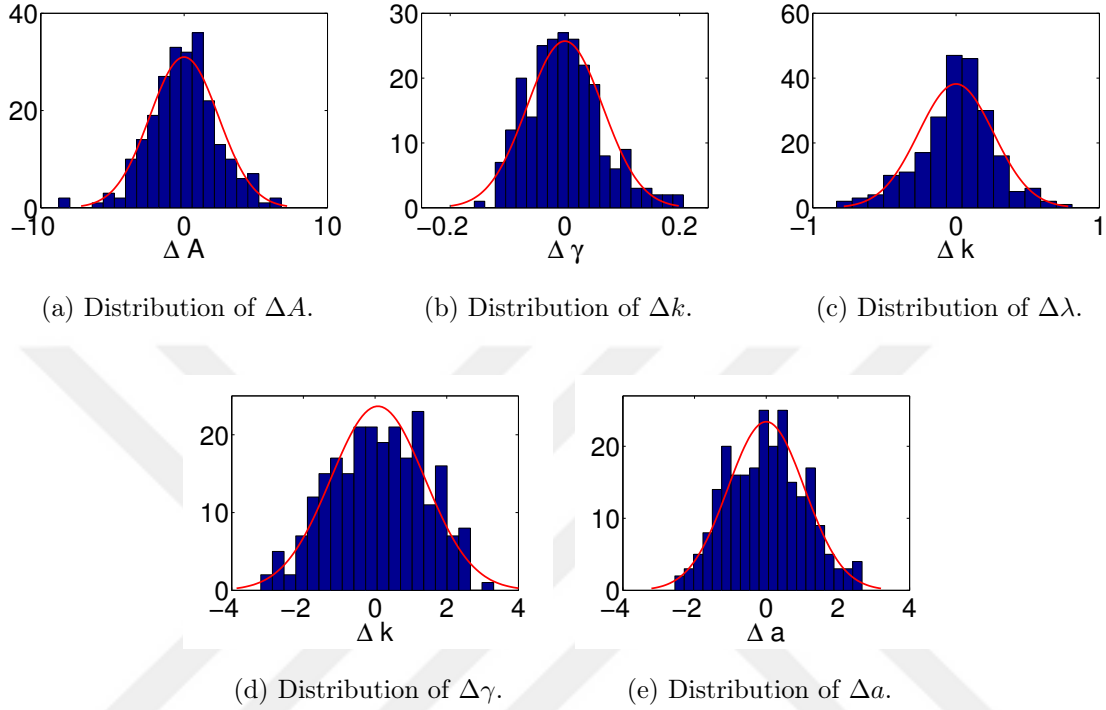


Figure 7.5: The deviations of the parameters from the regression models.

divergence (KLD) [104]. Let P and Q be two distributions and the divergence of Q from P can be calculated as

$$D_{KL}(P||Q) = \sum_i \ln \frac{P(i)}{Q(i)} P(i). \quad (7.16)$$

$D_{KL}(P||Q)$ takes lower values when the two distribution get closer, and it becomes 0 for the identical distributions.

We generate various distributions with MATLAB and use (7.16) to determine how close the generated distributions to the shadow fading are. The simulation results for the possible distributions can be found in Table 7.1. According to our results, the Weibull distribution best matches with the shadow fading for the surface ducts. Therefore, the shadow fading in the channel is modeled with log-Weibull distribution which is also known as Gumbel distribution.

Table 7.1: The KLD Results.

Type of Distribution	KLD
<i>Normal distribution</i>	1.3436
<i>Rayleigh distribution</i>	0.1954
<i>Weibull distribution</i>	0.0209
<i>Gamma distribution</i>	0.0283
<i>Generalized extreme value distribution</i>	0.0887
<i>Nakagami distribution</i>	0.0574

The PDF of the Weibull distribution can be represented as

$$f(x; \lambda, k) = \begin{cases} \frac{k}{\lambda} \left(\frac{x}{\lambda}\right)^{k-1} e^{-(x/\lambda)^k}, & x \geq 0 \\ 0, & x < 0 \end{cases} \quad (7.17)$$

where $k > 0$ is the shape parameter and $\lambda > 0$ is the scale parameter. We perform the multivariate regression analysis to find the distribution of k and λ as well.

We use the same methodology as in A and γ parameters. Again, the parameters with slight effects are neglected and only the final results are presented for the remaining multivariate fitting results. With the multivariate linear regression analysis, the k parameter can be given as

$$k = \alpha_k + \xi_k f^3 + \varrho_k \Delta m + \sigma_{\Delta k} z, \quad (7.18)$$

where z is the zero mean and unit variance Gaussian random variable. $\alpha_k = 0.1792$, $\xi_k = 3.36 \times 10^{-14}$ 1/MHz and $\varrho_k = 1.85$ m. The deviation of k from our model can be found in Figure 7.5(c), and this deviations can be modeled with Gaussian distribution with zero mean $\sigma_{\Delta k} = 0.26$.

The λ parameter can be modeled as

$$\lambda = \alpha_\lambda + \beta_\lambda \delta + \xi_\lambda f^3 + \varrho_\lambda \Delta m + \sigma_{\Delta \lambda} q, \quad (7.19)$$

where q is the zero mean and unit variance Gaussian random variable. $\alpha_\lambda = -1.94$, $\kappa_\lambda = 0.176$ 1/m, $\xi_\lambda = 1.09 \times 10^{-12}$ 1/MHz. Figure 7.5(d) shows the deviations of the λ from our model, and it has Gaussian distribution with zero mean and $\sigma_{\Delta\lambda} = 1.45$.

As noticed in Figure 7.4(b), the shadow fading values can be negative. However, since the Weibull distribution takes only positive values, we first shift the shadow fading to the right by Δa and perform the distribution fitting. To this end, Δa can be represented as

$$\Delta a \text{ dB} = \alpha_a + \beta_a \delta + \kappa_a f + \varrho \Delta m + \sigma_{\Delta a} w, \quad (7.20)$$

where w is the zero mean and unit variance Gaussian random variable. The resulting values of the $\alpha_a = 3.27$ dB, $\kappa_a = -0.14$ 1/m, $\xi_a = 3.2 \times 10^{-4}$ 1/MHz, and $\sigma_{\Delta a} = 1.07$ dB as can be seen in Figure 7.5(e).

7.5.4 The large-scale Path-Loss Model

By combining (7.9), (7.13), (7.15), and (7.20), the statistical large-scale path-loss equation is found as

$$\begin{aligned} PL &= A + 10\gamma_{\Delta h} \log d/d_0 + X_s \\ PL &= \alpha_A + \beta_A \Delta h + \kappa_A \delta + \xi_A f + \varrho_A \Delta m + \sigma_{\Delta A} x + \\ &\quad + 10(\alpha_\gamma + \xi_\gamma f^2 + \varrho_\gamma \Delta m + \sigma_{\Delta\gamma} y) \log d/d_0 + \\ &\quad W + \alpha_a + \beta_a \delta + \kappa_a f + \varrho \Delta m + \sigma_{\Delta a} z, \end{aligned} \quad (7.21)$$

where x , y and z are the zero mean unit variance independent normal random variables and W has the Weibull distribution with k in (7.18) and λ in (7.19).

7.6 Simulation Results

In this section, we present the simulation results with the PE methods and the RO methods. In addition, the results of the large-scale path-loss model is compared with the PE results. We present the link budget calculations for the surface ducts to predict the communication ranges for given channel parameters.

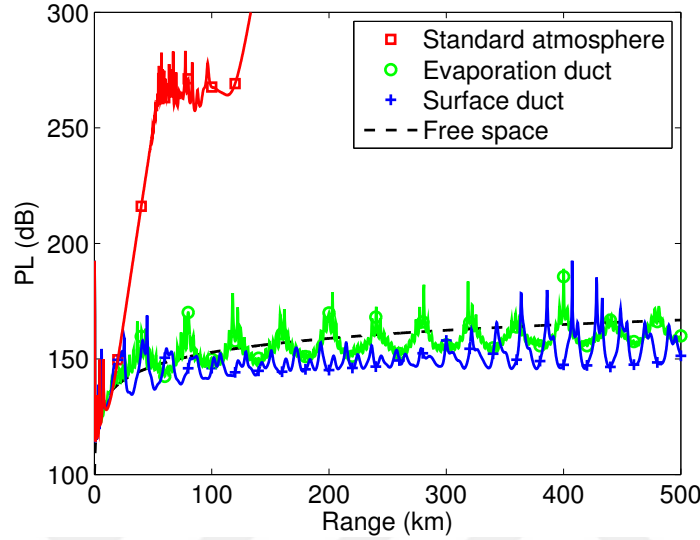
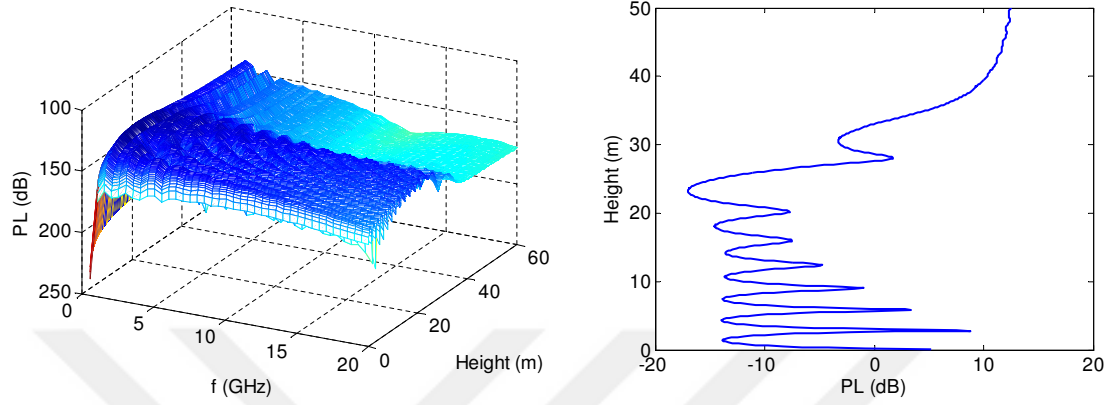


Figure 7.6: Path-loss vs. range for standard atmosphere, free space, and atmospheric ducts.

7.6.1 Simulation Results with PEM

We utilize PETOOL to perform the PE simulations as described in Section 7.4.1. Figure 7.6 presents the path-loss vs. range for different atmospheric ducts, the free space and the standard atmosphere at 10.5 GHz, 5 m transmitter and receiver, and horizontal diversity. As noticed, the curves for 20 m evaporation duct and 35 m surface duct ($\Delta M = 20$ M-unit) outperform the standard atmosphere in terms of path-loss values. In addition, the surface ducts have strong shadow fading compared to the evaporation ducts.

Figure 7.7(a) shows the path-loss for a 40 m surface duct (20 M-units) as a function of the carrier frequency and receiver height. As noticed, the lower frequencies 3 – 5 GHz have significantly higher trapping capabilities compared to the evaporation ducts [22,84]. This is one the main difference between the surface ducts and the evaporation ducts in terms of channel modeling because the evaporation ducts provide the lowest path-loss values near 10 GHz. Surface ducts show higher trapping capabilities over the evaporation ducts due to the sharp refractivity curve of the surface ducts as described



(a) Propagation-loss vs. receiver height and frequency at 100 km range. (b) Receiver height vs. propagation-loss at 100 km range.

Figure 7.7: Propagation-loss results for 35 m surface duct, 27 m transmitter height, 100 km range with PETOOL.

in Section 7.3.

Figure 7.7(b) presents the path-loss values as a function of the receiver height. As noticed, the path-loss values for the surface duct is significantly lower compared to the free space path-loss up to 18 dB. However, the power levels have fading with respect to the vertical height. This condition is caused by the constructive/destructive interference of the delayed signals. The delay spread of the channel should be examined in order to determine the fading behavior of the channel. Therefore, the next subsection presents the RO results to analyze the delay spread of the ducting channel.

7.6.2 Simulation Results with Ray-optics Methods

By using the calculated ray trajectories as in Figure 7.2(b), the delay spreads and AoAs of the received rays can be calculated. Figure 7.8(a) presents $1/T_{max}$ in Hz vs. the range for 27 m transmitter and receiver with a 40 m surface duct ($\Delta M = 20$ M-units) where T_{max} is the maximum delay spread. As noticed from the figure, $1/T_{max}$ value is higher than 200 MHz line up to 400 km. Since the channel bandwidth should

be much lower than $1/T_{max}$ for a channel to have flat fading, the ducting channel can be assumed as a flat fading channel up to 400 km with 20 MHz bandwidth.

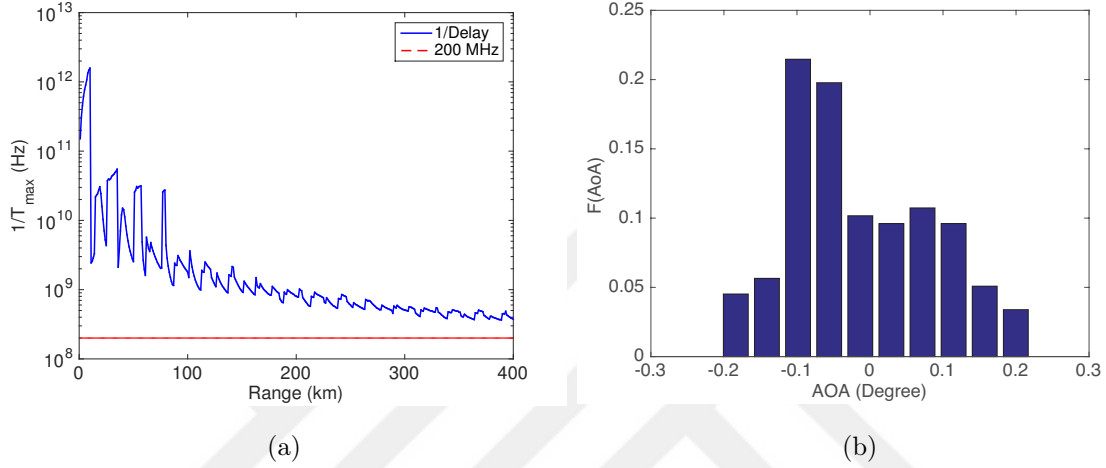


Figure 7.8: (a) $1/T_{max}$ spread, and (b) histogram of AoA for a 40 m surface duct.

Figure 7.8(b) presents the AoA of the ducting channel for 27 m transmitter and receiver with a 40 m surface duct ($\Delta M = 20$ M-units). As noticed, the AoA results are bounded by the analytical trapping beam-width value of 0.266° calculated previously in Section 7.4.2.

7.6.3 Large-scale Path-loss Model

We compare the results of our large-scale path-loss model with PETOOL for verification. The following channel parameters are utilized in the simulations: $f = 10$ GHz, $h_t = h_r = 20$, $\delta = 40$ m, $\Delta M = 20$ M-unit, horizontal polarization. Figure 7.9 includes the comparison of the results in the upper figures. As noticed, our model can predict the level of the path loss for each range. The lower figures show the distribution of the deviations from our proposed model which can be modeled as the small-scale fading. As noticed, the normalized deviations can be modeled with Rayleigh distribution, and the distribution becomes closer to the Rayleigh distribution with the increasing range.

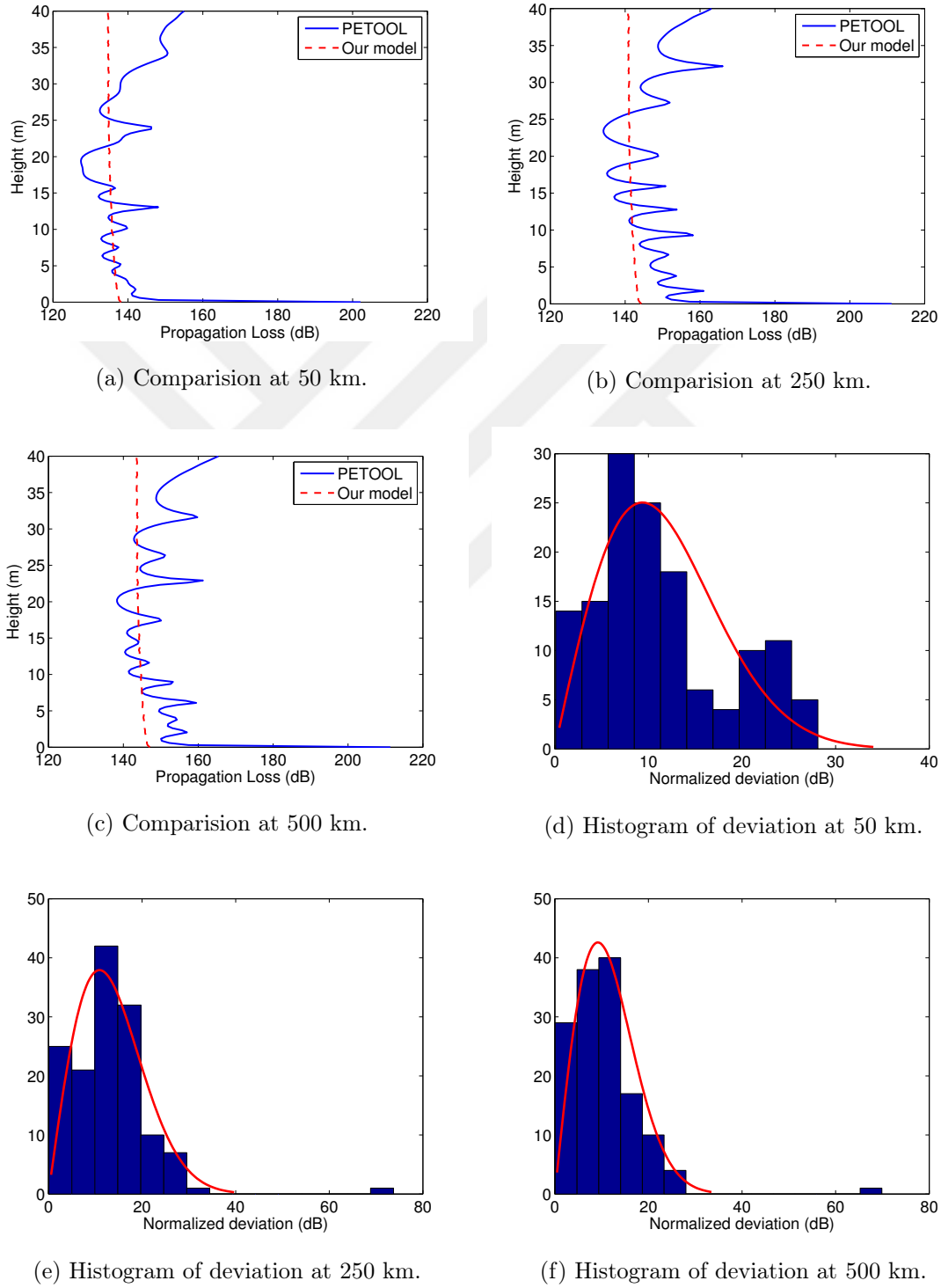


Figure 7.9: The comparison of our large-scale path-loss model and PETOOL results.

7.6.4 Link Budget Calculations

The link budget calculations are presented in this section, and the following parameters are utilized in the b-LoS channel with the surface ducts:

- Transmit power (P_t) = 30 dBm (1 W)
- Antenna gains ($G_{t,r}$) = 30 dB
- Cable losses (L_q) = 10 dB

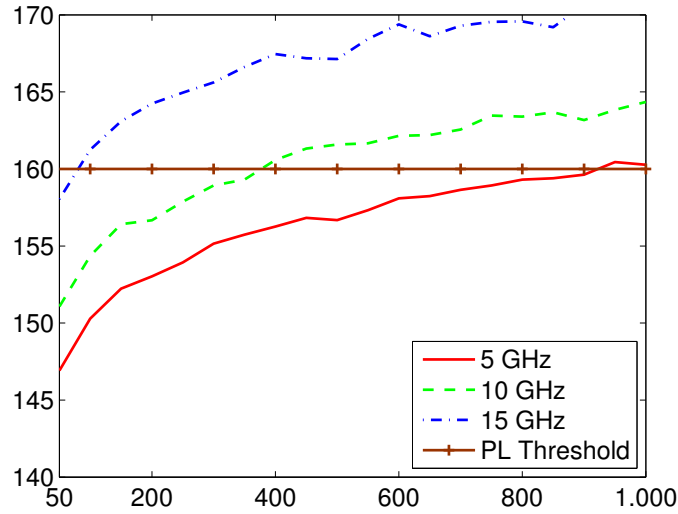


Figure 7.10: Link budget analysis for surface ducts.

Received power is given by

$$P_{rdBm} = P_{tdBm} + G_{tdB} - PL_{dB} + G_{rdB} - L_{qdB}. \quad (7.22)$$

The modern modems requires a minimum received power of -80 dBm to operate at Mbps data rates [38]. Thus, the maximum PL that can provide the communication link is found as 160 dB (PL threshold) by (7.22). Figure 7.10 presents the path-loss curves for different frequencies for 90% not exceeded of the time and PL threshold

for a 40 m surface duct ($\Delta M = 10$), 27 m transmitter and receiver. As noticed, 10 GHz can provide communication links up to 1000 km. Lower frequencies can provide higher ranges. However, since evaporation ducts also perform best near 10 GHz, the usage of this frequency range can increase the availability of the links.



Chapter 8

PATH-LOSS AND CORRELATION ANALYSIS FOR SURFACE DUCTS

8.1 Introduction

The trapping effect of atmospheric ducts provides significant received power enhancement compared to the standard atmospheric conditions. For this reason, atmospheric ducts can be utilized as a b-LoS communication medium, which is also denoted as *ducting channel* [22]. Although there are many studies on ducting channel wave propagation, there is no study, which considers antenna correlation and MIMO systems for ducting channels. Therefore, the main motivation of this chapter is to analyze MIMO ducting channels in the presence of surface ducts. In this chapter, we only focus on surface ducts because modified refractivity of surface ducts can be modeled with a simple bi-linear model as described in Section 8.2.

There are three main contributions of this work. Firstly, we develop a large-scale path-loss model for surface ducts to estimate the path-loss. By using the developed large-scale path-loss model, the distribution of fading in ducting channel is predicted. Second contribution is to analyze the correlation coefficient for space and polarization diversity systems in ducting channel for the first time in the literature. Since b-LoS communication systems generally operates at low SNR due to high path lengths, MIMO systems have high potential to increase the achievable data rates. Thirdly, we compare the performance of different diversity techniques in terms of achievable data rates by combining the proposed path-loss and correlation analyses. This way, we aim to show that ducting channel is suitable for high data rate applications.

The remainder of the chapter is organized as follows. Section 8.2 provides the

review of ducting channel modeling. Section 8.3 includes the proposed large-scale and small-scale path-loss models for ducting channel. In Section 8.4, we provide the correlation analysis for space and polarization diversity. Section 8.5 includes the capacity calculations with the proposed models. The achievable data rate results are presented in Section 8.6.

8.2 Channel Modeling in Ducting Channels

This section includes the propagation and atmospheric modeling for ducting channels.

8.2.1 Parabolic Equation Method

PE was originally developed in [83, 96], and utilizes par-axial approximation to the Helmholtz wave equation. PE methods became the dominant technique to model wave propagation in atmospheric ducts after the derivation of a computationally efficient solution based on SSF transform [83]. PE is especially advantageous to model complex boundary conditions and refractivity variations of the lower atmosphere [67]. Thus, PE methods are mostly utilized in ducting channel wave propagation analysis because wave propagation in ducting channel depends on multiple parameters: transmitter height, receiver height, duct height, duct strength, carrier frequency, polarization and surface conditions.

For 2D narrow angle forward scatter waves, PE is represented as [68]

$$\frac{\partial u(x, z)}{\partial x} = \frac{i}{2k} \frac{\partial^2 u(x, z)}{\partial z^2} + \frac{ik}{2} (m^2(x, z) - 1) u(x, z), \quad (8.1)$$

where x is the range in m, z is the height in m, $k = 2\pi/\lambda$ is the wave number, λ is the wavelength, and $u(x, z)$ is the reduced function. The modified refractivity index is $m = 1 + M10^{-6}$, where M is the modified refractivity. Horizontal and vertical polarizations can be modeled by (8.1) with appropriate boundary conditions.

In this chapter, PETOOL [70] is utilized to solve (8.1), and the path-loss (PL) is calculated as [67]

$$PL = 20 \log(4\pi/|u(x, z)|) + 10 \log(d) - 30 \log(\lambda), \quad (8.2)$$

where d is the range in m. In Section 8.3, PETOOL simulation results are utilized to develop large-scale path-loss model and analyze small-scale path-loss characteristics.

8.2.2 Atmospheric Modeling

In PE methods, atmospheric conditions are considered via modified refractivity, which is represented as

$$M = N + (h/R_0) \times 10^6 \text{ M-units}, \quad (8.3)$$

where h is the height in km and R_0 is the earth radius in km. Refractivity $N = (n - 1) \times 10^6$ N-units, where n is the atmospheric refractive index.

Trapping layer of atmospheric ducts is associated with negative modified refractivity gradient as in Figure 8.1. This condition causes propagating signals to be refracted back to the sea surface, and these signals are trapped between the sea surface and atmospheric duct. This way, trapped signals can propagate over-the-horizon with lower path-loss compared to the standard atmospheric conditions. This effect makes ducting channel a promising candidate for b-LoS communications especially in coastal and maritime environments [22].

Atmospheric ducts can be classified according to their formation process and properties: evaporation ducts, surface-based ducts, elevated ducts, and surface ducts [22]. The occurrence probability of surface and evaporation ducts are significantly higher than surface-based ducts and elevated ducts. Therefore, surface and evaporation ducts are more suitable for b-LoS communications. Surface ducts can be modeled with a bi-linear curve and depend on two parameters: duct height (δ) and duct strength (ΔM) as presented in Figure 8.1(a). Duct height (δ) represents the height of the trapping layer. Duct strength is the change in the modified refractivity from surface to trapping layer. On the other hand, evaporation ducts can be modeled with a logarithmic curve as in Figure 8.1(b). Ray tracing analysis for the bi-linear model is computationally efficient to estimate angle-of-arrivals and delay spreads of the ducting channels. For this reason, evaporation ducts are also approximated by a bi-linear

curve in the propagation analysis [106]. Thus, we only focus on surface ducts in the remainder of the chapter.

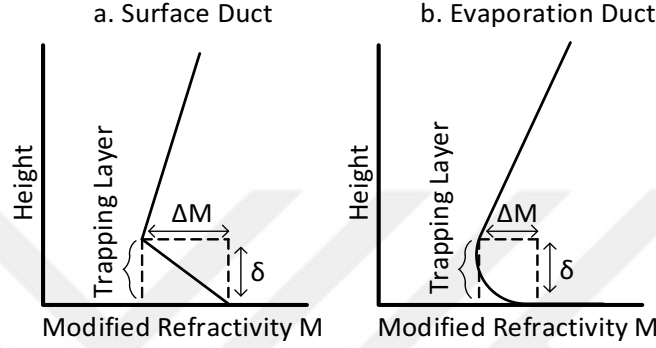


Figure 8.1: Modified refractivity profiles for (a) surface duct, and (b) evaporation duct.

Surface ducts are modeled with duct height and duct strength. Modified refractivity above the duct height is modeled as the standard atmospheric conditions (Modified refractivity increases with 118 M-units/km.). According to the experimental studies [79, 80], the surface duct formation has up to 80% probability of occurrence, and even in winter conditions with low evaporation rates this probability is above 60%. With the high formation probabilities, trapping effect of surface ducts can be utilized as a b-LoS communication medium.

8.3 Large-scale and Small-scale Path-Loss Model for Surface Ducts

This section provides the large-scale path-loss model for surface ducts and the distribution of the small-scale path-loss. To this end, we develop a large-scale path-loss model for the surface ducts based on regression analysis to PETOOL results as in [105]. Then, the deviations from our large-scale path-loss model are denoted as the small-scale path-loss, which is also known as *fading*.

Ducting channel communication is especially promising in places, where direct LoS communication is not available or expensive [21]. Therefore, ducting communication

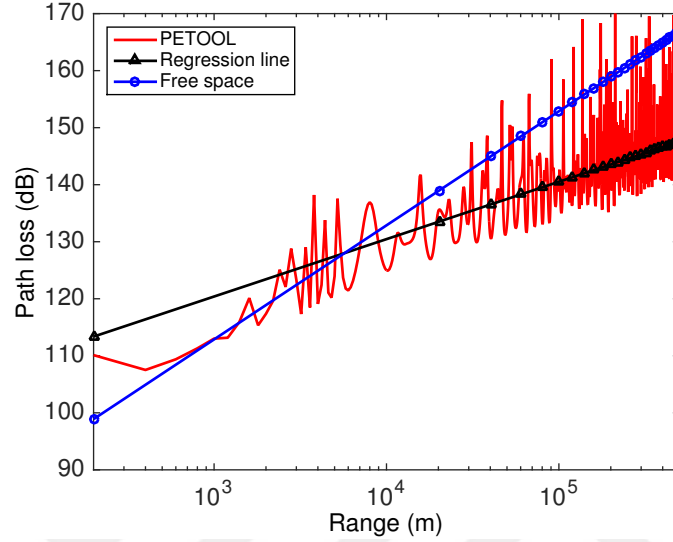


Figure 8.2: Path-loss results with PETOOL and regression model.

studies focus on b-LoS ranges. For a wireless channel, the generic large-scale path-loss equation is given as [27]

$$PL_{dB} = A + 10\gamma \log d/d_0 + X, \quad (8.4)$$

where X is the small-scale fading, γ is the path-loss exponent, d is the range in m, A is the path-loss in dB at d_0 , and d_0 , as defined in (5), is the fixed range in m beyond where b-LoS is observed [107].

$$d_0 = (4.12\sqrt{h_t} + 4.12\sqrt{h_r}) \times 10^3 \text{ m}, \quad (8.5)$$

where h_t and h_r are the transmitter and receiver heights in m, respectively.

For the large-scale path-loss model, we first generate path-loss results for ducting channel with PETOOL for the following parameters: frequency 10.5 GHz, transmitter height 27 m, receiver height 27 m, horizontal polarization, duct height 40 m, duct strength 20 M-units. Figure 8.2 includes the path-loss vs. range for fixed receiver height. Then, we apply regression analysis to fit the path-loss results presented in Figure 8.2 to (8.4) in the least square error sense as in [105, 108]. In this way, A and γ parameters in (8.4) are found as $A = 136$ dB and $\gamma = 1.16$. Figure 8.2 also includes

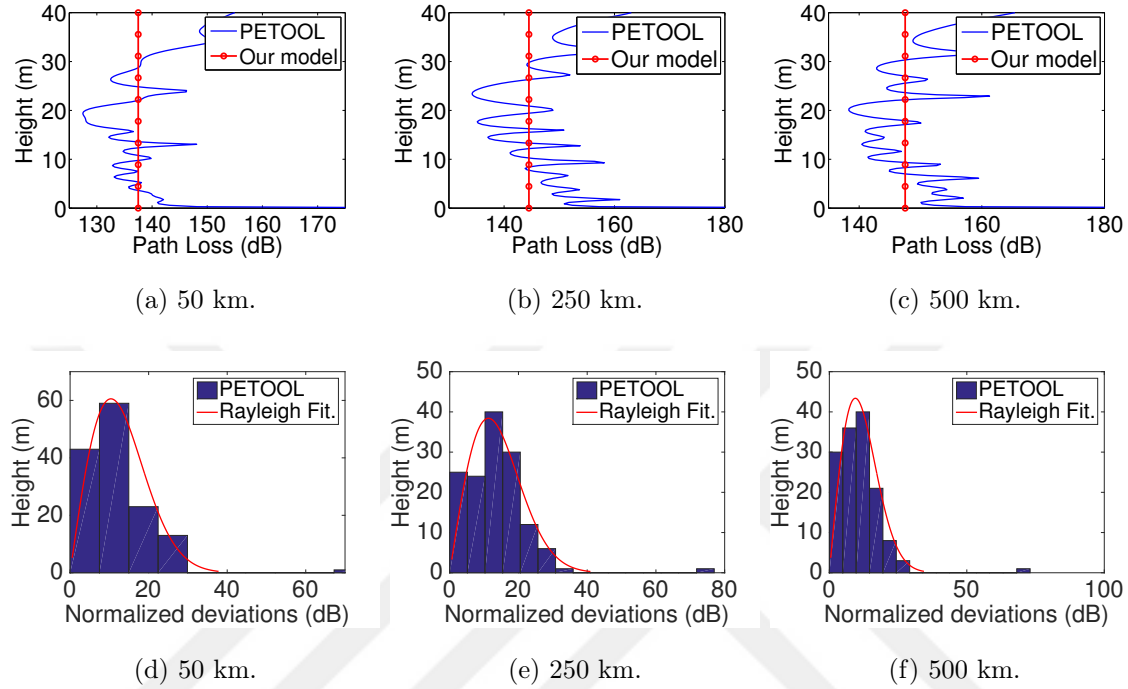


Figure 8.3: (a-c) The comparison of PETOOL and our model, (d-f) histograms of the deviations with Rayleigh distribution fitting.

the regression based model (straight line in Figure 8.2) and the free-space path-loss. As noticed, the surface duct can provide 20 – 30 dB lower path-loss compared to the free space conditions as also reported in [20].

Figure 8.3(a-c) present the path-loss estimates of PETOOL and the proposed regression based path-loss model for different ranges. As noticed, the level of the large-scale path-loss can be estimated with the developed model. The deviations from the regression model can be assumed as fading in the ducting channel. Figure 8.3(d-f) present the normalized distribution of the fading in the ducting channel for different ranges along with the Rayleigh distribution fitting. The Rayleigh distribution fitting in Figure 8.3(d-f) has standard deviation between 8 – 9 dB. To justify the Rayleigh distribution assumption, we utilize the two-sample Kolmogorov-Smirnov test [109] on MATLAB. According to the test, the presented histograms in Figure 8.3(d-f) can be

fitted with the Rayleigh distribution as low as 90% confidence level. As a result, the ducting channel with the surface ducts can be modeled as a Rayleigh fading channel.

8.4 Antenna Correlations for Space and Polarization Diversity

Ducting b-LoS communications suffer from high path-losses due to high communication ranges. Therefore, MIMO techniques have high potential to improve the achievable data rates. For this reason, we analyze the fading correlation for space and polarization diversity systems for ducting communication links.

8.4.1 Space Diversity

Space diversity can be achieved with horizontal or vertical placement. In this chapter, only vertical space diversity is considered because vertical space diversity becomes uncorrelated with a few meters separation. For calculating the space correlations, we generate path-loss results for vertically spaced transmitters, and calculate the correlation coefficient between the vertical variations in the path-loss at the desired range. Figure 8.4 shows the antenna correlation vs. vertical spacing for different carrier frequencies and different duct heights with transmitter height 27 m and duct strength rate 1 M-units/m (Duct strength is equal to duct strength rate times duct height.). As noticed, 1 m separation provides uncorrelated antennas (correlation coefficient close to 0) for 15 GHz. For 10 GHz, 1 – 2 m spacing yields low correlation values. The required spacing decreases as the carrier frequency increases. At the end, we can conclude that the vertical space diversity is promising for ducting channels.

8.4.2 Polarization Diversity

Polarization diversity is an advantageous method in wireless systems to provide higher data rates with lower cost [110,111] because polarization diversity can be implemented with dual-polarized antennas. This way, polarization diversity does not require an additional antenna. In order to utilize polarization diversity, the correlation between the polarizations should be lower than a certain threshold (≈ 0.5). To this end, we

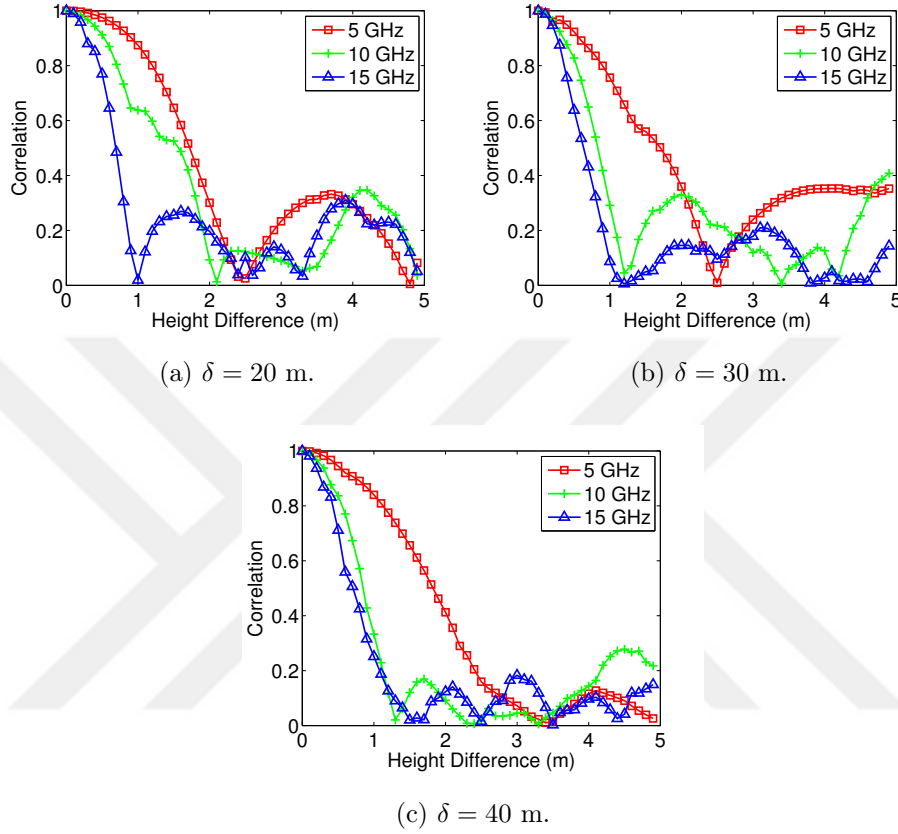


Figure 8.4: Antenna correlation for vertical diversity with different duct heights.

analyze the correlation between horizontal and vertical polarizations in b-LoS ducting links.

We also utilize PETOOL simulations to analyze the correlation in polarization diversity. For this case, we perform simulations for transmitters with horizontal and vertical polarizations. We use the following channel parameters: transmitter/receiver heights 27 m, horizontal polarization in the receiver, sea surface conditions, duct height 40 m and duct strength 10 M-units. Figure 8.5(a) presents the vertical path-loss results at 100 km range, where H and V represent horizontal and vertical polarizations, respectively. As noticed, the fading in different transmitter polarizations shows significant difference.

Fading correlation between polarizations (ρ_P) with respect to range is presented

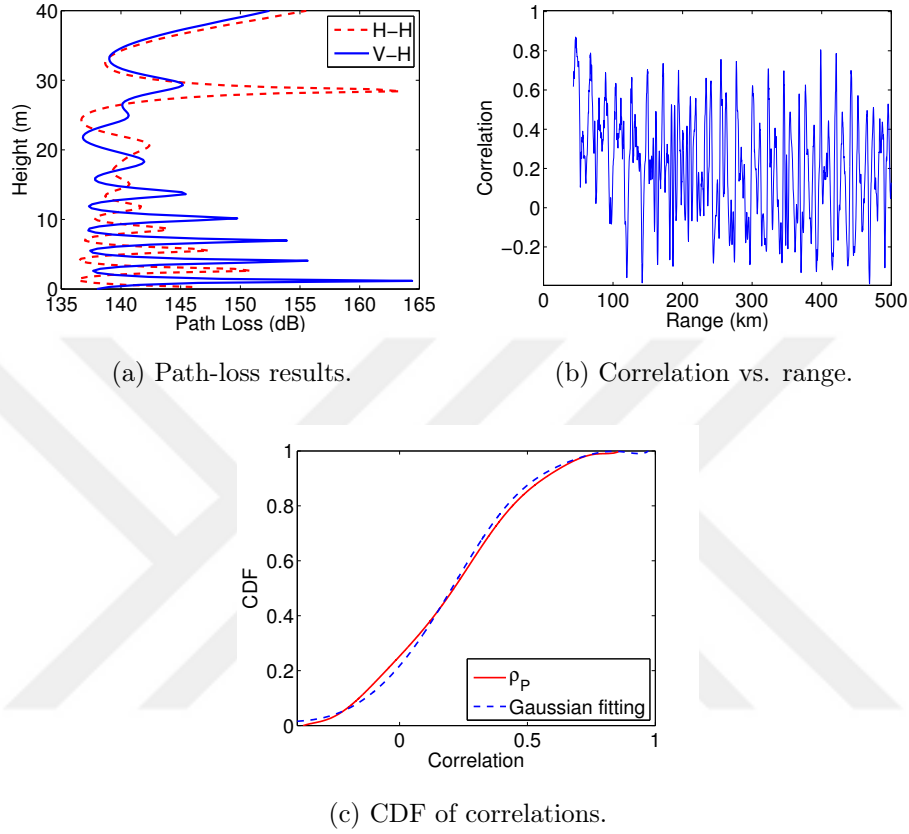


Figure 8.5: Fading correlation for polarization diversity.

in Figure 8.5(b). Figure 8.5(c) shows the CDF of the correlation between the polarizations. As noticed, 80% of the time the correlation is below 0.5. Although the standard deviation for the fading correlation is less between 50 – 100 km range, the overall fading correlation can be approximated by a Gaussian distribution with mean 0.2 and standard deviation 0.27 because ducting b-LoS communication is aimed to be utilized at b-LoS ranges, which are often higher than 100 km. As in Section III, we utilize the two-sample Kolmogorov-Smirnov test [109] to justify the fitted distribution in Figure 8.5(c), and this test yields that the fitted distribution has 95% confidence level. Since MIMO systems can achieve desired capacity gains below 0.5 correlation, polarization diversity is a promising method for ducting channel. The occurrence rate of surface ducts mostly varies between 70 – 80% depending on the season. Therefore,

the polarization diversity will be able to provide the expected gains up to 64% of the time. On the other hand, the correlation between horizontal and vertical polarizations at the receiver is close to one according to our simulations. At the end, polarization diversity is promising for ducting links as a transmitter diversity technique.

8.5 Channel Capacity in Surface Duct Links

In this section, we introduce capacity calculations to calculate achievable data rates for ducting channel by utilizing the proposed regression model and correlation results in Section 8.3 and 8.4, respectively. We utilize a narrow-band fading model for ducting channel with surface ducts because the previous ray-tracing analysis suggests that b-LoS ducting channel has delay spreads in the order of nanoseconds [100, 105, 112]. Therefore, multi-path components are not resolvable in bandwidth levels that we are interested (< 40 MHz). As described in Section 8.3, the ducting channel is assumed to have Rayleigh fading. As a result, we utilize flat *Rayleigh fading channel model*.

Suppose that N_T is the number of transmitters and N_R is the number of receivers, the channel can be modeled as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}, \quad (8.6)$$

where $\mathbf{H}_{N_R \times N_T}$ is the channel gain matrix, $\mathbf{x}_{N_T \times 1}$ and $\mathbf{y}_{N_R \times 1}$ are the transmit and receive vectors, $\mathbf{n}_{N_R \times 1}$ is the noise vector.

Since $\mathbf{H}$ contains the effect of both large-scale path-loss and small-scale path-loss, we utilize channel matrix normalization to separate these two effects. Thus, the normalized channel gain matrix $\tilde{\mathbf{H}}$ is represented as [41]

$$\tilde{\mathbf{H}} = \frac{\mathbf{H}}{g}, \quad (8.7)$$

where g is found as

$$g = \sqrt{\frac{1}{N_T N_R} \|\mathbf{H}\|_F^2}, \quad (8.8)$$

where $\|\cdot\|_F$ is the Frobenius norm.

With the channel matrix normalization, the average receive power in Watts is given by [41,42]

$$Pr_{av} = g^2 P_t = P_t G_t G_r (PL_L), \quad (8.9)$$

where P_t is the transmit power in W, G_t and G_r are the transmitter and receiver antenna gains, respectively. Lastly, PL_L is the large-scale path-loss, which will be directly calculated with the developed large-scale path-loss model $(PL_L)_{dB} = A + 10\gamma \log d/d_0$ dB as in (8.4), where A and γ are calculated as in Section 8.3. In this way, $g^2 P_t$ will be directly calculated by using the developed large-scale path-loss model by (8.9) for the given channel parameters: P_t , G_t , G_r . In this chapter, we assume that large parabolic reflector antennas are utilized to provide high antenna gains. However, high antenna gains can be provided via planar antenna arrays as proposed in [113].

The effects of small-scale path-loss are modeled with $\tilde{\mathbf{H}}$. The normalized channel gain matrix can be represented as [43]

$$\tilde{\mathbf{H}} = (\mathbf{R}^{RX})^{1/2} \mathbf{G} (\mathbf{R}^{TX})^{1/2}, \quad (8.10)$$

where $\mathbf{G}_{N_R \times N_T}$ is the complex Gaussian independent and identically distributed channel matrix. The root of the receiver antenna correlation matrix $((\mathbf{R}^{RX})^{1/2})$ can be computed with the Cholesky factorization. In the same way, $(\mathbf{R}^{TX})^{1/2}$ is the root of the transmitter antenna correlations $(\mathbf{R}^{TX})_{N_T \times N_T}$. The correlation matrices depend on the type of diversity. In this study, we assume that the channel is symmetrical for vertical space diversity for simplicity because the modeled surface duct is range independent and same number of antennas are utilized at both ends with the same orientations. Thus, the calculated correlation values are assumed to be valid for both transmitter and receiver in the vertical space diversity.

For space diversity: The transmitter correlation matrix for $N_T \times N_R$ space

diversity is shown as

$$\mathbf{R}^{TX} = \begin{bmatrix} 1 & \rho_{1,2}^S & \cdots & \rho_{1,N_T}^S \\ \rho_{2,1}^S & 1 & \ddots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ \rho_{N_T,1}^S & \cdots & \cdots & 1 \end{bmatrix}, \quad (8.11)$$

where $\rho_{i,j}^S$ is the correlation between antenna i and j and, they are determined by using the analysis in Figure 8.4 according to the channel parameters. The $N_R \times N_R$ receiver correlation matrix has also the same structure as (8.11).

For polarization diversity: The polarization diversity is only effective for transmitter side since polarizations have high correlation at the receiver side. Therefore, we assume that all receivers are horizontally polarized. For a dual-polarized antenna, the transmitter correlation matrix is given as

$$\mathbf{R}_P^{TX} = \begin{bmatrix} 1 & \rho_{1,2}^P \\ \rho_{2,1}^P & 1 \end{bmatrix}, \quad (8.12)$$

where $\rho_{1,2}^P = \rho_{2,1}^P$ is the correlation between horizontal and vertical polarizations. As shown in Figure 8.5, the correlation between polarizations can be modeled with a Gaussian random variable with mean 0.2 and standard deviation 0.27 as described in Section 8.4.2.

For space-polarization diversity: We also consider spatially separated dual-polarized transmitters. The transmitter correlation matrix for N_T number of dual-polarized transmitters is given as

$$(\mathbf{R}_{TX}^{SP})_{2N_T \times 2N_T} = \begin{bmatrix} [\mathbf{R}_P^{TX}]_{1,1} & \cdots & \rho_{1,N}^S [\mathbf{R}_P^{TX}]_{1,N_T} \\ \vdots & \ddots & \vdots \\ \rho_{N,1}^S [\mathbf{R}_P^{TX}]_{N_T,1} & \cdots & [\mathbf{R}_P^{TX}]_{N_T,N_T} \end{bmatrix} \quad (8.13)$$

where the correlation matrices for dual-polarized antennas $[(\mathbf{R}_P)_{2 \times 2}]_{\{1:N_T, 1:N_T\}}$ are determined randomly as in (8.12). Since dual-polarized antennas are utilized at the transmitter side, the total number of transmitters becomes $2N_T$ for N_T number of dual-polarized transmitters.

By using the matrix normalization in (8.9), the resulting channel capacity is calculated as

$$\begin{aligned} C &= \log_2 \det \left(\mathbf{I}_{N_R} + \frac{Pr_{av}}{N_T \sigma_n^2} \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H \right), \\ &= \log_2 \det \left(\mathbf{I}_{N_R} + \frac{g^2 P_t}{N_T \sigma_n^2} \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H \right), \end{aligned} \quad (8.14)$$

where σ_n^2 is the noise in the system, N_T is the total number of transmitters, $\mathbf{I}_{N_R}$ is $N_R \times N_R$ identity matrix, and g is utilized in the channel matrix normalization in (8.8). Since there is a power amplifier at the receiver, the amplified noise at the receiver can be found as

$$\sigma_n^2 = k(T_a + T_e)BN_F G_a, \quad (8.15)$$

where k is the Boltzman's constant, G_a is the amplifier gain, T_a is antenna noise temperature, and T_e is effective noise temperature, which is give as $N_F - 1 = T_e/T_a$ and N_F is the amplifier noise figure.

In the channel capacity calculations, we utilize uniform power allocation to the all eigenvalues by assuming the power level of the channel is high enough. By denoting $\gamma = \frac{g^2 P_t}{N_T \sigma_n^2}$, the capacity is simplified as

$$C = \sum_{i=1}^{N_{min}} \log_2 (1 + \gamma \lambda_i), \quad (8.16)$$

where $N_{min} = \min(N_T, N_R)$, λ_i 's are the nonzero eigenvalues of the Wishart matrix ($\tilde{\mathbf{H}} \tilde{\mathbf{H}}^H$ for $N_R \leq N_T$, or $\tilde{\mathbf{H}}^H \tilde{\mathbf{H}}$ for $N_R > N_T$).

8.6 Performance Evaluations

In this section, we provide the data rate results for the diversity techniques. We utilize PETOOL to determine the parameters for our large-scale path-loss model as in Section 8.3. Channel parameters that are utilized in the simulations are listed in Table 8.1. We perform 10000 iterations on MATLAB for the random channel gain matrix (8.10) and random polarization correlation coefficient (8.12).

Table 8.1: Ducting channel communication parameters.

Parameter	Value	Unit
<i>Range (d)</i>	300	km
<i>Carrier frequency (f)</i>	10.5	GHz
<i>Transmit power (P_t)</i>	1	W
<i>TX&RX height</i>	27	m
<i>Bandwidth (B)</i>	20	MHz
<i>Antenna gains ($G_{t,r}$)</i>	30	dBi
<i>Amplifier gain (G_a)</i>	50	dB
<i>Amplifier noise figure (N_F)</i>	1.32	dB
<i>Vertical spacing (d_v)</i>	1	m
<i>Spatial correlation (ρ^S)</i>	0.5	-

Figure 8.6 shows the capacity results for single-input-single-output (SISO), space diversity, polarization diversity and space-polarization diversity in ducting channels with 40 m surface duct and 10 M-units duct strength. As noticed, the polarization diversity significantly improves the data rates of the link as in Figure 8.6. Since dual-polarized transmitters can be mounted on a single antenna feeder, the cost of polarization diversity is significantly lower compared to the space diversity. However, the space diversity can provide higher diversity gains with the additional antennas at both sides. By utilizing 2 dual-polarized transmitters and 2 horizontally polarized receivers, the achievable data rates can be significantly improved. Therefore, polarization diversity and vertical space diversity are promising methods for ducting channel communication systems to provide higher data rate levels.

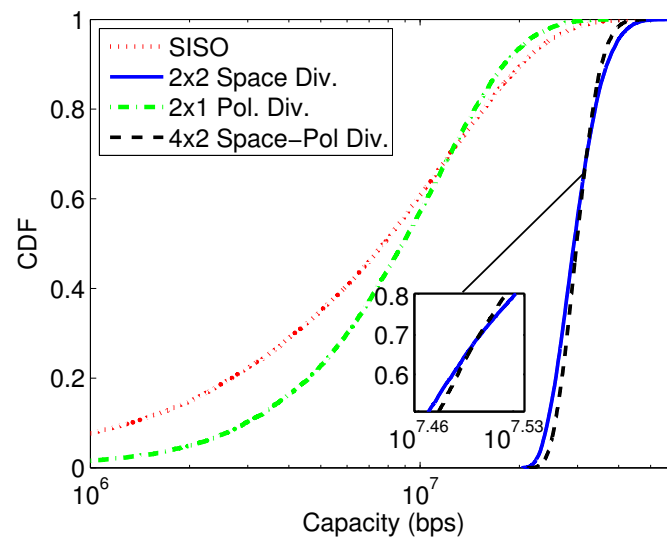


Figure 8.6: Ducting channel capacity with diversity techniques.

Chapter 9

B-LOS DUCTING CHANNELS: COHERENCE TIME AND RAIN ATTENUATION

9.1 Introduction

In ducting communications, both of the antennas should be in the duct layer in order to utilize the trapping effect. In this way, the trapped signals can propagate through b-LoS ranges because ducting layer avoids the spreading of signal energy to the atmosphere. Instead, signal energy is concentrated between surface and duct layer. Ducting channel communication can be utilized as a communication medium up to 400 km [22, 105]. Trapping effect of atmospheric ducts are effective between 5 – 10.5 GHz depending on duct properties. Since atmospheric ducts utilize such high frequency range and signal rays follow high distances close to the surface level, ducting channel may be exposed to rain attenuation. However, there is no study which considers rain attenuation in b-LoS ducting links.

There are plenty of rain attenuation models in the literature [16–18]. Most of the available studies consider the slant-path rain attenuation models for SATCOM [17, 18]. The closest work for this chapter is [23] in which a slant-path rain attenuation model is proposed for troposcatter considering nonuniform rainfall distribution. We utilize similar rain data generation as in [23], but the path characteristics are completely different in ducting channels. Thus, the model in [23] is not applicable to ducting channels and there is no study which considers rain attenuation in ducting links.

The main contributions of this chapter are two-fold. Firstly, coherence bandwidth and time of ducting channels are analyzed. In [105], the authors analyze delay spread of the channel with frequency independent ray-optics methods. In this chapter, coher-

ence bandwidth is determined via frequency dependent frequency correlation results. In addition, coherence time of the channel is estimated by utilizing angle constraints developed in [105]. Coherence bandwidth and time analysis can be utilized to determine the bandwidth and mobility levels in which the b-LoS channel has flat and slow fading. Secondly, a slant-path rain attenuation model for ducting channel is developed based on the ray analysis in [105] and synthetic rain distribution model in [55, 56]. Simulation results are presented for the rain loss and spectral efficiency of the system under rain. These results can guide the b-LoS system designer to determine the carrier frequency which provides better spectral efficiencies with or without rain to design more reliable links.

9.2 Atmospheric Ducts

Atmospheric duct is the layer in which modified refractivity gradient is turned negative to positive ($0 \geq \partial M / \partial z$) as in Figure 9.1(a). This layer traps the propagating rays with low grazing angles, and the trapped signal energy is concentrated between the surface and duct layer. In this way, propagating signals can reach b-LoS distances [20, 22]. Since duct formation is more probable in humid areas, ducting channel is a promising b-LoS communication medium especially in naval communication applications [22]. In maritime and coastal areas, the occurrence probability of atmospheric ducts varies between 60 – 80% according to the season [80].

Atmospheric ducts can be divided into four categories according to their formation process: evaporation ducts, surface-based ducts, elevated ducts, and surface ducts. In [22], the authors analyze path-loss of a 13 m evaporation duct and show that 10.5 GHz frequency has the lowest path-loss. In [105], the authors develop a large-scale path-loss model for a 40 m surface duct. According to their results, lower frequency ranges (3–5 GHz) has better trapping effects. The difference between frequency range is caused by duct height difference. As experimentally presented in [121], evaporation ducts with high duct heights also perform better around 3 GHz. Since modified refractivity profiles of evaporation duct and surface duct are similar as shown in Figure

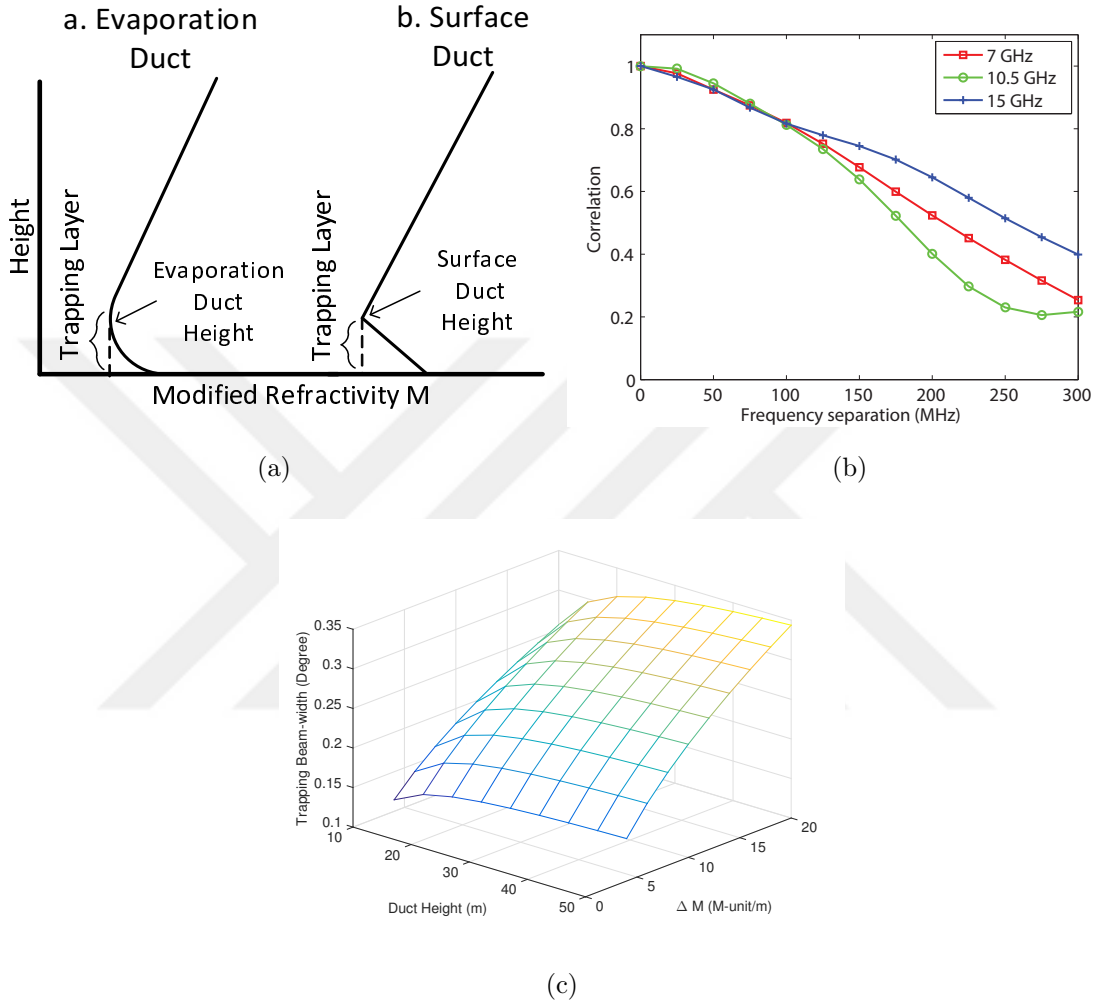


Figure 9.1: (a) Modified refractivity profile of evaporation duct and surface duct, (b) coherence bandwidth for ducting links, (c) trapping beam-width values for 5 m antenna height.

9.1(a), we can approximate evaporation ducts with a surface duct with the same duct height and duct strength. Duct strength is the change in modified refractivity from surface to trapping layer. For this reason, the analysis presented within this chapter is also valid for the evaporation duct with the same duct parameters.

9.3 Coherence Bandwidth and Time

In this section, the analysis for coherence bandwidth and coherence time of ducting links are presented.

9.3.1 Coherence Bandwidth

In [105], maximum delay spread of the channel is estimated via ray-optics methods which are independent of frequency. To this end, the coherence bandwidth of the ducting channel is analyzed with PE methods to observe the effects of frequency. For this purpose, we use PETOOL [70] which is a split-step-Fourier based PE wave propagation tool to estimate the path-loss in ducting links. PETOOL is calibrated with experimentally validated tools [70].

PETOOL results are utilized to estimate the coherence bandwidth in ducting channels. To this end, we generate two path-loss profiles that have some frequency separation and calculate the correlation between vertical cuts at the desired range to find the frequency correlation. Figure 9.1(b) shows the correlation vs. frequency separation for the surface duct with the following simulation parameters: 150 km range, 20 m transmitter and receiver height, duct height 40 m, duct strength 10 M-units and horizontal polarization. The channel can be assumed as flat when the frequency correlation is lower than 0.5. As in Figure 9.1(b), the ducting channel can be assumed as a flat-fading channel if the channel bandwidth is lower than 205 MHz for 7 GHz, 180 MHz for 10.5 GHz and 252 MHz for 15 GHz. According to the ray optics analysis presented in [105], ducting channels have coherence bandwidth levels more than 200 MHz. Therefore, our results are consistent with [105]. In addition, the frequency correlation do not show a notable pattern with the frequency because trapping effect also do not show direct relationship with frequency as in [22].

9.3.2 Coherence Time

For coherence time estimation, a simple Doppler spread approach is used based on the limited AoA range. Since ducting channel traps the signal at the ducting layer, only

the rays that have particular angle range can be received at the receiver. In [105], an analytical formula for the trapping angle range is derived. Upper and lower bound for the trapping angle in the elevation cut can be estimated with

$$\theta_{max,min}^T = \pm \sqrt{2 \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) (h_t - \delta)}, \quad (9.1)$$

where dn/dz is the refractive index derivative with respect to height, h_t is the transmitter height, δ is the ducts height, R_0 is the earth radius (≈ 6370 km) and $n(0)$ is the index of refraction at surface (≈ 1.00035) [22]. Atmospheric ducts are generally denoted with duct strength (ΔM) which is the change in the modified refractivity from bottom to top of the duct ($dn/dz = -10^{-6}(\Delta M/\delta - 0.157)$) [22].

Figure 9.1(c) shows the trapping beam-width values (θ_{max}) for 5 m transmitter height. Based on experimental observations [79,80], duct heights and strengths mostly vary between 10 – 40 m and 4 – 10 M-units, respectively. As in Figure 9.1(c), these duct height and strength ranges are associated with $0.2^\circ - 0.3^\circ$ beam-width in the elevation cut. Thus, ducting channel has very limited receive angle span.

In ducting channel, the elevation cut has very narrow beam-widths ($0.2^\circ - 0.3^\circ$). Therefore, Doppler spread due to mobility is very limited in ducting channels and it is given as

$$D_s = v \cos(\theta)/\lambda, \quad (9.2)$$

where v is the speed, λ is the wavelength and θ is the angle between the received ray and horizon. We assume that the receiver is moving away from the transmitter on the same line. For 10.5 GHz and $v = 10$ m/s, the Doppler spread becomes 349.9979 Hz for the 0.2° ray and 349.9952 Hz for the 0.3° ray. Therefore, the Doppler spread can be assumed as ≈ 350 Hz for $v = 10$ m/s and the estimated coherence time is ≈ 2.9 ms. Since the speed of naval vehicles will be not so high, coherence time of ducting links will be in the order of milliseconds. Thus, ducting channel will experience slow fading for MHz level bandwidths.

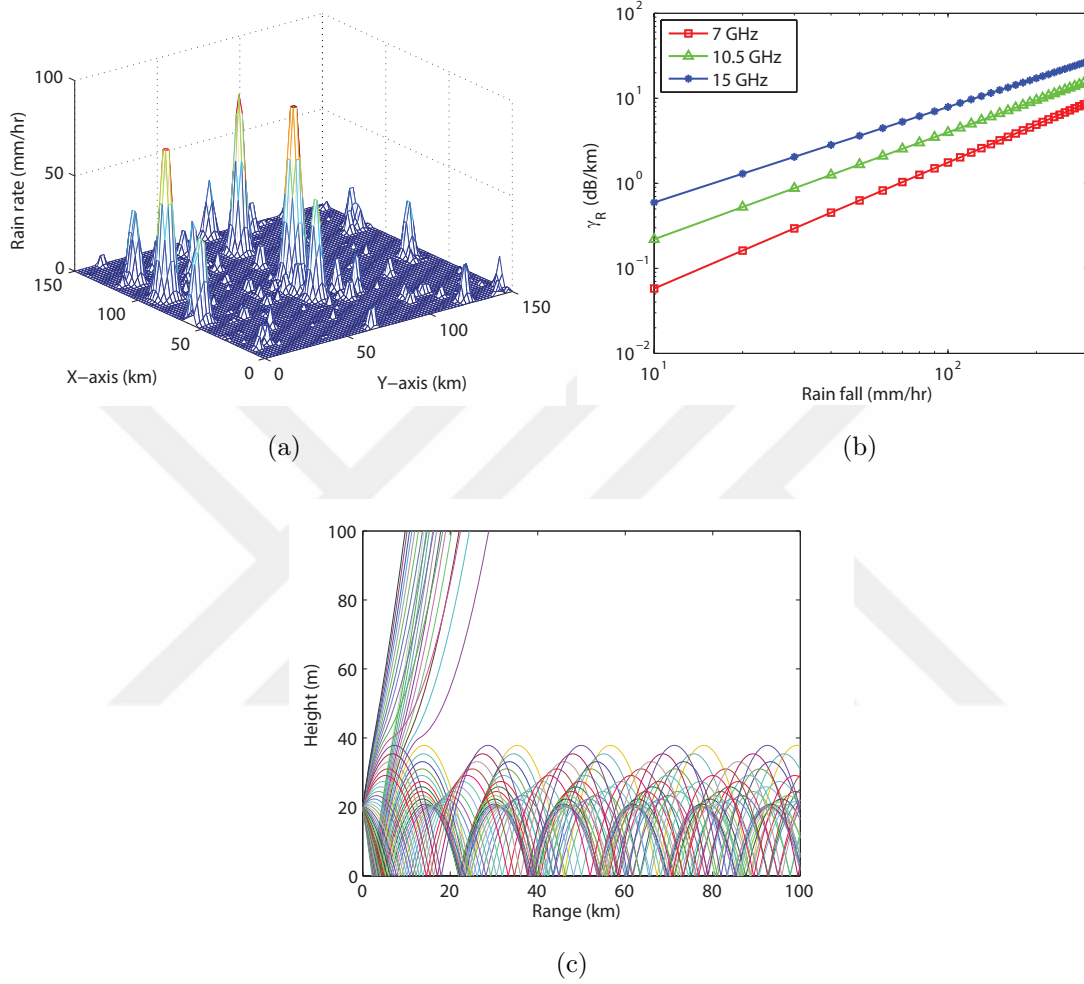


Figure 9.2: (a) Spatial distribution of rain cells, (b) rain specific attenuation, (c) ducting channel ray trajectories.

9.4 Rain Attenuation Model

Ducting links will utilize high frequency ranges as analyzed in [22]. Thus, b-LoS paths may experience strong rain attenuation. Path-loss and rain loss in the ducting links are mutually exclusive as outlined in [122]. Therefore, the total transmission loss can be determined by calculating path-loss and rain loss separately. To this end, a rain attenuation model for ducting links is developed in this section.

9.4.1 Rain Model

A synthetic rain model is utilized to generate rain cell spatial distribution as in [23]. To this end, HYCELL model [55, 56] is adapted with a slight modification. In the HYCELL model, rain distribution in a single rain cell is modeled with both Gaussian and exponential functions. Since the exponential function models only tail of the rain cell, rainfall rates associated with exponential function is very low. Thus, only Gaussian function is utilized to model the rain cells in this chapter, and rainfall distribution of a single cell is given as

$$R(x, y) = \begin{cases} R_G \exp[-(x^2/a_G^2 + y^2/b_G^2)], & \text{if } R > R_1 \\ 0, & \text{otherwise,} \end{cases} \quad (9.3)$$

where R_G is the peak rainfall rate, a_G and b_G are the distance where the rainfall rate decreases by a factor of $1/e$ in x and y axis. Rain cells are assumed as symmetrical, such that $a_G = b_G$. The minimum rain rate (R_1) is utilized as 1 mm/hr [55]. HYCELL model [55, 56] estimates spatial rain distribution based on local spatial rain probabilities which are provided by ITU-R P. 837-6 [58]. By utilizing the modified HYCELL model and local rain probabilities for Istanbul, Turkey [58], the spatial rain distribution is estimated as in Figure 9.2(a).

9.4.2 Slant-path Model

In this subsection, a slant-path rain attenuation model is developed for ducting channels by utilizing the rain model discussed in the previous part. In the slant-path model, ITU-R P.838 [59] rain specific attenuation is utilized and it represents the amount of path-loss in dB/km due to rain as

$$\gamma_R = kR^\gamma \text{ dB/km}, \quad (9.4)$$

where k and γ are empirical coefficients and their values are utilized as provided in [59]. Figure 9.2(b) presents rain specific attenuation values for different frequencies. In the slant-path model, total rain attenuation is calculated by multiplying the rain specific

attenuation with the rain path length. Thus, rain path length in ducting links is required to find the rain loss.

For rain path lengths, the ray trajectory analysis are utilized as introduced in [105]. Figure 9.2(c) shows the ray trajectories for the ducting channel with 20 m transmitter height, 40 m duct height and 10 M-units duct strength. The trapped rays follow longer paths compared to the range which is assumed as the horizontal distance between transmitter and receiver. However, the difference between ray path lengths and range is very low as presented in Figure 9.2(c). In the simulations, the average ratio of ray path length to the range is $\epsilon = 1.00000635548381$. Thus, the parabolic shape of the rays can be neglected and the rain attenuation calculations may utilize direct rays between transmitter and receiver. To this end, we multiply the calculated path lengths with ϵ .

In the proposed rain attenuation model, the rain attenuation for i th rain cell can be calculated as

$$\begin{aligned} A_{rain}^{(i)} &= \int_{d_s}^{d_e} kR(x)^\gamma \epsilon dx \\ &= \int_{d_s}^{d_e} R_G^\gamma \exp(-(x - d_m)^2/a_G^2)^\gamma \epsilon dx \text{ dB}, \end{aligned} \quad (9.5)$$

where d_s is the starting point of the rain cell, d_e is the end point of the rain cell and $d_m = (d_s + d_e)/2$ is the middle of the rain cell. By considering all of the rain cells in the communication path, the total amount of rain attenuation can be found as

$$A_{rain} = \sum_{i=1}^{N_R} A_{rain}^{(i)} \text{ dB}, \quad (9.6)$$

where N_R is the number of rain cells on the path.

9.5 Spectral Efficiency

The spectral efficiency results are also provided for ducting links in the presence of rain in Section 9.6. To this end, this section includes the spectral efficiency calculations by utilizing the path-loss model developed in [105]. In single-input single-output links,

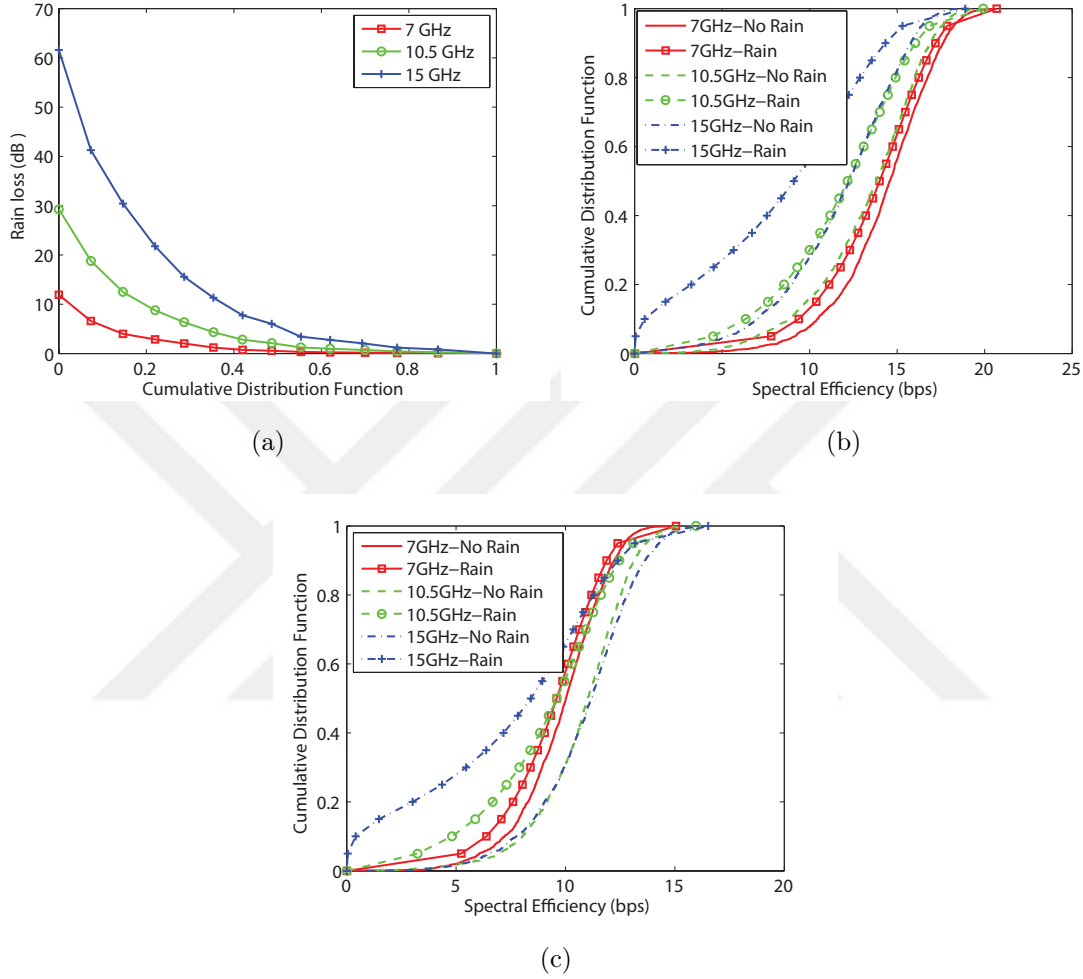


Figure 9.3: (a) Rain loss in ducting links, (b,c) Spectral efficiency in ducting links for scenario 1 and 2.

the spectral efficiency can be found as

$$R(t) = \log_2 (1 + \rho \times P_t / N) \text{ bps}, \quad (9.7)$$

where P_t is the transmit power, ρ is the path-loss which is calculated with the path-loss model in [105] and rain attenuation is also added to this term. N is the noise power which is modeled as thermal noise $N = kBT$ where T is the temperature in K, B is the bandwidth and k is the Boltzmann constant.

9.6 Simulation Results

Rain attenuation for each ray is calculated with the slant-path model for the rainfall distribution in Figure 9.2(a). The rain loss is calculated for each 500 m cut in y-axis of the synthetic rainfall distribution in Figure 9.2(a). In this way, CDF of rain loss is calculated with the following link parameters: 40 m duct height, 10 M-units duct height, 150 km range, 20 m antenna heights, 1.2 m antenna diameter (D_{ant}) and 0.65 antenna efficiency (η_{ant}).

Table 9.1: Simulation Parameters.

	Scenario 1		Scenario 2	
Frequency	T. Power	Bandwidth	T. Power	Bandwidth
7 GHz	30 dBm	20 MHz	45 dBm	13 MHz
10.5 GHz			30 dBm	20 MHz
15 GHz			21 dBm	28 MHz

Figure 9.3(a) shows the distribution of rain loss in a 150 km ducting link by utilizing the proposed slant-path rain model. As noticed, ducting channel can experience very high rain attenuations especially for frequencies higher than 10 GHz. By considering the high path-loss enhancements near 10.5 GHz [22], ducting channel links is prone to rain loss.

In addition, we present the spectral efficiency distributions for two scenarios in Figure 9.3(b) and 9.3(c). In the first scenario, the same channel parameters is utilized for each frequency as in Table 9.1 with 40 dB antenna gains. As noticed in Figure 9.3(b), rain attenuation may cause significant reduction in the spectral efficiency in high frequencies. In the second scenario, channel parameters are also changed in accordance with the carrier frequency as in Table 9.1. In addition, the antenna gains are calculated with the generic equation for the parabolic reflector antennas $G = 10 \log[\eta_{ant}(\pi D_{ant}\lambda^2)]$. Since higher antenna gains can be provided with higher carrier

frequencies, the spectral efficiency performance of 10.5 GHz and 15 GHz becomes better without rain as presented in Figure 9.3(c). By considering the significant spectral efficiency loss in 15 GHz due to rain, 10.5 GHz is expected to provide higher data rates.



Chapter 10

MULTI-STAGE BEAM-FORMING FOR B-LOS DUCTING CHANNELS

10.1 Introduction

Ducting channel b-LoS communication is a promising candidate especially in coastal and maritime regions because duct formation is highly probable in humid environments. Based on our previous analysis [105] and reviews [22], ducting channel can provide high data rate links with tens of Mbps data rate up to 500 km with transmission delays in the order of a few milliseconds.

We previously propose a path-loss model for surface ducts to estimate path-loss and fading characteristics of ducting channels [105]. In addition, we model AoA and delay spread of the channel through ray-tracing methods. In this study, we utilize the path-loss model in [105] for simulations.

Ducting channel communication has high path-losses because it is expected to operate at high ranges. Since trapped signals are concentrated in ducting layer, their AoA spread is very narrow. Therefore, antennas with sharp beam-widths can be used to provide high antenna gains. To this end, large planar antenna arrays can be utilized to provide sharp electronically steerable beams. The utilization of large antenna arrays, also known as massive MIMO, attracts significant research attention especially for LTE [114, 115] and millimeter waves [116, 117] attracts significant research attention. [116] provides Kronecker product model for planar antenna array in which azimuth and elevation beam-forming can be separated. [118] introduces virtual channel model for antenna arrays for virtual angle spreads. Since we have the knowledge of angular spreads in ducting channel through [105], we can use virtual channel

model in [118]. In addition, [117] provides multi-resolution codebook for hierarchical beam-search. Thus, planar arrays are promising for ducting channel to detect the trapped signal by adjusting the signal beam in both elevation and azimuth direction. Although sharp angular spread of ducting channel can be exploited via large antenna arrays, there is no work which considers beam-forming for b-LoS ducting channels.

The main contributions of this chapter are two-fold. Firstly, we propose a multi-stage limited feedback beam-forming method for ducting channel with planar antenna arrays. We decompose the beam-forming problem to azimuth and elevation beam-forming, and beam-forming vectors for azimuth and elevation are determined in two stage by using multi-resolution codebooks. In addition, we provide simulation results for the spectral efficiency of ducting b-LoS links for surface ducts by using the channel model developed in [105]. Although the proposed method is only tested for surface ducts, trapping characteristics of atmospheric ducts are similar, thus the proposed method can be used in all type of ducts.

The remainder of the chapter is organized as follows. Section 10.2 includes the properties of ducting channel and effective trapping beam-width. Section 10.3 provides the system model for ducting channel with planar antenna arrays. Multi-stage beam-forming method is proposed in Section 10.4. In Section 10.5, simulation results are presented.

Notation: We use the following notation throughout this chapter: a is a scalar; $\mathbf{a}$ is a column vector; $\mathbf{A}$ is a matrix and $\mathbf{a}_k$ is the k^{th} column of $\mathbf{A}$; $\|\mathbf{A}\|_F$ is the Frobenius norm of $\mathbf{A}$; $|\mathbf{A}|$ is the determinant of $\mathbf{A}$; $\mathbf{A}^T$ is the transpose of $\mathbf{A}$; $\mathbf{A}^H$ is the conjugate transpose of $\mathbf{A}$; $\otimes$ is the Kronecker product.

10.2 Ducting Channel

In this section, we provide a brief introduction to ducting channel b-LoS communication and modeling. In addition, we explain the requirement for planar antenna arrays to provide high data rate naval b-LoS links.

10.2.1 Ducting Channel Modeling

Ducting channel wave propagation are strongly affected by complex refractivity conditions. Therefore, PE techniques are utilized to model ducting channels with their capability of modeling complex boundary conditions and refractivity of the lower atmosphere [22, 67, 68]. In addition to PE methods, ray-optics methods are also utilized in hybrid methods [69] to estimate the AoA and delay spreads [72].

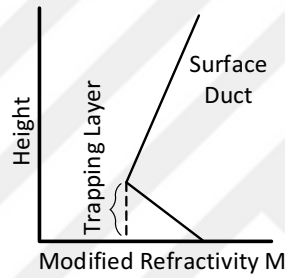


Figure 10.1: Modified refractivity profile of a surface duct.

We previously introduced a channel model for surface ducts to estimate path-loss, AoA, AoD and delay spreads [105]. Atmospheric ducts are divided into four types according to their physical formation process: evaporation duct, surface-based ducts, elevation ducts and surface ducts. Surface ducts can be modeled with a bi-linear curve as in Figure 10.2. We utilize PETOOL [70], which is a PE based wave propagation software, to develop the path-loss model.

In addition, we develop a ray-tracing method to estimate the ray trajectories as in Figure 10.2 to estimate delay spreads, AoAs, and AoDs. Based on our ray-tracing simulations in [105], ducting channel can be modeled as a narrow-band fading channel because maximum delay spread of the channel is considerably low compared to bandwidth values of interest 10 – 20 MHz. 40 surface duct with 10 M-units duct strength has 200 MHz one over maximum delay spread value at 400 km. In ducting channels, coherence bandwidth decreases with increasing range and duct height because path length difference increases with these factors. Since surface ducts are mostly between 5 – 30 m heights [79, 80], ducting channel experiences narrow-band fading.

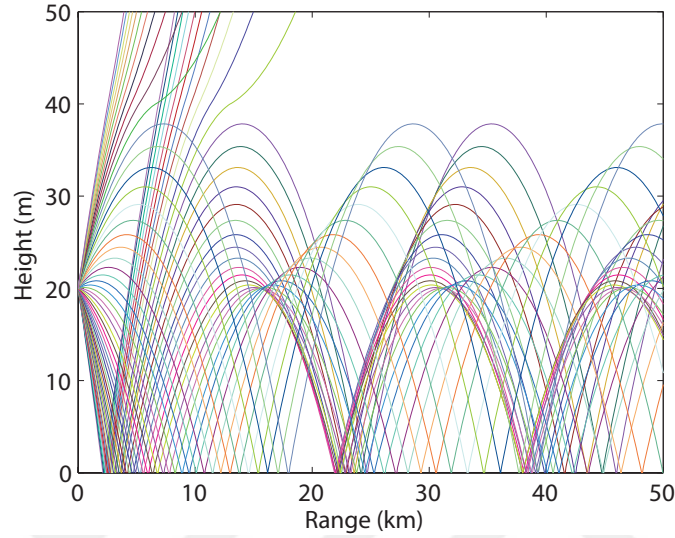


Figure 10.2: Ray trajectories for 40 m surface duct.

10.2.2 Effective Trapping Beam-width

Trapping effect of atmospheric ducts is only effective for the rays having certain angle range depending on channel condition: transmitter height, antenna orientation; atmospheric conditions: duct height and duct strength. Rays outside of this particular range cannot be trapped and propagate through the atmosphere. Since we can only exploit the path-loss enhancements of atmospheric ducts in this angle range, antenna beam-widths and orientation should be well adjusted.

In [105], we utilize the proposed ray trajectory formula to derive the span of effective trapping beam-width which can be represented as

$$\theta_{max,min}^T = \pm \sqrt{2 \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) (h_t - \delta)}, \quad (10.1)$$

where refractive index derivative with respect to height dn/dz , h_t is the transmitter height, δ is the ducts height, R_0 is the earth radius (≈ 6370 km) and $n(0)$ is the index of refraction at surface (≈ 1.00035) [22]. In this derivation, we assume that transmitter is in duct layer ($h_t < \delta$) because this is required condition to provide trapping effect. Since dn/dz is negative, we have real roots in (10.1). Instead of dn/dz ,

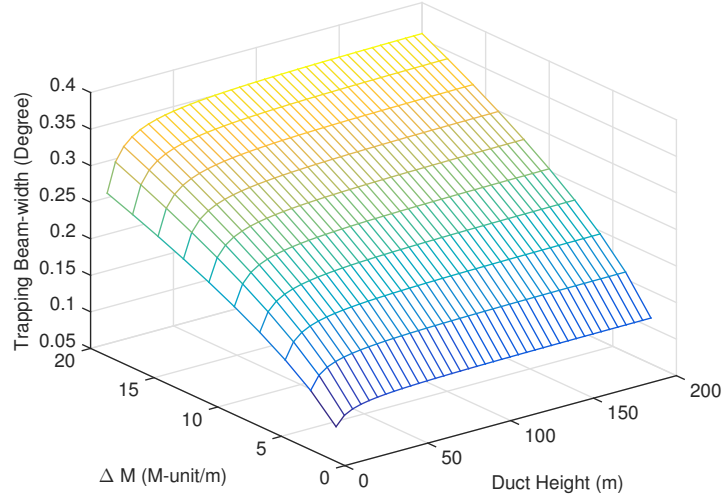


Figure 10.3: Trapping beam-width values for 5 m antenna height.

atmospheric ducts are characterized with duct strength (ΔM) which is the change in the modified refractivity from bottom to top of the duct ($dn/dz = -10^{-6}(\Delta M/\delta - 0.157)$) [22].

Figure 10.3 shows effective trapping beam-width values (θ_{max}) for 5 m transmitter height for different duct heights and duct strengths. Based on experimental observations, duct heights of 10 – 20 m and duct strengths of 4 – 10 M-units are highly probable [79, 80]. As in Figure 10.3, duct height of 10 – 20 m and duct strength of 4 – 10 M-units should have 3 dB elevation beam-width (Ψ_{el}) of $\approx 0.15^\circ - 0.25^\circ$. Therefore, transmitter elevation beam-width can be adjusted to this effective value to increase the utilization of the trapping effect. Receiver can focus its beam to this angle range to receive trapped rays and increase antenna gain. Therefore, we propose utilization of large planar antenna arrays to control both azimuth and elevation beams of the antenna. Since ducting channel requires very narrow elevation beams, high number of vertical row can be used to get sharper beams with high array gains.

10.3 System Model

In this section, we introduce system model for ducting channel with antenna arrays. As in Figure 10.4, both transmitter and receiver sides have planar arrays with $W_T \times H_T = N_T$ antenna elements where W_T is the number of horizontal elements, H_T is the number of vertical elements in transmitter. Similarly, receiver also has $W_R \times H_R = N_R$ antenna array.

As discussed in Section 10.2.1, ducting channel communication has narrow-band fading [105]. Since propagating signals have low spreading due to trapping effect, AoA in both elevation and azimuth is significantly narrow. Therefore, ducting channel can be modeled with the virtual channel representation as introduced in [118]. With the virtual channel representation, channel gain matrix is represented as

$$\mathbf{H} = \sqrt{\frac{N_T N_R}{L}} \sum_{i=1}^L \alpha_i \mathbf{a}_r^l(\theta_r, \phi_r) \mathbf{a}_t^l(\theta_t, \phi_t)^H \quad (10.2)$$

where N_T is the number of transmitter, N_R is the number of receiver, L is the number of paths, α_i is the complex gain of i 'th path. Since ducting channel has Rayleigh fading [105], amplitude of α_i is Rayleigh distributed. $\mathbf{a}_{t,r}$ are the array response vectors for transmitter/receiver arrays where θ is the azimuth angle and ϕ is the elevation of the signals. General expression for planar array response vector can be written as

$$\begin{aligned} \mathbf{a}(\theta, \phi) = & \frac{1}{\sqrt{N}} [1, \dots, e^{jkd(m \cos(\phi) \sin(\theta) + n \sin(\phi))}, \dots, \\ & \dots, e^{jkd((W-1) \cos(\phi) \sin(\theta) + (H-1) \sin(\phi))}]^T, \end{aligned} \quad (10.3)$$

where N is the total number of elements in the planar array, W is the horizontal and H is the vertical element number ($W \times H = N$), k is the wavenumber ($k = 2\pi/\lambda$). d is the separation between array elements which is used as $\lambda/2$ in this chapter. (10.3) can be decomposed as

$$\mathbf{a}^{az}(\theta, \phi) = [1, \dots, e^{jkd(m \cos(\phi) \sin(\theta))}, \dots, e^{jkd(W-1) \cos(\phi) \sin(\theta)}]^T, \quad (10.4)$$

$$\mathbf{a}^{el}(\phi) = [1, \dots, e^{jkd(n \sin(\phi))}, \dots, e^{jkd(H-1) \sin(\phi)}]^T \quad (10.5)$$

$$\mathbf{a}(\theta, \phi) = \frac{1}{\sqrt{N}} (\mathbf{a}^{az}(\theta, \phi) \otimes \mathbf{a}^{el}(\phi)) \quad (10.6)$$

As noticed, $\mathbf{a}_{el}$ only depends on elevation angle. Although, $\mathbf{a}_{az}$ depends on both azimuth and elevation angle, elevation angle span in the ducting channel is between -0.5° - 0.5° with respect surface even at extreme cases as shown in Figure 10.3. Therefore, narrow elevation span of ducting channel can be used to adjust azimuth and elevation beams separately as will be discussed in Section 10.4.

At the end, channel gain matrix can be simplified as

$$\mathbf{H} = \sqrt{\frac{1}{L}} \sum_{i=1}^L \alpha_i (\mathbf{a}_r^{az}(\theta_r, \phi_r) \otimes \mathbf{a}_r^{el}(\phi_r)) (\mathbf{a}_t^{az}(\theta_t, \phi_t) \otimes \mathbf{a}_t^{el}(\phi_t))^H \quad (10.7)$$

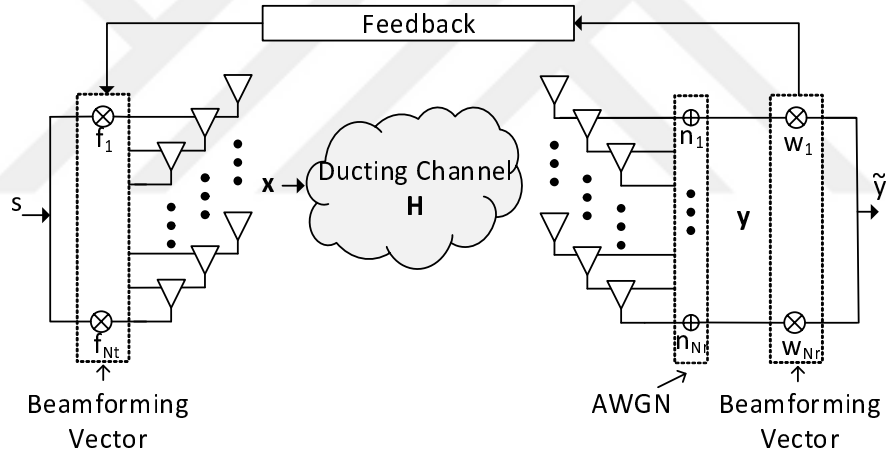


Figure 10.4: Ducting channel beam-forming system model.

In Figure 10.4, block diagram of ducting channel system with planar array is presented. Transmitted signal can be represented as

$$\mathbf{x} = \mathbf{f}s, \quad (10.8)$$

where s is the transmitted symbol, $\mathbf{f}_{N_T \times 1}$ is the transmitter beam-forming vector. Received signal can be given as

$$\mathbf{y} = \sqrt{\rho} \mathbf{H} \mathbf{f} s + \mathbf{n}, \quad (10.9)$$

where $\mathbf{y}_{N_R \times 1}$ is the received signal, $\mathbf{n}_{N_R \times 1}$ is the noise vector. $\rho = PL \times P_T$ is the average received power, where P_T is the transmit power and PL is the path-loss which

is directly calculated with Equation (22) in [105]. Output of the receiver planar array becomes

$$\tilde{y} = \sqrt{\rho} \mathbf{w}^H \mathbf{H} \mathbf{f} s + \mathbf{w}^H \mathbf{n}, \quad (10.10)$$

where $\mathbf{w}_{N_R \times 1}$ is the receiver beam-forming vector.

We consider single-user to single-user link. However, the proposed technique can be also extended for multiple user case by precoding matrices [119]. Therefore, spectral efficiency for single-user to single-user link can be found as

$$R = \log_2 (|1 + \rho / (N_T \sigma_n^2) \mathbf{w}^H \mathbf{H} \mathbf{f} \mathbf{f}^H \mathbf{H}^H \mathbf{w}|) \quad (10.11)$$

where σ_n^2 is the noise power which is modeled as thermal noise. (10.11) is a well-known equation and optimal beam-forming vectors are found via singular value decomposition. For $\mathbf{H} = \mathbf{U} \mathbf{\Lambda} \mathbf{V}^H$, optimal beam-forming vectors are $\mathbf{f}_{opt} = \mathbf{v}_1$ and $\mathbf{w}_{opt} = \mathbf{u}_1$. However, full CSI is required at both end to achieve the optimal performance. Considering the high path-losses and required high transmission powers for military b-LoS communication, transmission of full CSI is not possible in real world conditions. Thus, we propose a limited feedback multistage beam-forming algorithm for ducting channels.

10.4 Multistage Beam-forming Algorithm

Elevation angle span of ducting channel is very narrow as seen in Figure 10.3. Therefore, both receiver and transmitter is required to point their beams horizontal to the surface with very narrow beam-widths. In this way, transmitter avoids wasting the transmitting power by maximizing the amount of trapped signal energy. For 20 m surface duct with 6 M-units duct strength, trapping angle range in the elevation cut becomes -0.25° - 0.25° . To focus the antenna beam-width to this range, 200 vertical antenna columns is required. The number of antenna elements in the azimuth cut is determined to provide desired antenna gains because the channel model in [105] is 2D. As an example, the response of $W = 20 \times H = 200$ planar array is presented in Figure 10.5 by using phased array toolbox in MATLAB. As noticed, 200 vertical elements

can achieve 0.25° beam-width (Ψ_{el}). Since directivity is not so sharp in azimuth cut, less number of antennas can be placed in horizontal axis to reduce the system complexity. However, it is not possible to transmit CSI for such a large antenna array, thus we propose a limited feedback beam-forming algorithm.

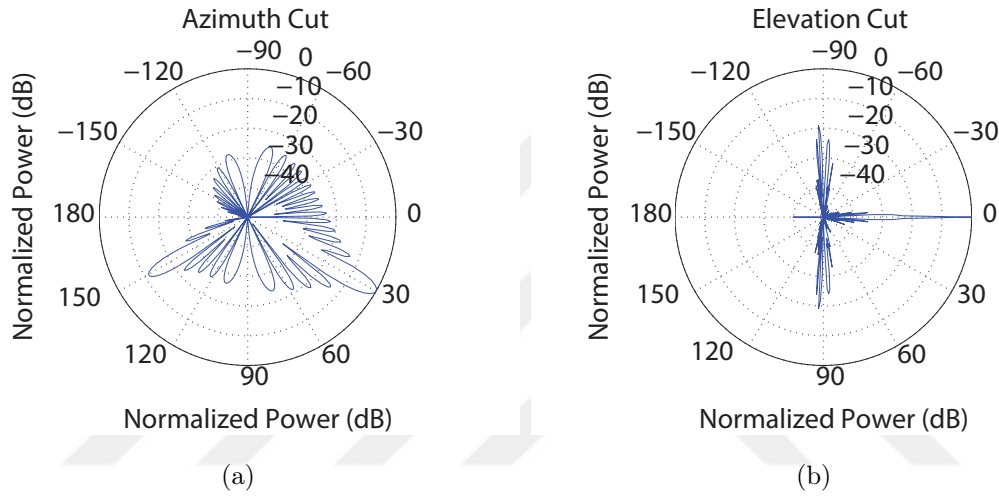


Figure 10.5: Planar array response for (a) azimuth cut ($\Psi_{ez} = 2.52^\circ$), and (b) elevation cut ($\Psi_{el} = 0.25^\circ$).

(10.4) depends on both azimuth and elevation angle, and (10.5) depends on only elevation angle. Although (10.4) is dependent to both angles, we can separate azimuth and elevation beam-forming thanks to sharp elevation range. Consider planar arrays on yz axis and antennas are separated in x axis. For this link, ϕ in (10.4) and (10.5) becomes zero. However, waves in oceans or sea can cause changes in this angle and significantly affect the link quality. To avoid this problem, system will utilize only some portion of vertical elements when scanning azimuth direction to detect desired unit. In this way, elevation beam-width of the system will be considerably increase as shown in Figure 10.6 for 20×10 planar antenna array. With the increasing beam-width, receiver can detect the transmitter in wavy oceans. The number of active vertical elements can be adjusted according to operation conditions. After finding azimuth vector, all of the vertical elements are start to operate to focus the beam in

elevation cut to desired direction and increase array gain.

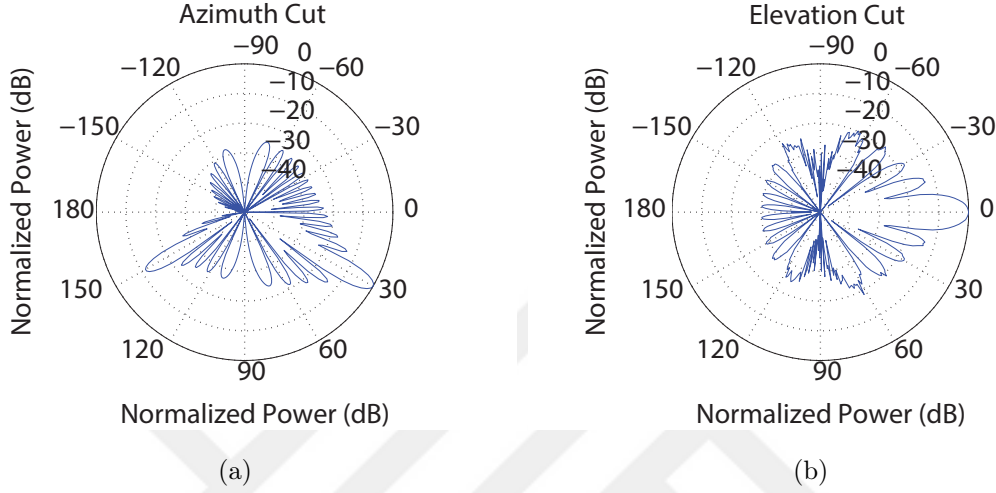


Figure 10.6: Planar array response for (a) azimuth cut ($\Psi_{az} = 2.52^\circ$), and (b) elevation cut ($\Psi_{el} = 5.12^\circ$).

In [116], 3D correlation matrix is decomposed as the Kronecker product of azimuth and elevation correlations. That is, azimuth and elevation beam-forming can be performed separately, and still similar eigenvalues can be maintained. Therefore, beam-forming vectors can be decomposed as

$$\mathbf{f}_{est} = \mathbf{f}^{az} \otimes \mathbf{f}^{el}, \mathbf{w}_{est} = \mathbf{w}^{az} \otimes \mathbf{w}^{el} \quad (10.12)$$

where $(\mathbf{f}, \mathbf{w})_{est}$ are the estimated transmit/receive beam-forming vectors with the proposed method. Azimuth and elevation beam-forming vectors is determined in two stage.

10.4.1 Azimuth Beam-forming Stage

In azimuth beam-forming stage, H_S number of vertical elements are used to get higher elevation beam-width. Therefore, elevation beam-forming vectors for azimuth

scanning stage can be formulated as

$$\begin{aligned}\mathbf{f}_S^{el} &= [1, e^{jkd(n \sin(\phi_T))}, \dots, e^{jkd(H_S-1) \sin(\phi_T)}, 0, \dots, 0]^T \\ \mathbf{w}_S^{el} &= [1, e^{jkd(n \sin(\phi_R))}, \dots, e^{jkd(H_S-1) \sin(\phi_R)}, 0, \dots, 0]^T\end{aligned}\quad (10.13)$$

where $\phi_{T,R}$ are the elevation angles which makes the beam horizontal to surface. In the simulations, we assume that planar arrays are placed on yz axis, thus $\phi_{T,R} = 0$. However, planar array can have any orientation or changing orientation because it is adaptively changed during second stage.

We assume that receiver has perfect CSI. Therefore, azimuth beam-forming vectors can be found by

$$\{\mathbf{f}_{est}^{az}, \mathbf{w}_{est}^{az}\} = \arg \max_{\forall \mathbf{w}_k \in \mathcal{W}, \forall \mathbf{f}_k \in \mathcal{F}} \|(\mathbf{w}_k \otimes \mathbf{w}_S^{el})^H \mathbf{H}(\mathbf{f}_k \otimes \mathbf{f}_S^{el})\| \quad (10.14)$$

where $\mathcal{F}$ and $\mathcal{W}$ are the azimuth codebooks for transmitter and receiver.

For codebook design, we utilize multi-resolution codebook approach used in [117] to provide faster search. In this approach, AoD range is divided into hierarchical sub-levels ($\{2\pi u/N\}_{u \in \mathcal{I}_{az}(s,k,m)}$). Consider S level codebook and $s = 1, \dots, s, \dots, S$ represent the codebook levels. In each level, there are $K^{(s-1)}$ subsets as shown in Figure 10.7(a). In each level, there are K number of beam-forming vector ($m = 1, \dots, k, \dots, K$) and represented as [117]

$$(\mathbf{f}_{s,k,m}^{az})^H \mathbf{a}_t^{az}(\theta_u) = \begin{cases} C_s & \text{if } u \in \mathcal{I}_{az}(s, k, m) \\ 0 & \text{otherwise} \end{cases} \quad (10.15)$$

where C_s is used for normalization and,

$$\mathcal{I}_{az}(s, k, m) = \{N/K^s(K(k-1) + m - 1) + 1, \dots, N/K^s(K(k-1) + m) + 1\} \quad (10.16)$$

Let $\mathbf{A}_T^{az} = [\mathbf{a}^{az}(\theta_1), \mathbf{a}^{az}(\theta_2), \dots, \mathbf{a}^{az}(\theta_N)]$, the beam-forming vectors can be designed as

$$(\mathbf{A}_T^{az})^H \mathbf{f}_{(s,k)} = C_s \mathbf{g}_{(s,k)}, \quad (10.17)$$

where $\mathbf{g}_{(s,k)}$ is a $N \times 1$ column vector which has 1's for $u \in \mathcal{I}_{az}(s, k, m)$ and 0's for $u \notin \mathcal{I}_{az}(s, k, m)$. Therefore, solution to (10.17) can be approximated as $\mathbf{f}_{(s,k)} = (\mathbf{A}_T^{az}(\mathbf{A}_T^{az})^H)(\mathbf{A}_T^{az})\mathbf{g}_{(s,k)}$. The codebook design for the receiver is the same. In the codebook search in (10.14), algorithm tries to find a lower level maximizing vector by throwing lower portions. When sub-levels are smaller than the upper level, the algorithm settles to the solution. At the end, the estimated transmitter azimuth beam-forming vector is sent back to transmitter.

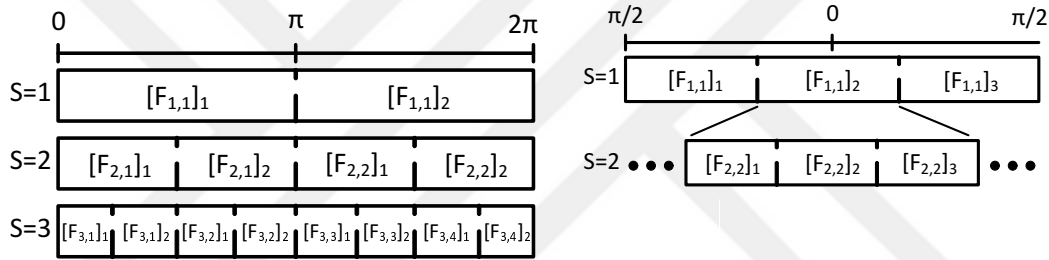


Figure 10.7: (a) Multi-resolution azimuth codebook for $N = 8$ and $K = 2$, (b) Elevation codebook for $N = 9$ and $K = 3$.

10.4.2 Elevation Beam-forming Stage

After azimuth beam-forming vectors are determined, unused elements in the vertical elements starts to operate to focus the antenna beams to the effective trapping range in elevation. Elevation beam-forming vectors are determined by

$$\{\mathbf{f}_{est}^{el}, \mathbf{w}_{est}^{el}\} = \arg \max_{\forall \mathbf{w}_k^{el} \in \mathcal{K}, \mathbf{f}_k^{el} \in \mathcal{L}} \|(\mathbf{w}_{est}^{az} \otimes \mathbf{w}_k^{el})^H \mathbf{H}(\mathbf{f}_{est}^{az} \otimes \mathbf{f}_k^{el})\|. \quad (10.18)$$

where $\mathcal{K}$ and $\mathcal{L}$ are the elevation codebooks for transmitter and receiver.

Since we assume that planar array is on yz axis, elevation beam is required to be centered at 0. For this reason, the major difference in elevation codebook design is the number of beam-forming vectors in each stage K . If K takes a even value, we will not have any code which centers at 0 as in Figure 10.7(a). However, beam codes can be centered on 0 for an odd K as in Figure 10.7(b). For sub-level angle range

$(\{\pi v/N - \pi/2\}_{v \in \mathcal{I}_{el}(s,k,m)})$, elevation codebook can be found with

$$(\mathbf{f}_{s,k,m}^{el})^H \mathbf{a}_t^{el}(\phi_v) = \begin{cases} 1 & \text{if } v \in \mathcal{I}_{el}(s,k,m) \\ 0 & \text{otherwise} \end{cases} \quad (10.19)$$

For the elevation codebook, the same methodology is utilized as in azimuth case. At the end, elevation beam-forming vector is also fed back to transmitter and spectral efficiency for the proposed method can be calculated by (10.11) and (10.12).

10.5 Simulation Results

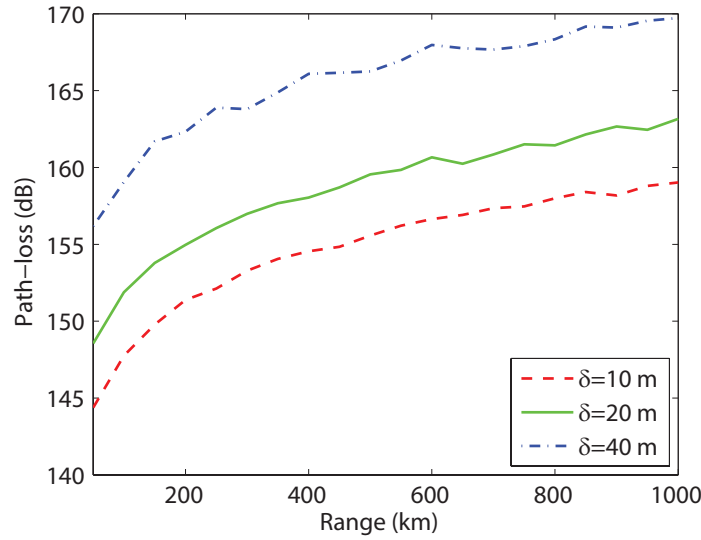


Figure 10.8: Ducting channel path-loss ($\Delta M = 10$ M-units)

In this section, we provide simulation results for the proposed method by using the ducting channel model [105]. The simulations are performed in MATLAB. 1000 realizations are used for path-loss and 1000 realizations are utilized for channel gain matrix. Figure 10.8 shows the path-loss for different duct heights for 5 m transmitter and receiver, 10 M-units duct strength, and 10.5 GHz frequency. Lower duct heights have lower path-loss because signal spreading increases with increasing duct height. We utilize $N = 256$ and $K = 2$ azimuth codebook, and $N = 729$ and $K = 9$ elevation

codebook to support lowest of 0.25° elevation beam-width. The codebook size can be adaptively changed according to operation environment. 20×200 planar antenna arrays are used at both sides, and only 20 vertical rows are used in the azimuth beam-forming stage.

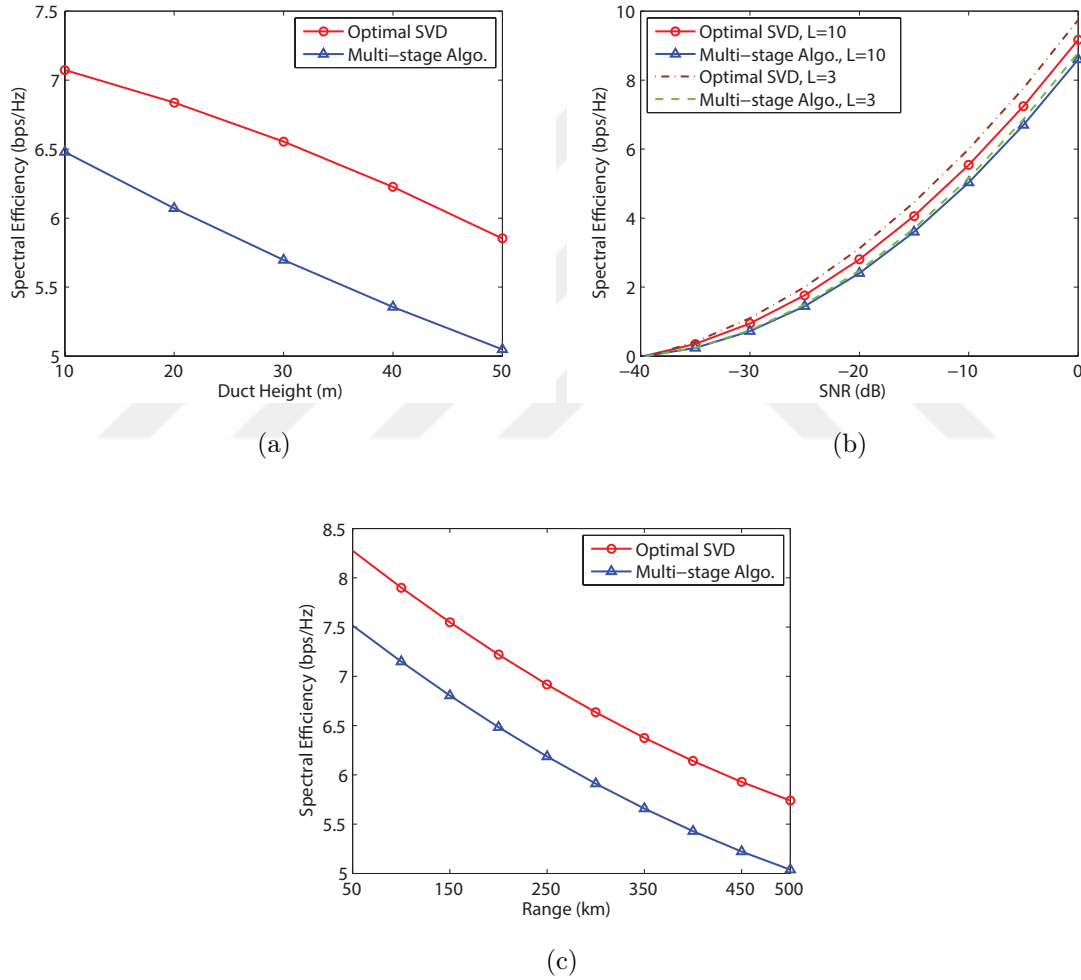


Figure 10.9: (a) Spectral efficiency vs. duct height ($L = 10$), (b) Spectral efficiency vs. SNR (10 m duct height), (c) Spectral efficiency vs. range ($L = 10$, 10 m duct height).

Figure 10.9(a) shows the simulation results for spectral efficiency of the ducting channel with the following channel parameters: 1 W transmit power, 5 m transmitter

and receiver, 10 M-units duct strength, 10.5 GHz frequency, number of paths $L = 10$, bandwidth 10 MHz, noise temperature 300 K ($\sigma_n^2 = kTB$ where k is the Boltzman constant, T is the temperature and B is the bandwidth), 200 km range. As expected, the spectral efficiency decreases with duct height because signal spreading increases as the duct height increase. Figure 10.9(b) shows spectral efficiency vs. SNR for 200 km link with 10 m duct height. As noticed, the proposed multi-stage beam-forming method is close to the optimal case for different number of received paths. As the number of paths increase, the performance gets lower because the beam-forming vector becomes wider. Lastly, Figure 10.9(c) presents spectral efficiency vs. range for 10 m duct height and $L = 10$. As expected, spectral efficiency decreases with range. However, the link can maintain 5 b/Hz spectral efficiency at 500 km. Therefore, we can conclude that ducting channel can be utilized in high data rate b-LoS applications. In addition, the proposed limited feedback method has close performance to optimal case as seen in Figure 10.9(b).

Chapter 11

COVERAGE AND AREA SPECTRAL EFFICIENCY IN DUCTING AD HOC NETWORKS

11.1 Introduction

Modern naval communications require alternative techniques for b-LoS communications because available b-LoS communication techniques have coverage and security problems. On the other hand, ducting channel communication can be utilized for direct communication medium up to 400 km with millisecond transmission delays and low probability of detect/intercept [105]. Since the trapping effect of atmospheric ducts significantly limits the signal spreading, the detect/intercept probability of the link by a plane or unmanned aerial vehicle is considerably low [61]. For these reasons, b-LoS ducting communication is a promising alternative for naval communication to provide a network between different fleets that do not have line-of-sight links as shown in Figure 11.1. We assume that the communication within a fleet (unmanned-air vehicle to navy ship, and navy ship-navy ship) is performed via conventional methods. However, ducting channel is utilized for the communication between different fleets.

Modern military applications require infrastructure-free communication between units. Thus, ad hoc networking method provides flexible networks without requiring any infrastructure. Although atmospheric modeling and propagation within the duct layer have been attracted research attention, there is no work which considers ducting ad hoc networks. Therefore, the main motivation of this work is to develop a model to analyze ducting ad hoc networks in terms of coverage and ASE.

The main contributions of this chapter is two-fold. Firstly, we provide a system model for ducting b-LoS ad hoc networks to perform coverage probability and ASE

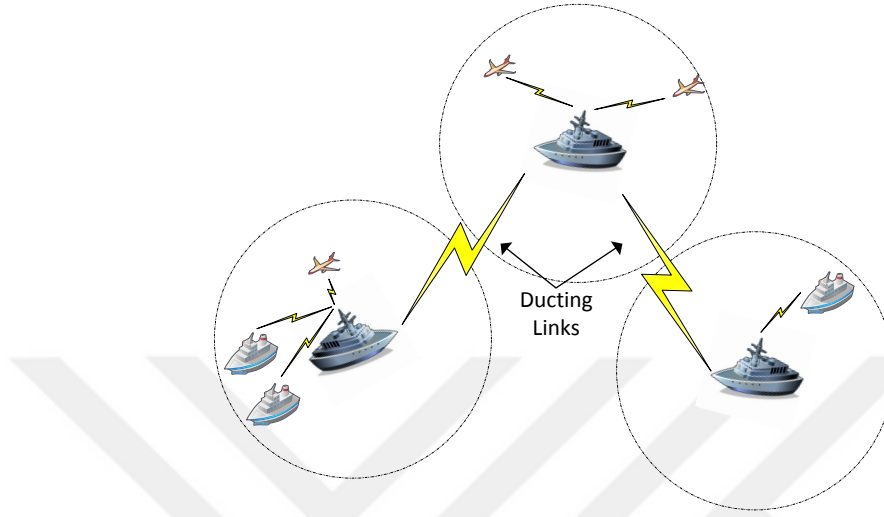


Figure 11.1: Naval ducting ad hoc networks between fleets.

analysis. For the system model, we utilize a recent channel modeling approach to the surface ducts [105]. In addition, link budget analysis is utilized to determine the communication range by using the experimental duct height statistics [79, 80]. Secondly, we compare the performance of the main directional MAC algorithms to determine the main challenges and possible solutions for the ducting ad hoc networks.

The remainder of the chapter is organized as follows. Section 11.2 includes the summary for atmospheric ducts, path-loss model introduced in [105], and switched beam antenna model utilized in this chapter. In Section 11.3, the system model for the coverage and ASE is proposed. In addition, link budget analysis and MAC protocols for ducting ad hoc networks are discussed in Section 11.3. Simulations results are provided in Section 11.4.

11.2 Ducting Channel and b-LoS Antennas

In this section, we provide an introduction for atmospheric ducts and a brief summary for the path-loss model proposed in [105]. In addition, switched beam antenna model which will be utilized in the simulations is also developed based on the ray analysis provided in [105].

11.2.1 Atmospheric Ducts

Atmospheric ducts are caused by the rapid changes in the refractivity of the troposphere, and form a layer in which the propagating signals are guided back to the surface. In this way, the signal energy is concentrated between the duct layer and surface instead of spreading through the upper layers of the troposphere. This effect is called as *trapping effect* of the atmospheric ducts, and the layer between the surface and duct is called as ducting layer.

Atmospheric ducts can be divided into four categories according to their refractivity model and formation process: evaporation ducts, surface-based ducts, elevated ducts, and surface ducts [22]. Instead of refractivity, modified refractivity is highly used when modeling atmospheric ducts as

$$M = N + (h/R_0) \times 10^6 \text{ M-units}, \quad (11.1)$$

where R_0 is the radius of the earth in km, h is the height in km, and refractivity $N = (n - 1) \times 10^6$ N-units, where n is the atmospheric refractive index.

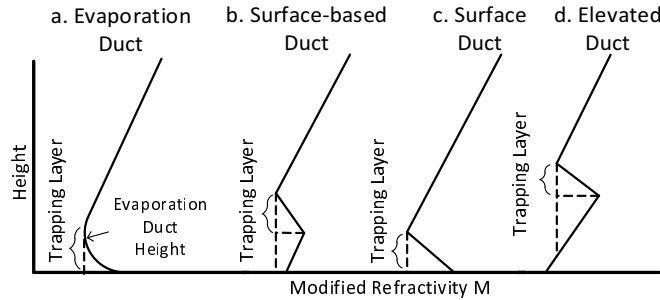


Figure 11.2: Modified refractivity profiles for atmospheric ducts.

Figure 11.2 shows the modified refractivity models for the atmospheric ducts. The height in which the modified refractivity gradient turns positive to negative is called as duct height. Trapping layer is formed between the surface and duct height as shown in Figure 11.2. The signals having particular grazing angles are trapped in the duct layer, and these signals can propagate over-the-horizon with significantly

lower path-loss compared to standard atmosphere. As shown in Figure 11.2, there are four types of atmospheric ducts. Surface-based and elevation ducts have lower formation probability compared to the other types. Thus we only consider evaporation and surface ducts for this chapter. Evaporation ducts are caused by humidity difference in the air-sea boundary, and they are present at the 90% of the time in maritime and coastal areas [62, 63]. Evaporation duct heights are generally between 10 – 20 m [22]. Furthermore, surface ducts are caused by the humidity difference due to air movements. Surface duct heights also vary between 10 – 20 m most of the time. However, surface ducts can reach higher heights with lower probabilities. Although the modified refractivity of evaporation ducts are modeled with a logarithmic curve, evaporation ducts can be approximated as a surface duct which has a bi-linear modified refractivity model as shown in Figure 11.2 [113]. Evaporation and surface ducts are present almost all the time in maritime and coastal areas [19]. With this approximation, we can utilize recently developed path-loss and ray analysis [105] to analyze naval ducting ad hoc networks.

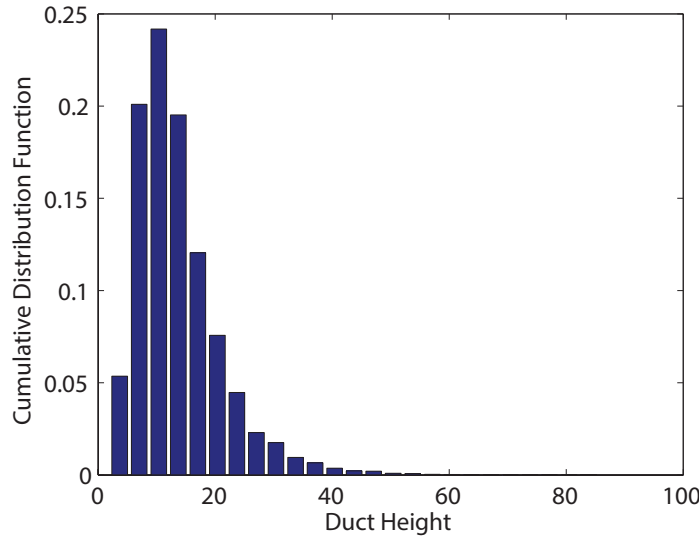


Figure 11.3: Duct height distribution modeled with a log-normal distribution with $\mu = 2.5$ m and $\sigma = 0.5$.

Surface ducts have a bi-linear modified refractivity profile. This profile can be represented with duct height (δ) and duct strength (ΔM). Duct height is the height in which the modified refractivity gradient changes its sign, and duct strength is the amount of change in the modified refractivity from surface to top of the duct. Experimental studies provides duct height and duct strength for coastal areas in [79, 80]. Therefore, we utilize a statistical fitting to the provided summer time duct heights in [80].

We observe that the duct height distribution presented in [80] (Figure 2 and 4 in [80]) is close to the log-normal distribution with $\mu = 2.5$ m and $\sigma = 0.5$ as in Figure 11.3. In addition, duct strength value varies between 4 – 10 M-units in [79]. Hence, we utilize this distribution in link budget analysis to determine more realistic range values in Section 11.3. We assume that the modified refractivity decreases by 118 M-units/km above duct height as in standard atmosphere conditions.

11.2.2 Path-loss Model for Surface Ducts

Ducting channel wave propagation studies are dominated by PE techniques which can handle complex refractivity conditions. The detailed overview of PE methods can be found in [22, 67, 68]. There are available wave propagation tools for modeling atmospheric ducts. In recent studies [22, 105], PETOOL [70] is utilized to model path-loss in atmospheric ducts because PETOOL is a free software and calibrated with an experimentally validated wave propagation software: AREPS [69]. Figure 11.4 contains the path-loss results for a 20 m surface ducts, 10.5 GHz frequency and 5 m transmitter/receiver heights. As noticed in the figure, surface duct provides significant path-loss enhancement compared to free space path-loss up to 10 – 25 dB, and the similar path-loss enhancements were observed in the experimental studies as well [20, 21, 120].

In [105], a path-loss model for surface ducts are proposed. The authors perform regression analysis to PETOOL results to develop a statistical path-loss model in [105]. In this model, large-scale path loss equation is fitted by the regression analysis

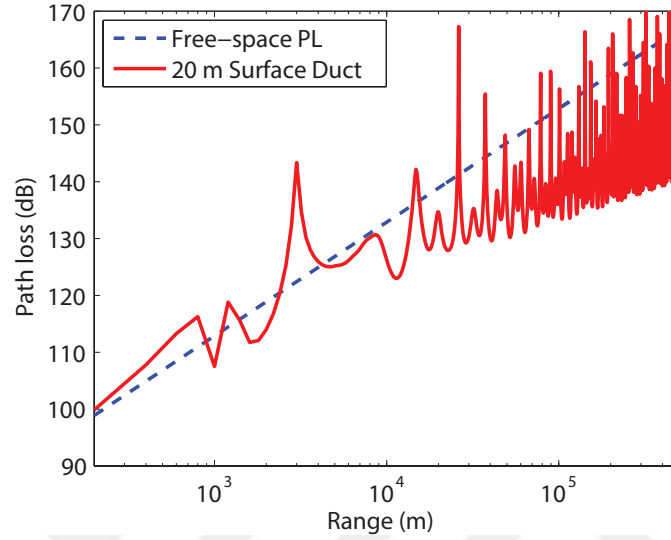


Figure 11.4: Path-loss results with PETOOL and free-space path-loss.

and it can be given as [27]

$$PL_{dB} = A + 10\gamma \log d/d_0 + X \text{ dB}, \quad (11.2)$$

where A is the path-loss (dB) at d_0 , d is the range (m), γ is the path-loss exponent, X_s is the shadow fading and X is the small-scale fading. A and γ depend on transmitter height, duct height, duct strength, and carrier frequency along with random variables. X_s represents Weibull distribution, and X can be modeled as Rayleigh fading as suggested in [105]. Minimum b-LoS range (d_0) is given by maximum LoS range equation for standard refractivity conditions as

$$d_0 = 4.12\sqrt{h_t} + 4.12\sqrt{h_r} \text{ km}, \quad (11.3)$$

where $h_{t,r}$ are the transmitter and receiver heights in m. Since this path-loss model depend on many random parameters, it is not possible to derive tractable equations for the coverage and rate distributions. Thus, ducting ad hoc networks are analyzed with Monte Carlo simulations in Section 11.4.

11.2.3 Antenna Model for b-LoS Links

In this subsection, we provide the antenna model that is utilized for ducting channel by analyzing the AoA conditions. In addition to PE based wave propagation analysis, ray-tracing methods are also promising for estimating delay and AoA of the ducting channels [69, 72].

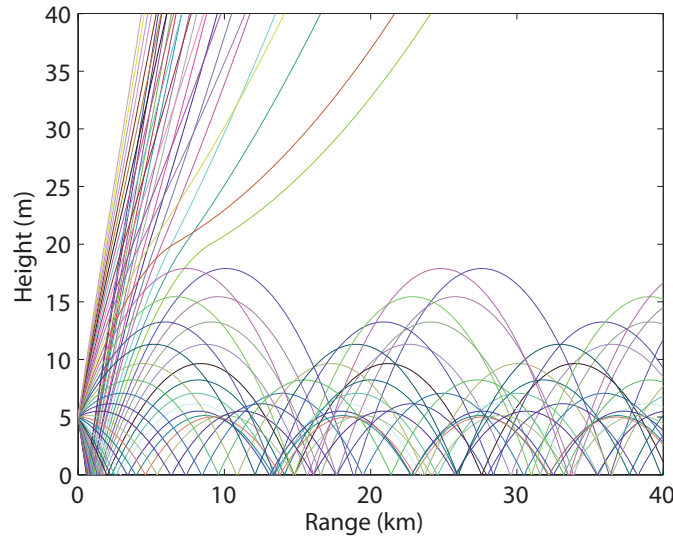


Figure 11.5: Ray trajectories for a 20 m surface duct.

In Figure 11.5, propagating rays in a 20 m surface duct are presented for $\Delta M = 6$ M-units and 10 m transmitter/receiver heights by using the ray-tracing method proposed in [105]. As noticed in Figure 11.5, only the rays having a certain angle range can be trapped by the duct. The rays outside of this range cannot be trapped and their power is wasted. To maximize the trapped power, the transmitter and receiver should focus their beams to the effective trapping beam-width. To this end, analytical expression for the effective trapping beam-width is provided in [105]

$$\theta_{max,min}^T = \pm \sqrt{2 \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) (h_t - \delta)}, \quad (11.4)$$

where refractive index derivative dn/dz ($dn/dz = -10^{-6}(\Delta M/\delta - 0.157)$) [22], δ is the ducts height in m, h_t is the transmitter height in m, R_0 is the earth radius (≈ 6370

km) and $n(0)$ is the index of refraction at surface (≈ 1.00035) [22]. In Figure 11.6, the effective trapping beam-widths in the elevation cut are presented as a function of duct strength and duct height for 5 m transmitter height. Since the duct heights can reach up to 40 m as discussed in 11.2.1, antenna beam-width can be designed according to the effective trapping beam-width of $\approx 0.15 - 0.25^\circ$.

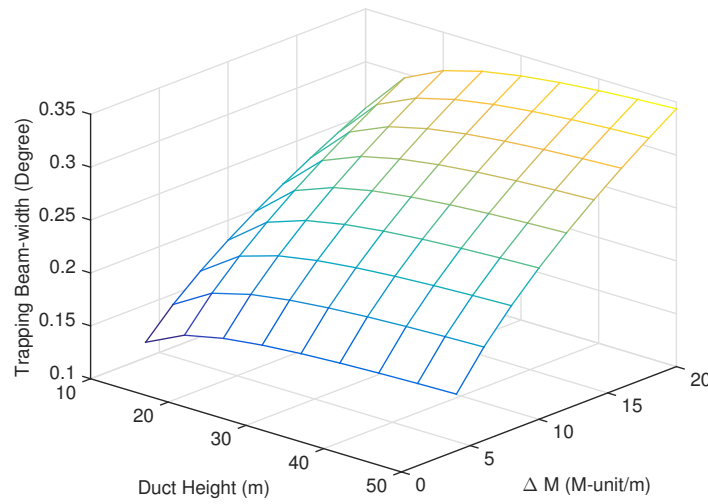


Figure 11.6: Effective trapping beam-widths (antenna height 5 m).

In [113], planar antenna arrays are suggested for the ducting communication links to focus the antenna beam in both azimuth and elevation to maximize the achievable data rates. According to their analysis, 200 vertical antenna columns are required to provide -0.25° - 0.25° 3dB beam-width in the elevation cut, and 20 horizontal antenna elements are required to provide high antenna gains as assume in the link budget calculations in [105]. Antenna response for of $W = 20 \times H = 200$ planar array is presented in Figure 11.7 by using the phased array toolbox in MATLAB. As noticed, elevation cut has very sharp beam, and azimuth can scan for the intended receiver.

In this chapter, we utilize a simplified switched beam antenna instead of the adaptive planar array that is utilized in [113]. However, we use the adaptive antenna array analysis to determine the parameters for the switched beam antenna. We assume

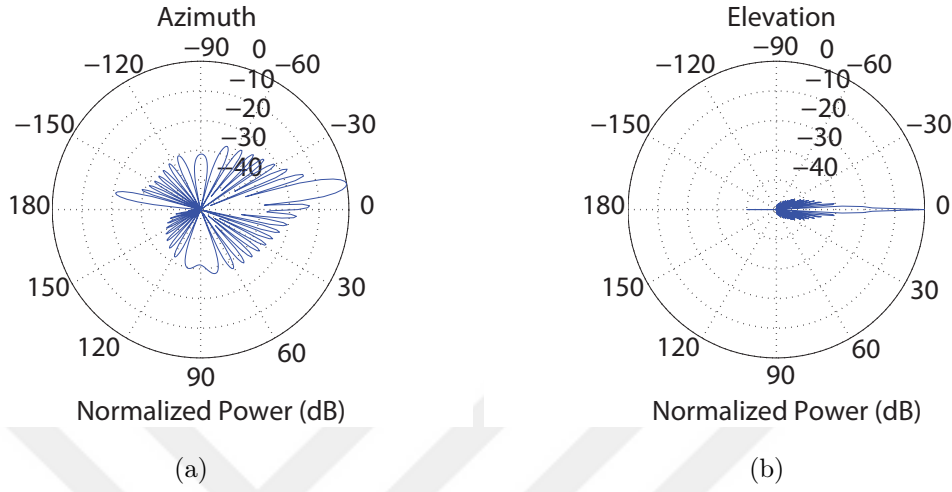


Figure 11.7: $W = 20 \times H = 200$ planar array response for (a) azimuth cut (2.5° 3dB beam-widths), and (b) elevation cut (0.25° 3dB beam-widths).

that the main lobe has the directional gain G_d , and the remaining region is assumed to have the omni directional gain G_o as shown in Figure 11.8. The width of directional beams are assumed as 5° because $W = 20 \times H = 200$ planar array provides 2.52° 3 dB beam-width. Associated directional gain is calculated as 35 dB, and assumed the same for all the beams. The side-lobe gain is assumed constant and assumed as 15 dB because the gain difference in Figure 11.7(a) is close to 20 dB on average as determined with the phased array toolbox in MATLAB. We also assume that the antennas are placed horizontally to send the signals with the calculated angle range in the elevation cut. Since we use a simplified antenna to analyze the ducting ad hoc networks, we do not consider interference cancellation. However, it is a significant open research issue for ducting communications.

11.3 Coverage Probability and Area Spectral Efficiency Analysis

Ducting communication is a promising candidate for military communications which require infrastructure free networking. For this reason, ad hoc networking is a necessity for ducting b-LoS networks. Since ducting b-LoS links utilize directional antennas

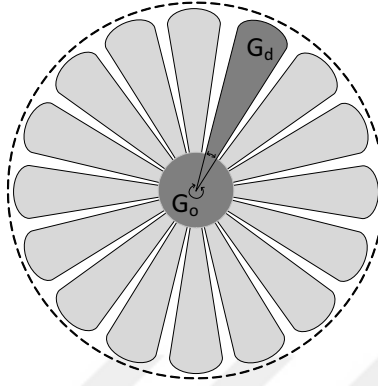


Figure 11.8: Switched beam antenna in the azimuth cut.

as described in Section 11.2.3, high spatial reuse can be provided to increase the overall throughput. However, the system performance can be significantly reduced due to interference. In ducting links, the trapping effects provide lower path-loss which enables b-LoS links. In the same way, the interfering nodes also have low path-loss and they can interfere with a station 400 km away. Therefore, the interference coordination in ducting b-LoS networks imposes new challenges.

In this section, we include the link budget analysis to analyze the communication range, the system model for the coverage and ASE analysis, and MAC protocols to handle the interference in the network.

11.3.1 Range Analysis

Communication range is an important metric for ad hoc network design. For this reason, we provide link budget calculations for the utilized link parameters. In [105], the communication range for a 40 m surface duct is calculated as 400 km with 10 GHz carrier frequency. In this study, we extend the link budget analysis by including the statistical duct height distribution as shown in Section 11.2.1. In this chapter, we select the carrier frequency as 10.5 GHz because lower duct heights (10 – 20 m) have the lowest path-loss near this frequency as shown in [22, 121] unlike the results presented in [105] for 40 m duct. Since this duct height range is more probable as in the

experimental studies [79, 80], 10.5 GHz is more promising for ducting communication links.

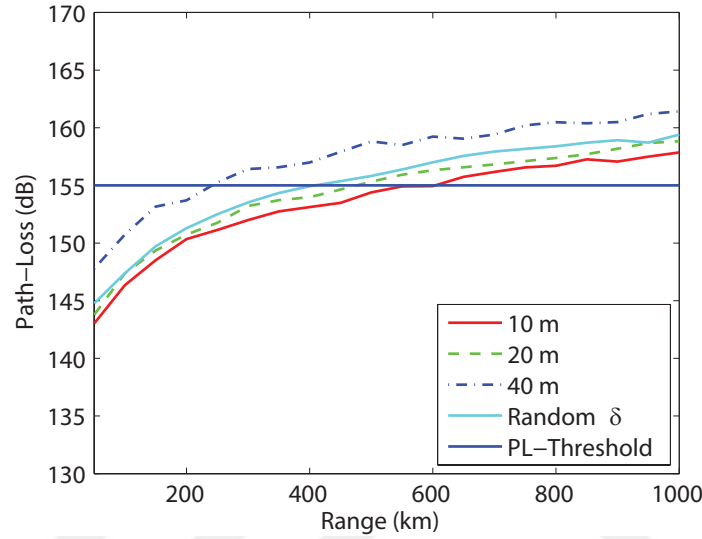


Figure 11.9: Path-loss results vs. range for different duct heights.

For the transmission range analysis, we perform link budget calculations with the following link parameters:

- Transmit power (P_t) = 30 dBm (1 W),
- Transmitter antenna gains (G_t) = 35 dB,
- Receiver antenna gains (G_r) = 35 dB,
- Cable losses (L_q) = 5 dB,
- Receiver sensitivity = -60 dBm.

With these channel parameters, the path-loss threshold for the ducting link is found as 155 dB, and the path-loss value is required to be lower than this value in order to maintain the transmission. Figure 11.9 shows the path-loss vs. range results for 90%

non-exceeding of the time with 5 m transmitter and receiver height, 10.5 GHz carrier frequency, and 6 M-units duct strength. As noticed, lower duct heights provides higher ranges because the signal spreading is further limited with the decreasing duct height. In addition, we provide the results for the log-normal duct height distribution as discussed in Section 11.2.1. The random duct height distribution maintain the ducting link up to 400 km range. Although random duct heights are mostly between 10 – 20 m as in Figure 11.3, the range level is close to 20 m duct height because when the duct height is lower than the antenna height the path loss becomes 180 – 200 dB as analyzed with PETOOL [70].

11.3.2 System Model for Ducting Ad Hoc Networks

In this subsection, we introduce the system model for ducting ad hoc networks to analyze the coverage probability and ASE of the network. We utilize the path-loss model for the surface ducts which is proposed in [105] as summarized in Section 11.2.2, and every node is equipped with a switched beam antenna which can scan 360° as described in Section 11.2.3. The antennas are placed at 5 m from the sea surface, and it may be required to put one antenna each side of a navy ship to 360° coverage. Since surface ducts are present up to 80% of the time and evaporation duct are present up to 90% of the time [22, 79], the received power enhancements with the atmospheric ducts occur almost all the time.

The distributions of the b-LoS nodes are determined via homogeneous Poisson point process (PPP) on the 2D space which is assumed as a flat ocean. The node density are controlled via intensity parameter λ and the node locations are distributed uniformly over $1000 \times 1000 \text{ km}^2$ region. PPP is highly utilized while analyzing ad hoc networks [123, 124]. We assume that the network nodes can start to transmit data with probability p or stay silent with probability $1 - p$. In this way, the effective node probability becomes $\lambda := p\hat{\lambda}$. Figure 11.12 presents random node distribution generated with PPP with $\lambda = 5 \times 10^{-5}$, and $\lambda = 10^{-4}$. In the uncoordinated case, the nodes utilize this ALOHA type protocol. However, how other MAC algorithms

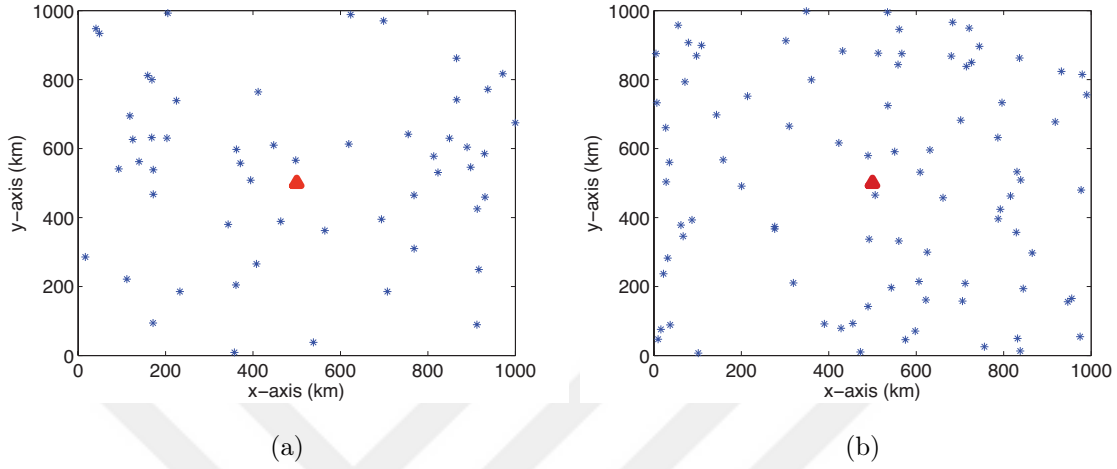


Figure 11.10: Random b-LoS node distribution with (a) $\lambda = 5 \times 10^{-5}$, and (b) $\lambda = 10^{-4}$

are modeled is discussed in the next subsection.

We assume that the b-LoS nodes utilize the switched beam antenna model developed in Section 11.2.3. In our model, the intended receiver is placed in the middle of the plane, and a transmitter with d km away from the receiver is communicating with it. After the transmitter sends its packet to the intended receiver, the receiver sends an acknowledgment packet upon successful transmission as in IEEE 802.11. Since the total antenna beam-width in the azimuth cut is $\theta \approx 2.5^\circ$ [11.7], there will be 72 beam in the each node. The active node between the transmitter and receiver is determined randomly. Interfering nodes also transmits to a randomly determined direction with probability p . The signal-to-interference-plus-noise ratio (SINR) is given by

$$SINR = \frac{P_t G_0 PL_0}{N_0 + \sum_{j \in \mathcal{I}} P_t G_j PL_j} \quad (11.5)$$

where P_t is the transmit power, PL_i is the path-loss for i th node which depends on distance and duct characteristics. G_0 is the total antenna gain for the transmitter and intended receiver, and we assume that the transmitter and intended receiver has perfect alignment $G_0 = G_d \times G_d$. G_j 's are the interfering stations total antenna gain. Since the interfering stations choose their direction of transmission randomly with

uniform distribution, the total antenna gain for the interfering stations can be given by

$$G_j = \begin{cases} G_d G_d \text{ w. p. } p_{dd} = (\theta/\pi)^2 \\ G_d G_o \text{ w. p. } p_{do} = 2(\theta/\pi)(\pi - \theta/\pi) \\ G_o G_o \text{ w. p. } p_{oo} = (\pi - \theta/\pi)^2. \end{cases} \quad (11.6)$$

For the two-way coverage probability, the link should satisfy the SINR constraint in both transmitter and receiver side. Since the acknowledgment packets are send through the control channel and requires much lower data rates, we only check the minimum received power constraint for the receiver to transmitter link as discussed in Section 11.3.1. Therefore, we only define the coverage probability with transmitter to receiver SINR constraint as

$$P_c = \mathbb{P}[SINR > \gamma], \quad (11.7)$$

where γ is the SINR threshold. In the coverage probability analysis, 90% availability is considered to determine γ .

In ad hoc networks, throughput is an important metric which show the utilization of the wireless medium. For this reason, we use ASE of the system as introduced in [124]. ASE is defined as

$$\eta = \lambda_\epsilon \log_2(1 + \gamma)(1 - \epsilon) \quad (11.8)$$

where ϵ is the outage, λ_ϵ is the maximum user density that can be supported with γ threshold. We find ASE through Monte Carlo simulations for the described ducting ad hoc networks.

11.3.3 MAC Protocols

Conventional ad hoc networks utilize omni-directional antennas, and the utilization of the wireless medium is significantly limited. On the other hand, smart antenna based ad hoc networking have potential to improve the network throughput by offering

spatial reuse with the directional beams. To this end, smart antenna based ad hoc networking has been attracted significant research attention [125–127]. Although the directional antennas may increase the spatial reuse, they impose some challenges as well such as deafness, directional hidden node problem and underutilization of the medium due to queuing problems [125,127]. To solve these issues, many methods are proposed [128–130]. In [128], it is assumed that node locations are known through GPS and each node sends their location to their neighbors. Omni-directional CTS is utilized to avoid collisions at the receiver. Since CTS is sent omni-directionally, the range of the CTS message will be shorter. To solve this problem, directional MAC (DMAC/DA) [129] is proposed in which directional RTS/CTS (Request-to-Send/Wait-to-Send) is utilized. In addition, receiver send wait to send (WTS) packets to the neighbors to inform the nearby nodes. This method provides robustness against the interference. However, this technique results in excessive overhead due to control packets. Instead of RTS/CTS packet, tone DMAC utilizes tones to inform the neighbors about the communication. There are number of different algorithms, and a good review of them can be found in [125].

In this chapter, we only focus on the coverage probability and ASE because we aim to show that high number of nodes can be supported in ducting ad hoc networks with high communication ranges. In addition, we provide simulation results for the coverage probability and ASE for the following MAC protocols:

- **ALOHA-type:** In ALOHA-type protocol, the nodes send their data with a probability p to a randomly chosen direction. If the channel is not available or transmission is not successful, the node tries it again after some back-off time. We also assume that the receiver sends an acknowledgment upon a successful transmission. As discussed in Section 11.3.2, the minimum path-loss requirement is controlled for the acknowledgment messages because the size of the control packets are considerably lower compared to the data packets. This protocol is expected to suffer from high interference due to uncoordinated nature of this technique.

- **Single-user to Single-user:** In this case, we aim to show the performance for a interference free scheme in which only one transmitter uses the wireless medium at a time without any spatial reuse. Although this protocol forms an upper bound for the coverage probability, it will suffer from the ASE due to underutilization of the channel.

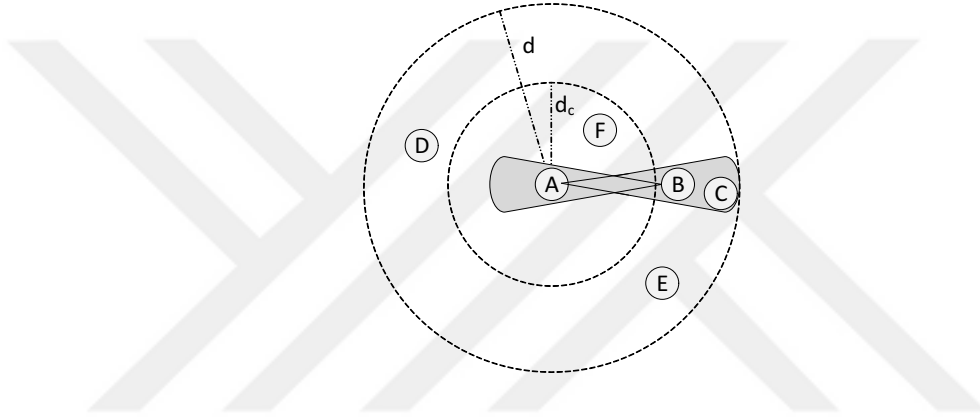


Figure 11.11: An example node distribution.

- **DMAC/B:** In directional MAC/beam (DMAC/B), we assume that the receiver sends control packets to inform the neighbors that are closer than the control range (d_c) about the incoming transmission. The nodes that take the control message do not send any packet in the direction of receiver beam. In Figure 11.11, there is a data transmission between transmitter B and node A. If node C is not aware of this communication and start to transmit data to node D, there will be significant interference at node A. For this reason, DMAC/B solves this interference problem, and do not avoid the data transmission between node F and node E because they are near orthogonal to the data transmission between node A and B. In addition, if node C and node E are communicating with each other while node A sends the control package, node C will be unaware of this transmission. In this case, there will be interference due to unheard control packets. We assume that the control packets reach to the neighboring nodes with a probability p_d .

Table 11.1: Simulation parameters.

Parameter	Value	Unit
Carrier frequency (f)	10.5	GHz
Bandwidth (B)	10	MHz
Transmitter height (h_t)	5	m
Receiver height (h_r)	5	m
Duct strength (ΔM)	6	M-units
Transmit power (P_t)	30	dBm
Directional gain (G_d)	35	dB
Side-lobe gain (G_o)	15	dB
Control packet drop prob. (No)	0.9	-
Noise temperature	-173	dBm
Control range (d_c)	300	km

- **DMAC/D**: In directional MAC/distance (DMAC/D), the control packets are sent to every neighbor that are closer than d_c km from node A, and all the receiving neighbors becomes silent with probability p_d as in DMAC/B. In this protocol, the main aim is to avoid the interference coming from the side-lobes. Since all the neighbors are blocked, this protocol will have higher coverage probability. However, it will suffer from the ASE due to lower spatial reuse compared to DMAC/B.

11.4 Simulation Results.

In this section, we provide simulation results for ducting ad hoc networks for the coverage probability and ASE. Simulations are performed for 1000×1000 km² region in MATLAB with the channel parameters in Table 11.1. The node density controlled with a homogeneous PPP with the intensity parameter λ , and each antenna is

equipped with a switched beam antenna as discussed in Section 11.2.3. 100000 iteration is performed for each scenario. In the simulations, the nodes randomly select a direction and start to transmit their data. However, some of the nodes are blocked in DMAC/B and DMAC/D to handle the interference between the nodes. In this case, the actual node density becomes lower because some of the active nodes are blocked. This effect is shown in ASE results.

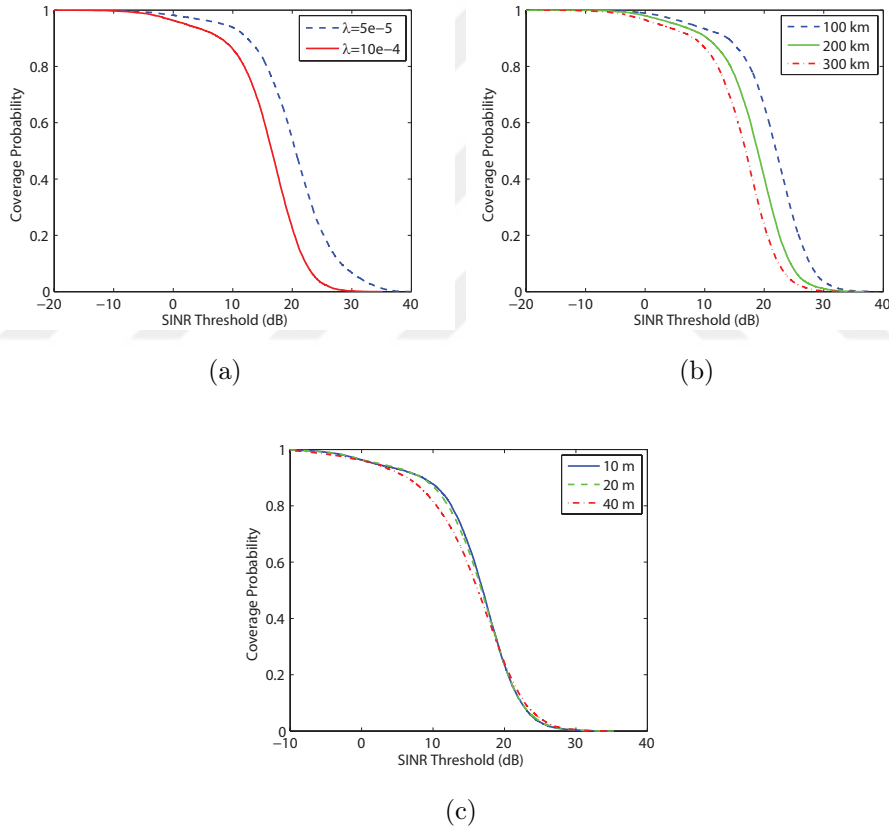


Figure 11.12: Coverage probability results for (a) node intensity with 300 km range, (b) range with 20 m duct height and (c) duct heights with 300 km range.

Figure 11.12 shows the coverage probability results for different scenarios under ALOHA-type MAC protocol. As noticed from Figure 11.12(a), higher node density results in lower coverage probability results for 300 km range and 20 m duct height. Figure 11.12(b) shows the coverage probability results for different ranges,

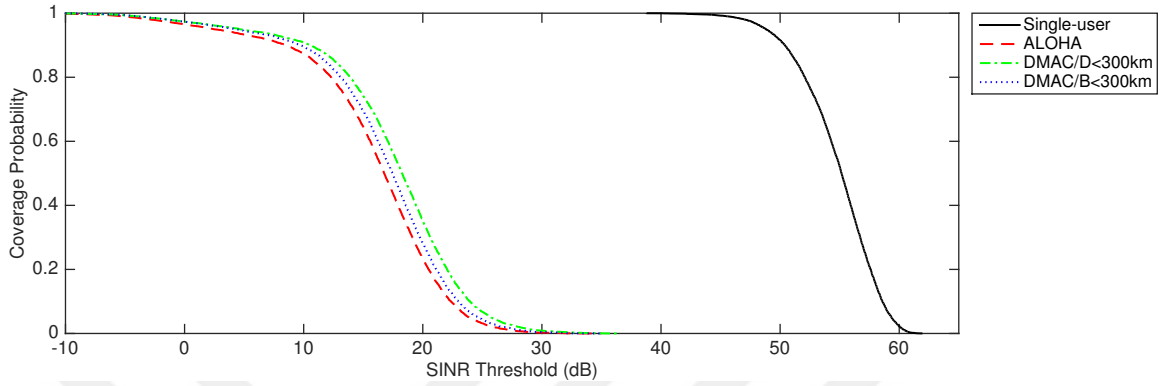
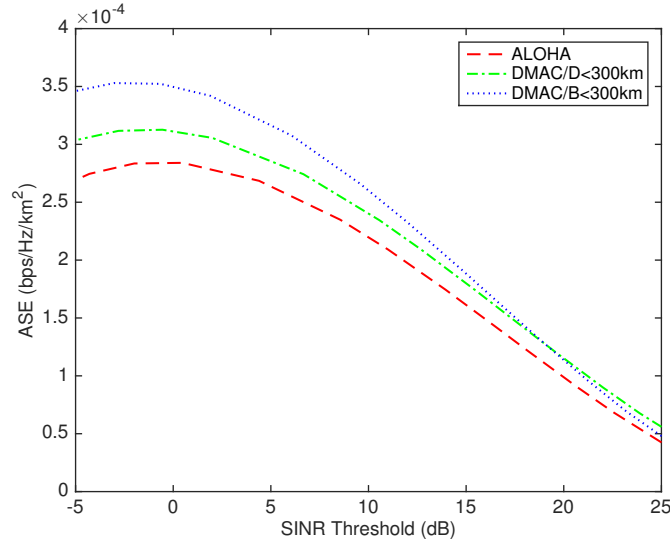


Figure 11.13: Coverage probability for the MAC protocols.

and the coverage probability lowers with the increasing range. Most importantly, Figure 11.12(c) presents the coverage probability for different duct heights. In Figure 11.9, it is shown that lower duct heights provide lower path-loss and higher communication range. However, the coverage probability for 10 m and 20 m curves are almost intersecting because interfering stations also have lower path-loss and this effect results in more interference at the intended receiver. For this reason, 10 m duct height cannot provide higher coverage probability compared to 20 m duct height.

Figure 11.13 shows the coverage probability results for the MAC protocols that are listed in Section 11.3.3 for 300 km range and 20 m duct height. As noticed, single-user to single-user case is interference free and provide significantly higher coverage probability even at very high SINR values. This results show that ducting ad hoc networks are interference limited. In addition, interference management techniques also provide better results for coverage probability. Although DMAC/D blocks higher number of nodes compared to DMAC/B, the coverage probability do not show significant deviation from DMAC/B in which only the direct interference beams are blocked. Since the number of blocked users are lower in DMAC/B, it is more promising for providing higher throughputs.

Figure 11.14 presents the ASE results for ducting ad hoc networks with 300 km range, 20 m duct height and 10% outage probability. As noticed, ALOHA-type proto-

Figure 11.14: Area spectral efficiency for $\epsilon = 0.1$.

col has the lowest ASE due to uncoordinated interference. Although DMAC/D provides better coverage probability results by blocking side-lobe interference, DMAC/B has better ASE results due to higher node density. For this reason, DMAC/B is more promising to provide higher network throughputs. We do not include single-user case because it has significantly low ASE. For 10% outage probability, SINR threshold becomes 50 dB, and it results in 1.8×10^{-5} bps/Hz/km² ASE for single-user to single-user case.

Chapter 12

CONCLUSION

In the proposed thesis, it is shown that troposcatter and ducting effects of the lower troposphere can be utilized in modern military communication systems to provide secure, reliable and high data rate links. This thesis mostly focus on the physical layer modeling of these propagation mechanisms, and also the concept of ducting ad hoc networks is introduced. To sum up, the following topics has been investigated:

- Physical layer channel model for troposcatter
- Investigation of main communication parameters for troposcatter
- Correlation analysis for main diversity techniques in troposcatter links
- Rain attenuation in troposcatter communications
- Physical layer channel model for ducting communication
- Investigation of main communication parameters for ducting links
- Correlation analysis and MIMO ducting communications
- Multi-stage Beam-forming for b-LoS Ducting Channels
- Rain attenuation in ducting communications coverage probability and area spectral efficiency

The following sections includes the detailed conclusions that we reached in each chapter.

12.1 *b-LoS Communications with Troposcatter*

In this topic, we provide a short summary of the troposcatter literature by reviewing the channel modeling approaches and possible diversity techniques. In addition, we discuss possible application areas of troposcatter communications. According to our reviews, we revealed that troposcatter communication is gaining momentum especially for fixed link military communications. Thus, Chapter 2 provides the state-of-the-art review of the troposcatter literature by outlining open research issues and application areas.

12.2 *A Ray-based Channel Modeling Approach for Troposcatter*

In Chapter 3, we propose a ray-based troposcatter channel modeling approach. In addition, we present important results for the transmission-loss, coherence bandwidth, RMS delay spread and correlation between vertically placed antennas for the first time with the real world conditions. According to our results, troposcatter communication can be used in b-LoS applications up to 300 km. In addition, b-LoS troposcatter communication can provide high data rates with high availability rates. For these reasons, troposcatter communication can be used as a communication medium for b-LoS applications which require high data rates and reliability. In addition, we present simulation results for coherence time and coherence bandwidth of the troposcatter channel. According to our analysis, the coherence time of the troposcatter channel is in the order of milliseconds. Therefore, the mobility in the troposcatter communication can be promising and, it is a significant open research area.

12.3 *Fading Correlation Model for MIMO-OFDM Troposcatter Communications*

Under this topic, RSM for fading correlation is proposed to investigate the fading correlation between antennas for space, frequency, angle and space-frequency diversity. In addition, we compare these diversity techniques with their distribution of achievable

data rates. According to our results, space-frequency diversity systems can provide more than 10% increase in the achievable data rates. Since the implementation of frequency diversity does not require additional parabolic reflector, space-frequency diversity systems are much more economical compared to adding additional antennas. Furthermore, the simulation results shows that the b-LoS troposcatter communication can provide more than 8 Mbps at all times. Therefore, troposcatter communication is a promising candidate for high data rate b-LoS communications with the employment of diversity techniques.

12.4 Rain Attenuation Model for Troposcatter Communication Links

In Chapter 5, we developed a rain attenuation model to predict the amount of rain loss in troposcatter links for non-homogeneous rain cells. According to our results, ITU-R P.452 underestimates the rain loss in the link. In addition, 4.7 GHz do not show significant changes with rain. However, 15 GHz shows significant rain losses. Although higher frequency ranges are able to provide more bandwidth, lower frequency ranges are more promising for troposcatter applications to provide higher link reliability.

12.5 b-LoS Communications with Ducting Layer

Ducting layer b-LoS communication is a promising communication method especially for military naval applications to empower tactical decisions in NCOs. Since ducting layer b-LoS communications show strong dependence on the refractive index which is highly affected by the atmospheric conditions, well-developed channel models are required to determine the characteristics of the channel. To this end, this chapter review the channel modeling approaches by outlining the open research areas, and possible application areas of the ducting layer b-LoS systems. More importantly, this work also provides the path-loss exponent for an ideal evaporation duct.

12.6 Channel Model for Surface Ducts: Path-Loss, Delay Spread, and AOA

In this topic, we develop a large-scale path-loss model for the surface ducts and analyze delay spreads and AoA of the ducting channel. According to our results, the ducting channel can be used as communication medium up to 1000 km at 5 GHz, and the small-scale fading in the channel shows Rayleigh fading. In addition, our RO methods can be utilized to estimate the fading behavior of the channel. Upper and lower bounds for the AoA of the channel can be estimated with the effective trapping beam-width equation.

12.7 Path-loss and Correlation Analysis for Surface Ducts

In Chapter 8, we develop a large-scale path-loss model for the ducting channel in the presence of atmospheric ducts. In addition, the distribution of fading and correlation between antennas for space and polarization diversity are analyzed. According to our results, polarization diversity is an effective diversity technique for ducting channels. In addition, 2×2 ducting channel communication can provide more than 22 Mbps at 90 percent of the time, thus modern high data rate applications can utilize ducting effects especially in maritime and coastal areas. In conclusion, this work reveals that ducting channel is suitable for high data rate communication applications.

12.8 b-LoS Ducting Channels: Coherence Time and Rain Attenuation

Under this topic, the coherence bandwidth and coherence time of the ducting links are analyzed. Our analyses show that ducting channel is expected to experience slow and narrow-band fading up to a few hundreds of MHz bandwidths. In addition, a slant-path rain attenuation model is proposed and, the amount of rain loss in the ducting links is presented. According to our results, 10.5 GHz is more promising to provide higher data rates.

12.9 Multi-stage Beam-forming for b-LoS Ducting Channels

In Chapter 10, we propose a limited feedback multi-stage beam-forming algorithm for b-LoS ducting channel for the first time in the literature. Based on our simulations, our proposed method provides 5 – 8% less performance compared to optimal conditions. Since it is not possible to maintain full CSI in a communication channel with high path-loss and very large antenna array, our method is promising for ducting channel b-LoS communications thanks to the limited feedback requirement. In addition, b-LoS ducting channel can achieve 5 b/Hz spectral efficiency even at 500 km. Therefore, high data rate naval b-LoS communications can be provided with atmospheric ducts with millisecond transmission delays.

12.10 Coverage and Area Spectral Efficiency in Ducting Ad Hoc Networks

In this topic, we develop a system model for analyzing ducting ad hoc networks by using the proposed path-loss and ray-tracing methods. Based on these methods, we perform link budget analysis by using the experimental duct statistics. According to our results, ducting channel provides high data rate direct communication links up to 400 km. In addition, the coverage and ASE analysis show that the network is interference limited, and the interference handling techniques are required in ducting channel. For this reason, we compare the basic DMAC protocols. According to our results, DMAC/B has potential to provide higher ASE at the expense of coverage probability. However, network simulation is required to further compare these techniques. In addition, power control techniques can be proposed in order to limit the interference between nodes and maintain higher spatial reuse. At the end, there are many open research problem on ducting ad hoc networks.

Chapter 13

APPENDICES

Appendix A: The Scattering Cross-section Calculations

The Rayleigh scattering is valid for the scattering by the atmospheric turbulence because the scattering particles are much lower compared to the wavelength between 4.7 – 10 GHz range. Beyond this frequency range, the Rayleigh scattering theory deviates from the real world values as reported in ITU-R-R P.452 [15]. Since the modern troposcatter equipments generally operate on 4.4 – 5 GHz [1], we focus on 1 – 10 GHz frequency range.

The derivations for the scattering cross-section are mostly taken from [12]. The scattering cross-section for a unit volume is given by

$$\sigma(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = |f(\hat{\mathbf{o}}, \hat{\mathbf{i}})|^2 / dV_c, \quad (13.1)$$

where $f(\hat{\mathbf{o}}, \hat{\mathbf{i}})$ is the scattering amplitude for the incoming ray $\hat{\mathbf{i}}$ and the scattered ray $\hat{\mathbf{o}}$ as seen in Figure 13.1(a). dV_c is the unit volume.

The scattering amplitude for Rayleigh scattering approximation is given by [12]

$$f(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = \frac{k^2}{4\pi} \int_{dV_c} \{ -\hat{\mathbf{o}} \times [\hat{\mathbf{o}} \times E(\hat{\mathbf{r}})] \} \epsilon(\hat{\mathbf{r}}) \exp(-ik\hat{\mathbf{r}} \cdot \hat{\mathbf{o}}) dV, \quad (13.2)$$

where $\times$ is the cross product, $\cdot$ is the dot product, $k = 2\pi/\lambda$ is the wavenumber, $\epsilon(\hat{\mathbf{r}})$ is the permittivity at $\hat{\mathbf{r}}$, and $E(\hat{\mathbf{r}})$ is the field at $\hat{\mathbf{r}}$ as seen in Figure 13.1(b).

Since the permittivity fluctuations is small for the atmospheric turbulences, we use the Born approximation [12]. In this way, $E(\hat{\mathbf{r}}) = \hat{\mathbf{e}}_i \exp(ik\hat{\mathbf{i}} \cdot \hat{\mathbf{r}})$, where $\hat{\mathbf{i}}$ is the polarization of the incident ray. Then, the scattering amplitude is simplified as

$$f(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = \hat{\mathbf{o}} \sin(\chi) \frac{k^2}{4\pi} \int_{dV_c} \epsilon(\hat{\mathbf{r}}) \exp(i\hat{\mathbf{k}}_s \cdot \hat{\mathbf{r}}) dV, \quad (13.3)$$

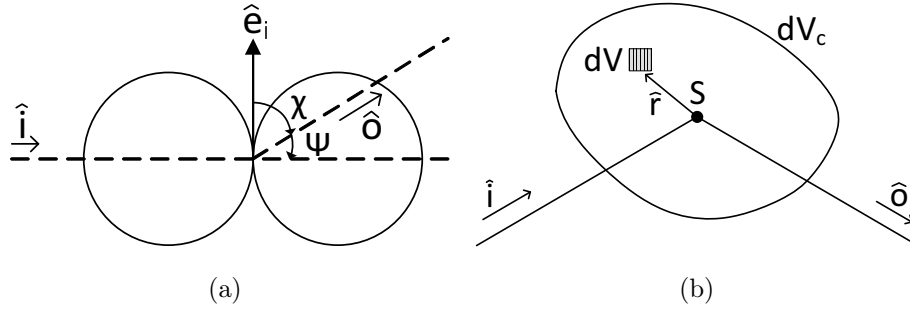


Figure 13.1: (a) Radiation pattern for Rayleigh scattering and, (b) Unit scattering common volume.

where $\sin(\chi) = \{-\hat{\mathbf{o}} \times [\hat{\mathbf{o}} \times \hat{\mathbf{e}}_i]\}$, χ is the angle between the polarization $\hat{\mathbf{e}}_i$ as in Figure 13.1(a). $\hat{\mathbf{k}}_s = k(\hat{\mathbf{i}} - \hat{\mathbf{o}})$ with magnitude $k_s = 2k \sin(\Psi/2)$, where Ψ is the scattering angle as seen in Figure 13.1(a). In this model, we use linear polarization at the both of the antennas. Therefore, the polarization of the incoming ray is always on the $\hat{\mathbf{e}}_i$ vector, and we can assume that $\sin(\chi) = \cos(\Psi)$.

By combining (13.1) and (13.3), the scattering cross-section is given as

$$\sigma(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = \frac{k^4 \cos(\theta)^2}{(4\pi)^2 dV_c} \int_{dV_c} \int_{dV_c} B_n(r_d) \exp[i\hat{\mathbf{k}}_s \cdot (\hat{\mathbf{r}}_1 - \hat{\mathbf{r}}_2)] dV_1 dV_2, \quad (13.4)$$

where $B_n(r_d) = \langle \epsilon(\hat{\mathbf{r}}_1) \epsilon(\hat{\mathbf{r}}_2) \rangle / 4$, and $r_d = |\hat{\mathbf{r}}_1 - \hat{\mathbf{r}}_2|$.

Since the correlation distance for $B_n(r_d)$ is much smaller compared to the dimensions of the unit volume dV_c , we can simplify the double integral term in (13.4) as [12]

$$\int_{dV_c} \int_{dV_c} B_n(r_d) \exp[i\hat{\mathbf{k}}_s \cdot (\hat{\mathbf{r}}_1 - \hat{\mathbf{r}}_2)] dV_1 dV_2 = \int_{dV_c} dV_1 \int_{\infty} B_n(r_d) \exp[i\hat{\mathbf{k}}_s \cdot \hat{\mathbf{r}}_d] dV_2. \quad (13.5)$$

In this way, the inside integral in (13.5) becomes the spectral density of the turbulence which is given as

$$\Phi(k_s) = \left(\frac{1}{2\pi} \right)^3 \int_{\infty} B_n(r_d) \exp[i\hat{\mathbf{k}}_s \cdot \hat{\mathbf{r}}_d] dV. \quad (13.6)$$

By combining (13.4) and (13.6), the scattering cross-section is represented as

$$\sigma(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = \frac{2\pi k^4 \cos(\theta)^2}{dV_c} \int_{dV_c} \Phi(k_s) dV. \quad (13.7)$$

We model $\Phi(k_s)$ with the Kolmogorov spectrum technique which assumes the turbulence locally homogeneous [12]. Therefore, the spectrum characteristics becomes constant at the unit volume dV_c , and the scattering cross-section is further simplified as

$$\sigma(\hat{\mathbf{o}}, \hat{\mathbf{i}}) = 2\pi k^4 \cos(\theta)^2 \Phi(k_s). \quad (13.8)$$

In our model, we represent the scattering cross-section as a function of α, β and n because the directions are found with the angles in our approach. In the same way, the Kolmogorov spectrum is represented with $\Phi(\alpha, \beta, n)$.

Appendix B: Integral Bounds

In (4.8), the integral bound is given as the whole ring defined by the 10 dB beam-width. However, the lower part of the ring should be excluded from the integral as also shown in Figure 13.2. To this end, we define a function $\zeta(r, \theta)$ to exclude this area from the integral, and it is given as

$$\zeta(r, \theta) = \begin{cases} 0, & \text{if } r > r^{10dB} \cos(\pi/2 - \theta + \alpha_0), \text{ and} \\ & \pi - \alpha_0 + \varphi > \theta > \alpha_0 + \varphi \\ 1, & \text{otherwise} \end{cases} \quad (13.37)$$

where $\varphi = \arcsin(r^{3dB}/r^{10dB})$.

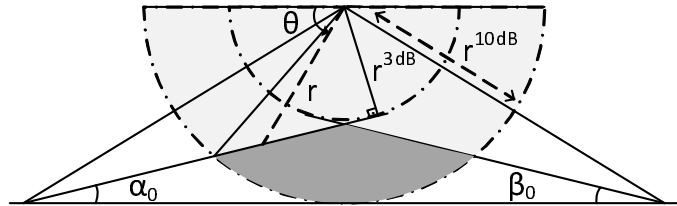


Figure 13.2: The integral boundaries.

Appendix C: Antenna Gains

Since RSM model utilize the 10 dB beam-width of the antennas, we include the antenna gains in the correlation calculations. To this end, we utilize Gaussian antenna pattern [6]. For the RSM method, we also need to represent the antenna gains as a function of r and θ . To this end, we utilize Figure 4.1 and geometrical approximations that are introduced in Section 4.3.3, thus the antenna gains is represented as

$$G_t(r, \theta) = e^{\left(-\frac{\left(\alpha_0 + \frac{w_t^{3dB}}{2} - \arctan\left(\frac{r \sin(\theta)}{R_t - r \cos(\theta)} \right) \right)^2}{(0.6 \times w_t^{3dB})^2} \right)} \quad (13.38)$$

$$G_r(r, \theta) = e^{\left(-\frac{\left(\beta_0 + \frac{w_t^{3dB}}{2} - \arctan\left(\frac{r \sin(\pi - \theta)}{R_t - r \cos(\pi - \theta)} \right) \right)^2}{(0.6 \times w_t^{3dB})^2} \right)} \quad (13.39)$$

Appendix D: $\xi(r, \theta)$ function

To simplify the correlation equations, we define $\xi(r, \theta)$ to represent the sine multiplication, and this function is given as

$$\begin{aligned} \xi(r, \theta) &= \sin(\Psi_1)^{-\frac{11}{3}} \sin(\Psi_2)^{-\frac{11}{3}}, \\ &= \left(\frac{\cos\left(\frac{\Psi_1 + \Psi_2}{2}\right) - \cos\left(\frac{\Psi_1 - \Psi_2}{2}\right)}{2} \right)^{-\frac{11}{3}}, \end{aligned} \quad (13.40)$$

Since the separation between antennas is very small in the troposcater communication between $70 - 100\lambda$, there will be a small difference between the scattering angles for the different antenna pairs in the space diversity. In frequency and angle diversity, the antennas are mounted on the same parabolic reflector therefore, the scattering angles will be same therefore, we assume $\Psi_1 = \Psi_2$. Thus, the ξ can be further simplified as

$$\xi(r, \theta) = [(\cos(\Psi_1/2 + \Psi_2/2) - 1)/2]^{(-11/3)}. \quad (13.41)$$

The scattering angles are calculated with the following equation for both of the antennas,

$$\begin{aligned} \Psi(r, \theta) = & \arctan \left(\frac{h - r \sin(\theta)}{d/2 - r \cos(\theta)} \right) + \\ & + \arctan \left(\frac{h - r \sin(\theta)}{d/2 + r \cos(\theta)} \right). \end{aligned} \quad (13.42)$$

Appendix E: Derivation of the Eikonal Equation

To derive (7.5), the earth radius transformation is utilized as shown in Figure 13.3 [100]. Therefore, the well-known Snell's equation for the stratified atmosphere is given as

$$R_0 n(z) \cos(\theta) = (R_0 + \Delta R) n(z + \Delta z) \cos(\theta + \Delta \theta). \quad (13.43)$$

Since the index of refraction is modeled with a bi-linear curve for the surface ducts, (13.43) is represented as

$$R_0 n(z) \cos(\theta) = (R_0 + \Delta R) \left(n(z) + \frac{dn}{dz} \Delta z \right) \cos(\theta + \Delta \theta). \quad (13.44)$$

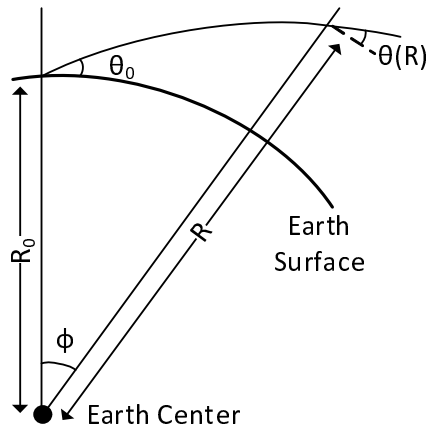


Figure 13.3: Ray trajectory for refractive path.

We know that Δz and $\Delta \theta$ will be small because the step sizes in the simulations will be low. Therefore, their squares and multiplications will give zero $\Delta z^2 = \Delta \theta^2 =$

$\Delta z \Delta \theta = 0$. For small $\Delta \theta$, $\cos(\Delta \theta) = 1$ and $\sin(\Delta \theta) = \Delta \theta$. By knowing these conditions, (13.44) is simplified as

$$\begin{aligned}
 R_0 n(z) \cos(\theta) &= R_0 n(z) \cos(\theta) \cos(\Delta \theta) + R_0 \frac{dn}{dz} \Delta z \cos(\theta) \\
 &\quad \cos(\Delta \theta) + \Delta R n(z) \cos(\theta) \cos(\Delta \theta) + \Delta R \frac{dn}{dz} \Delta z \cos(\theta) \cos(\Delta \theta) \\
 &\quad - R_0 n(z) \sin(\theta) \sin(\Delta \theta) - R_0 \frac{dn}{dz} \Delta z \sin(\theta) \sin(\Delta \theta) \\
 &\quad - \Delta R n(z) \sin(\theta) \sin(\Delta \theta) - \Delta R \frac{dn}{dz} \Delta z \sin(\theta) \sin(\Delta \theta). \quad (13.45)
 \end{aligned}$$

$$\begin{aligned}
 R_0 n(z) \sin(\theta) \Delta \theta &= R_0 \frac{dn}{dz} \Delta z \cos(\theta) + \Delta R n(z) \cos(\theta) \\
 \tan(\theta) \frac{\Delta \theta}{\Delta z} &= \frac{1}{n(z)} \frac{dn}{dz} + \frac{1}{R_0} \quad (13.46)
 \end{aligned}$$

where $\frac{\Delta \theta}{\Delta z} \tan(\theta) = \frac{\Delta \theta}{\Delta z} \frac{\Delta z}{\Delta x} = \frac{\Delta \theta}{\Delta x}$ [103]. As Δx goes to 0, $\frac{\Delta \theta}{\Delta x} = \frac{d\theta}{dx}$. Since $\theta = \frac{dz}{dx}$, the Eikonal equation for the Earth radius transformation can be represented as

$$\frac{d^2 z}{dx^2} = \frac{1}{n(z)} \frac{dn}{dz} + \frac{1}{R_0}. \quad (13.47)$$

Appendix F: Derivation of Ray Trajectory Equation

The ray trajectory equation (7.6) can be solved as

$$\frac{d^2 z}{dx^2} = \frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0}, \quad (13.48)$$

$$z'' = \frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0}, \quad (13.49)$$

$$z' = \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) x + \theta, \quad (13.50)$$

$$z = \frac{1}{2} \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right) x^2 + \theta x + h_t. \quad (13.51)$$

Appendix G: Derivation of Trapping Beam-width Derivations

In the trapping beam-width calculations, it is assumed that antennas has 0 elevation angle. The procedure is designed to find the maximum beam-width that can be

trapped by the surface ducts with respect to duct height (δ) and transmitter antenna height (h_t) as shown in Figure 13.4.

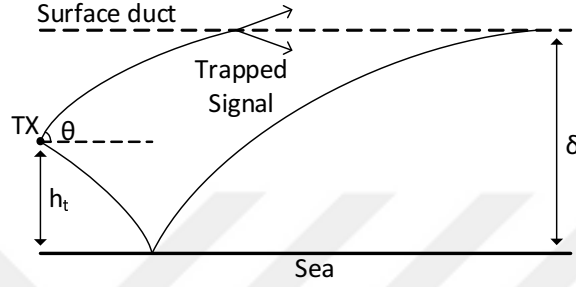


Figure 13.4: Trajectory of a trapped ray.

Starting from (7.7), let's denote $K = \frac{1}{2} \left(\frac{1}{n(0)} \frac{dn}{dz} + \frac{1}{R_0} \right)$ and we can write the following equations

$$z = Kx^2 + \theta x + h_t, \quad (13.52)$$

$$z + \Delta z = K(x + \Delta x)^2 + \theta(x + \Delta x) + h_t. \quad (13.53)$$

If we subtract the above equations, the result is given as

$$\Delta z = \Delta x^2 K + 2x \Delta x K + \theta \Delta x, \quad (13.54)$$

$$\frac{\Delta z}{\Delta x} = \Delta x K + 2x K + \theta. \quad (13.55)$$

At the boundary layer, the trapped ray will have zero Δz for very small step sizes because the trapped signal will first become parallel with the surface and then propagate downward. In addition, for small step sizes, $K \Delta x$ term will be nearly zero. Therefore, (13.55) is simplified as

$$-2xK = \theta^T, \quad (13.56)$$

$$x = -\frac{\theta^T}{2K}. \quad (13.57)$$

Therefore, at the duct height (boundary $z = \delta$), (13.52) is represented as

$$z = \delta = Kx^2 + \theta^T x + h_t. \quad (13.58)$$

By combining (13.57) and (13.58), the maximum and minimum effective trapping beam-widths are found as

$$\delta = \frac{1}{4K}\theta^{T^2} - \frac{1}{2K}\theta^{T^2} + h_t, \quad (13.59)$$

$$\delta - h_t = -\frac{1}{4K}\theta^{T^2}, \quad (13.60)$$

$$\theta^{T^2} = 4K(h_t - \delta), \quad (13.61)$$

$$\theta_{max,min}^T = \pm\sqrt{4K(h_t - \delta)}, \quad (13.62)$$

$$\theta_{max,min}^T = \pm\sqrt{2\left(\frac{1}{n(0)}\frac{dn}{dz} + \frac{1}{R_0}\right)(h_t - \delta)}. \quad (13.63)$$

Since the transmitter located within duct both $\frac{dn}{dz}$ and $h_t - \delta$ terms are negative, so this equation will have two real roots.

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