

THE CHARACTERIZATION OF PHASE CENTER OF MULTI-BAND GNSS
ANTENNA AND THE DESIGN OF V2V ANTENNA, ALONG WITH CHANNEL
ANALYSIS IN RURAL PROPAGATION SCENARIOS

by

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All the thanks are due to
God,
to my lovely and great wife
Mia,
my newborn daughter
Nancy,
my parents
Jianguo Liu and Min Gao
and my parents-in-law
Jinhu Gu and Xuezhen Ni

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Ran Liu

ABSTRACT

THE CHARACTERIZATION OF PHASE CENTER OF MULTI-BAND GNSS ANTENNA AND THE DESIGN OF V2V ANTENNA, ALONG WITH CHANNEL ANALYSIS IN RURAL PROPAGATION SCENARIOS

by

RAN LIU

Adviser: Daniel N. Aloia, Ph.D.

The wireless communication network has been vastly improved in the past decades. A high-precise positioning system and reliable wireless network has become an essential part of our daily life, especially Cellular Vehicle-to-Everything (C-V2X). It got most benefit development in Telematics Control Unit (TCU) of Vehicle E/E architecture. The TCU is main including the positioning and wireless connectivity. This research work focus on the automotive multi-band GNSS positioning antenna and Vehicle-to-Vehicle connectivity antenna.

High-precision Global Navigation Satellite System (GNSS) positioning is a promising technology that will be employed in future automotive navigation systems. This system will comprise the United States' Global Positioning System (GPS), the Russian Federation's Global Orbiting Navigation Satellite System (GLONASS), China Beidou Navigation Satellite System (BDS), and the European Union's Galileo. The significant enhancement in accuracy and precision is achieved through multiband signal transmission, carrier phase correction, and Real Time Kinematic (RTK). However, due to the current limitations of large-sized and expensive survey-grade antenna solutions, widespread adoption of this technology in the automotive sector remains challenging. In Chapter 3, it presents a cost-effective

small-sized multi-band ceramic GNSS patch antenna from its design stage to real sample fabrication. Furthermore, an investigation into this patch antenna reveals absolute phase center variation measured within an indoor range for accurate received signal phase error correction. Additionally, the study explores the integration of this low-cost antenna solution into a standard multi-band automotive antenna product. We evaluate this product both independently within an indoor range setting and on a typical vehicle roof at an outdoor range facility. Utilizing these evaluation findings enables us to achieve precise receiver position estimation with minimal phase motion errors.

Vehicle-to-Everything (V2X) communications will be an integral component of future autonomous drive decision systems, necessitating the establishment of a robust communication channel between end-to-end devices. However, existing cellular electromagnetic (EM) wave propagation models do not adequately suit V2X direct communication applications due to variations in antenna placement on vehicles. To address this issue, the Chapter 4 focuses on analyzing the propagation channel in a rural Vehicle-to-Vehicle (V2V) scenario for vehicular communication through antenna position experiments at different heights. By employing the ray-tracing algorithm, this research constructs a simulation model for a rural scenario using Commercial-off-the-Shelf (COTS) EM modeling software that calculates path loss received power and delay spread within a given propagation channel. Furthermore, it conducts real-world vehicle measurements to validate our simulation results, which demonstrate good agreement between simulated and measured receiver power. This study considers two types of antennas located at three different relative heights between the two vehicles and provides valuable insights into V2V antenna characteristics, antenna placement, and vehicle communication channel analysis.

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CHAPTER ONE

INTRODUCTION

On 2014, Society of Automotive Engineers (SAE) published first taxonomy and definitions for terms related to automation systems for on-road motor vehicle, Figure 1.1. High precision GNSS technology in C-V2X (Cellular Vehicle-to-Everything) is a system that utilizes advanced satellite positioning to provide precise location data for vehicles. This technology facilitates vehicle-to-vehicle (V2V), vehicle-to-infrastructure (V2I), vehicle-to-pedestrian (V2P), and vehicle-to-device (V2D) communication within the surrounding environment. By using high precision GNSS capabilities, C-V2X has the potential to improved safety measures, reduced traffic congestion, enhanced mobility, revenue growth, and environmental benefits.

Improved road safety: Through integrating C-V2X with high precision GNSS technology, vehicles can receive accurate and reliable information regarding their own position as well as nearby objects or other vehicles. Such communication aids in enhancing safety by alerting drivers about potential hazards like pedestrians or other vehicles present in their vicinity. According to recent statistics released by WHO in December 2018 [5], annual global road traffic fatalities have reached 1.35 million cases, which the leading cause of death among individuals aged 5-29 years old [5]. Road traffic injuries are also ranked as the eighth leading cause of death across all age groups [5]. In 2021, there were 42,915 fatalities resulting from motor vehicle crashes reported by USDOT (United States Department of Transportation) and NHTSA (National Highway Traffic Safety Administration), marking a sixteen-year peak [6]. An European study conducted in 2018 found that Advanced

Driver Assistance Systems (ADAS) could potentially reduce accidents by approximately 17% by year 2030 [7].

Reduced traffic congestion: C-V2X can help reduce traffic congestion by enabling vehicles to communicate with traffic signals and other infrastructure. This allows for more efficient traffic flow, as vehicles can be routed around congested areas or given priority at intersections. C-V2X can also help reduce the number of accidents. Urban mobility is complex, and in many cities, traffic is getting worse. From 2010 to 2016, congestion rose in London by 14%, in Los Angeles by 36%, in New York by 30%, and in Paris by 9% [8].

Enhanced mobility: By providing drivers with more accurate and reliable information about road conditions, traffic patterns, and alternative routes, C-V2X and high precision GNSS can improve mobility and enable new forms of transportation like ride-sharing and on-demand transportation. By enabling vehicles to communicate with infrastructure like traffic signals and road sensors, C-V2X and high precision GNSS can help improve the management of transportation infrastructure. This can include optimizing traffic flow, with seamless mobility, private cars and robotaxis could provide about 30% of passenger kilometers in 2030, compared with 40% for private cars today. Residents could “mix and match” rail transit and low-cost, point-to-point autonomous travel in robo-taxis and autonomous shuttles [9].

Revenue growth: By using C-V2X with high precise GNSS technology in Advanced Driver-Assistance Systems (ADAS) revenue growth, companies can develop and implement the technology, offer value-added services. By doing this, they can generate new revenue streams and enhance the customer experience. Vehicles with LiDAR-based products contain approximately \$1,500 to \$2,000 in component costs and even more for cars with SAE J3016 Level3 (L3) and L4 option

[2]. Based on consumer interest in Autonomous Driving (AD) features and commercial solutions available on the market today, ADAS and AD could generate between \$300 billion and \$400 billion in the passenger car market by 2035, according to McKinsey analysis [10].

Environmental benefits: C-V2X can help reduce emissions and improve air quality by enabling more efficient driving and reducing congestion. Drivers with real-time information about road conditions and traffic patterns, C-V2X can help them drive more smoothly and avoid unnecessary acceleration and braking. This can reduce fuel consumption. It could cost 25% to 35% less per trip, increase the number of point-to-point trips by 50%, and if autonomous vehicles are electric, reduce green house gas emissions by up to 85% [9].

	SAE Level 0	No Driving Automation	Zero autonomy; the driver performs all driving tasks
	SAE Level 1	Driver Assistance	Vehicle is controlled by the driver, but some driving assist features may be included in the vehicle design.
	SAE Level 2	Partial Driving Automation	Vehicle has combined automated functions, like acceleration and steering, but the driver must remain engaged with the driving task and monitor the environment at all times.
	SAE Level 3	Conditional Driving Automation	Driver is a necessity but is not required to monitor the environment. The driver must be ready to take control of the vehicle at all times with notice
	SAE Level 4	High Driving Automation	The vehicle is capable of performing all driving functions under certain conditions. The driver may have the option to control the vehicle.
	SAE Level 5	Full Driving Automation	The vehicle is capable of performing all driving functions under all conditions. The driver may have the option to control the vehicle.

Figure 1.1: Taxonomy and definitions for terms related to driving automation systems for on-road motor vehicles, SAE J3016 [2]

Table 1.1: Road safety number in the world and the USA

	Worldwide, 2018	USA, 2021
Road safety	1.35 million deaths	42,915 deaths
Environment	24% CO2 from vehicle	22% CO2 from vehicle
Traffic congestion	N/A	4.2 billion lost hours 2.8 billion gallons of wasted fuel

1.1 Problem Statement

The rapid advancement of the automotive industry and the widespread adoption of vehicles have significantly impacted individuals' lives and resulted in property damage due to traffic congestion and accidents. High precision GNSS technology in C-V2X (Cellular Vehicle-to-Everything) is a system that utilizes advanced satellite positioning to provide precise location data for vehicles. This technology is designed by Intelligent Transportation System (ITS), operates on a global frequency. The entire network is illustrated in Figure 1.2, which includes V2N, vehicle-to-vehicle (V2V), vehicle-to-infrastructure (V2I) and vehicle-to-pedestrian (V2P). V2N enables long-range communication through User-to-UTRAN (Uu) cellular evolved node B (eNB) base station to ITS platform. For the short-range direct communication, V2V, V2I and V2P can transmit low latency information using the direct communication PHY (PC5) side-link air interface.

What is the consideration for an automotive vehicle's antenna

- GNSS antenna - precise positioning
 - Receiver antenna motion error-> antenna performance
 - Automotive industrialization -> reliability and installation

-
- The diagram illustrates an ITS architecture. At the top, two satellites are shown. A central cloud represents the 'Network', which is connected to an 'ITS Platform' and an 'Internet' cloud. Below the network, a radio tower labeled 'eNB' is connected to two roadside units: 'RSU (eNB)' and 'RSU (UE)'. The ground level shows a road with several vehicles (green and white cars) and a pedestrian. Communication links are indicated by arrows: yellow lightning bolts for 'Uu' (satellite-to-ground), blue lightning bolts for 'PC5' (vehicle-to-vehicle), and black arrows for 'GNSS' (satellite-to-vehicle). Specific communication types are labeled: 'V2N' (vehicle-to-network), 'V2I' (vehicle-to-infrastructure), 'V2V' (vehicle-to-vehicle), and 'V2P' (vehicle-to-pedestrian).

1.2 Research Questions Proposal

5

- How can the Phase Center Variation (PCV) and Phase Center Offset (PCO) of the automotive receiver antenna affect the motion error and furthermore, impact GNSS carrier phase localization?
- Is there any low-cost method to acquire antenna phase calibration file and implement the calibration data into the GNSS TCU system?
- What is the challenge in vehicular network channel analysis?
- What is the V2V performance as function of antenna type and height?

1.3 Motivation and Challenges

There are some challenges in antenna analysis in the real automotive scenario. The motivation of this thesis including,

- a multi-band L1/L2/L5 GNSS antenna phase analysis in an ideal ground plane and an automotive roof location
- a GNSS system accuracy analysis in pseudorange and carrier phase range
- a V2V antenna performance analysis in three different heights of vehicle
- vehicular channel analysis in rural propagation scenario

1.4 Thesis Outline

The Telematic Control Unit (TCU) architecture depicted in Figure 1.3. Positioned at the core is the common stack, while peripheral circuitry houses functional devices and sensors, such as 4G/5G cellular connectivity, GNSS antenna signal input and V2X service.

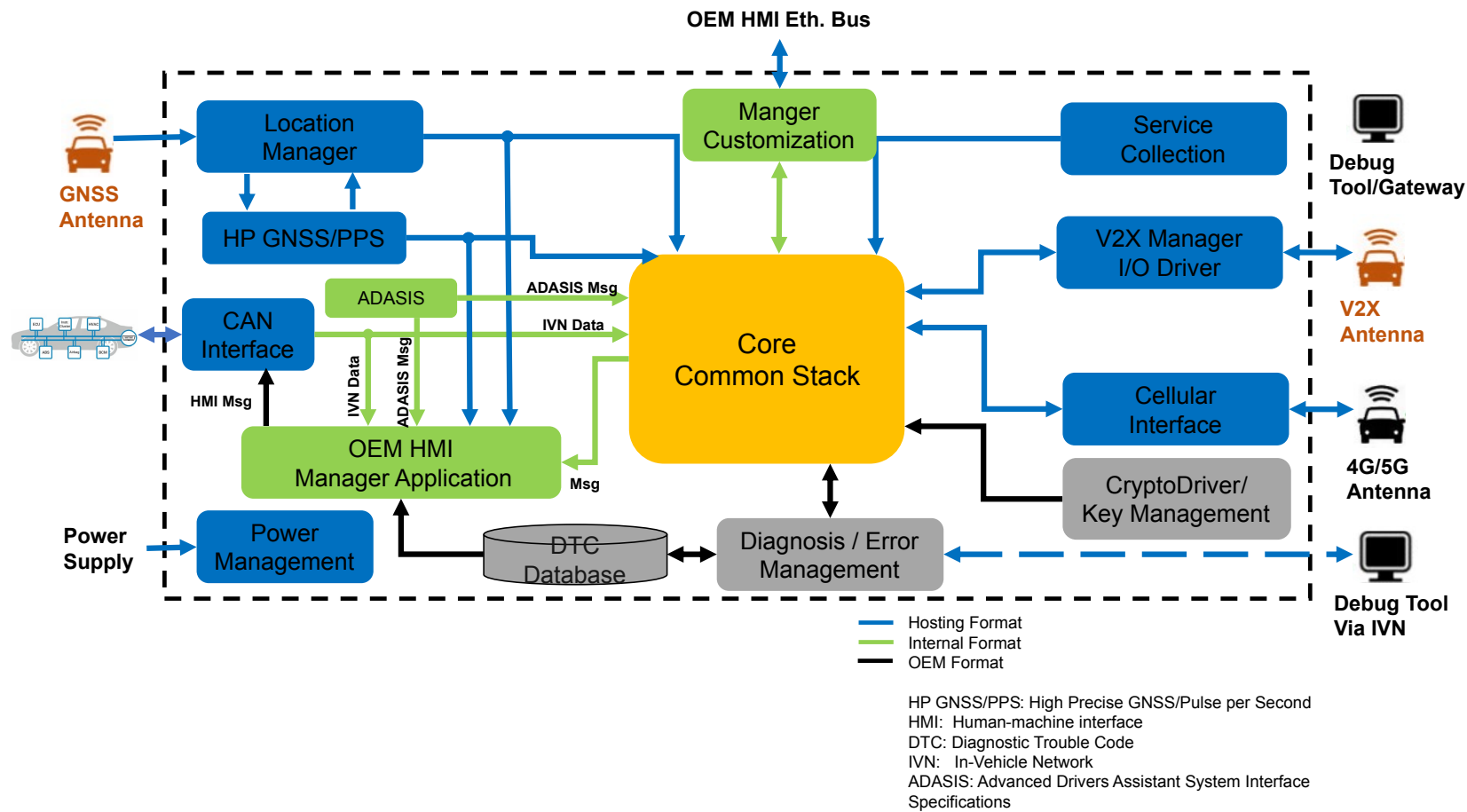


Figure 1.3: TCU architecture

This thesis primarily focuses on the GNSS antenna and V2X antenna for providing wireless access to the TCU. Chapter 2 presents a compact automotive GNSS antenna design along with its phase center characteristic. The design process of a low-cost small size multi-band ceramic GNSS patch antenna is described, leading to the development of an actual product. A further investigation of this patch antenna demonstrates the absolute phase center variation measured in an indoor range enabling received signal phase error correction. Furthermore, this low-cost antenna solution is examined when integrated into a standard multi-band automotive antenna product. The evaluation includes testing both indoor range and on a typical vehicle roof at an outdoor range. Utilizing this evaluation data allows for accurate estimation of receiver position by achieving motion error-free results through proper correction methods applied to raw GNSS data during static period outdoor testing.

Chapter 3 presents a state-of-the-art automotive V2V antenna, covering its design and real-world measurement. An automotive-grade antenna with small size and low-cost for mass production is introduced, along with an analysis of its performance and experiments on antenna positioning at various heights. A rural scenario simulation model is constructed using a Commercial-off-the-Shelf (COTS) electromagnetic modeling software package, employing the ray-tracing algorithm to compute path loss, received power, and delay spread in a given propagation channel. Furthermore, a real-world vehicle measurement campaign is conducted to validate the simulation results. The simulated and measured receiver power exhibited good agreement, considering two different types of antenna placed at three varying relative heights on the vehicles. This research provides valuable insights into V2V antenna characteristics, placement strategies, and analysis of vehicular communication channels.

CHAPTER TWO

FUNDAMENTALS

2.1 The Critical Scenarios for V2V Safety

Based on the Society of Automotive Engineers (SAE) J2945/1 On-Board System Requirements for V2V Safety Communications [3] and Chinese Society of Automotive Engineers (CSAE) Cooperative Intelligent Transportation System; Vehicular Communication: Application Layer Specification and Data Exchange Standard [11], the critical scenarios for V2V safety are listed in,

- Forward Crash Warning (FCW)
- Emergency Electronic Brake Lights (EEBL)
- Blind Spot Warning/Lane Change Warning (BSW/LCW)
- Intersection Movement Assist (IMA)
- Left Turn Assist (LTA)
- Control Loss Warning (CLW)

2.1.1 Forward Crash Warning (FCW)

The Remote Vehicle-1 (RV-1) is stopped on the road, the Host Vehicle (HV) is in the same lane by the RV-1, the HV receives a FCW signal to avoid rear-end crash with the stopped RV-1. The basic performance requirements,

- HV speed: 0-130 km/h
- Communication distance ≥ 300 m

- Data refresh frequency ≤ 10 Hz
- System latency ≤ 100 ms
- Position accuracy ≤ 1.5 m



Figure 2.1: FCW-stopped RV in the same lane [3]

2.1.2 Emergency Electronic Brake Lights (EEBL)

The Host Vehicle (HV) is driving on the road, HV follows by the Remote Vehicle-2 (RV-2). In front of RV-2, there is a RV-1 has a hard-breaking event with V2V capability. RV-2 may or may not have C-V2V communication, but HV could receive EEBL signal via non-LOS transmission. The basic performance requirements,

- HV speed: 0-130 km/h
- Communication distance ≥ 150 m
- Data refresh frequency ≤ 10 Hz

- System latency ≤ 100 ms
- Position accuracy ≤ 1.5 m

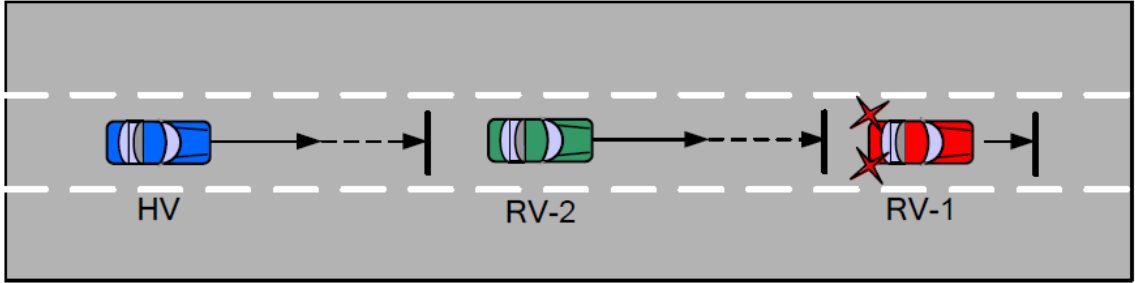


Figure 2.2: EEBL-abruptly slowing RV[3]

2.1.3 Blind Spot Warning/Lane Change Warning (BSW/LCW)

The HV is driving in the lane, and RV-1 is alongside the HV within the blind-spot zone. At this moment, the HV receives a BSW/LCW signal from RV-1 to warn about the occupied lane. Basic performance requirements remain unchanged,

- HV speed: 0-130 km/h
- Communication distance ≥ 150 m
- Data refresh frequency ≤ 10 Hz
- System latency ≤ 100 ms
- Position accuracy ≤ 1.5 m



Figure 2.3: BSW-RV in blind-spot zone [3]

2.1.4 Intersection Movement Assist (IMA)

- HV speed: 0-70 km/h
- Communication distance ≥ 150 m
- Data refresh frequency ≤ 10 Hz
- System latency ≤ 100 ms
- Position accuracy ≤ 5 m

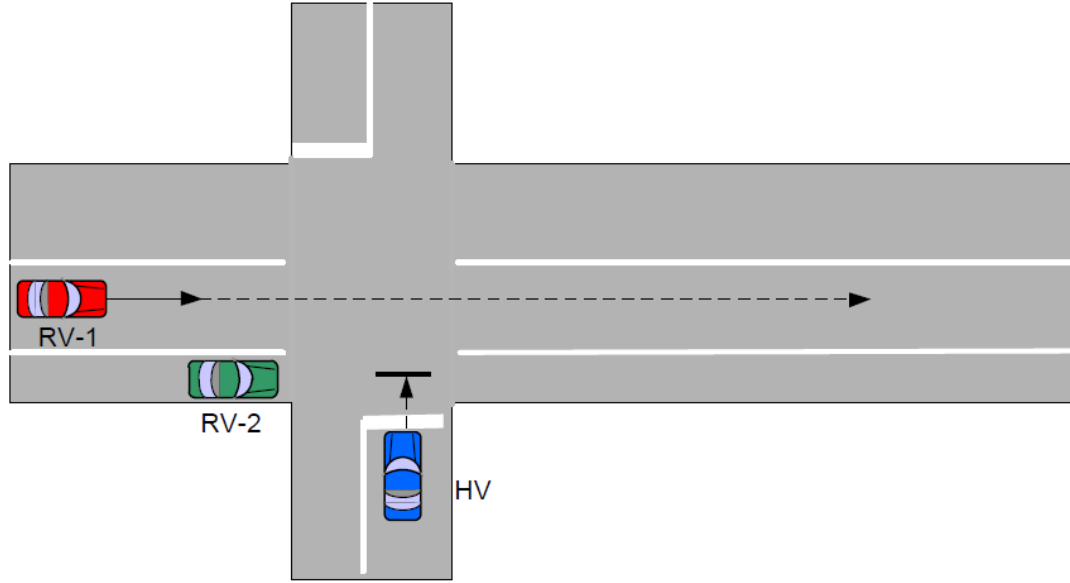


Figure 2.4: IMA-stopped HV at the intersection [3]

2.1.5 Left Turn Assist (LTA)

The HV is approaching an intersection to make a left turn. At this moment, RV-1 is also approaching the intersection with its view obstructed by RV-2. To avoid a potential crash with RV-1, the HV could receive LTA signals through non-line-of-sight transmission. These are the basic performance requirements,

- HV speed: 0-70 km/h
- Communication distance ≥ 150 m
- Data refresh frequency ≤ 10 Hz
- System latency ≤ 100 ms
- Position accuracy ≥ 5 m

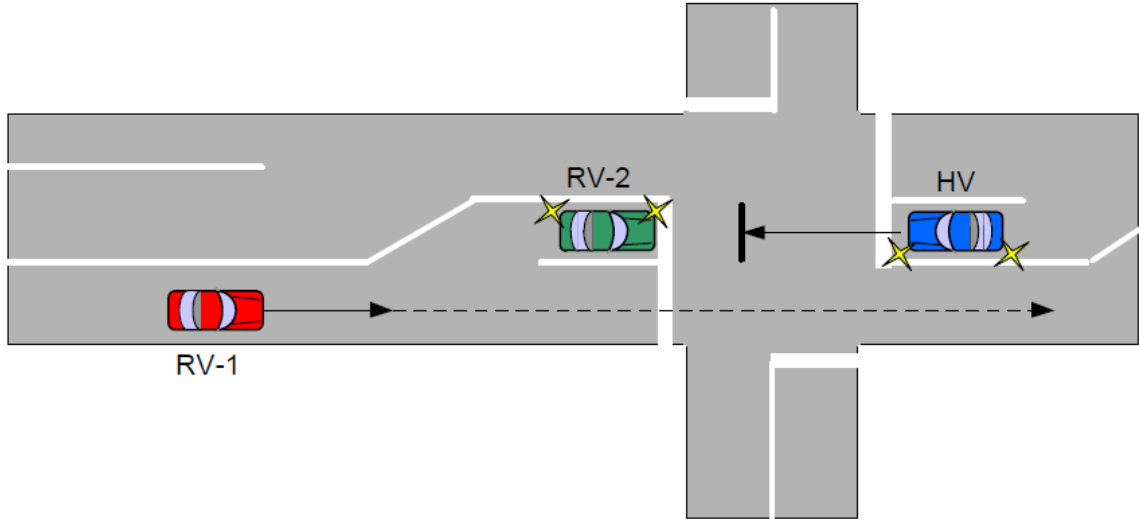


Figure 2.5: LTA-left turn across path [3]

2.1.6 Control Loss Warning (CLW)

The CLW safety application gives the driver warnings to HV an emergency control loss event, such as activation of the Antilock Brake System (ABS), traction control system and stability control system, by the RV in front of HV in the same or opposite direction. The basic performance requirements,

- HV speed: 0-130 km/h
- Communication distance ≥ 300 m
- Data refresh frequency ≤ 10 Hz
- System latency ≤ 100 ms
- Position accuracy ≤ 5 m

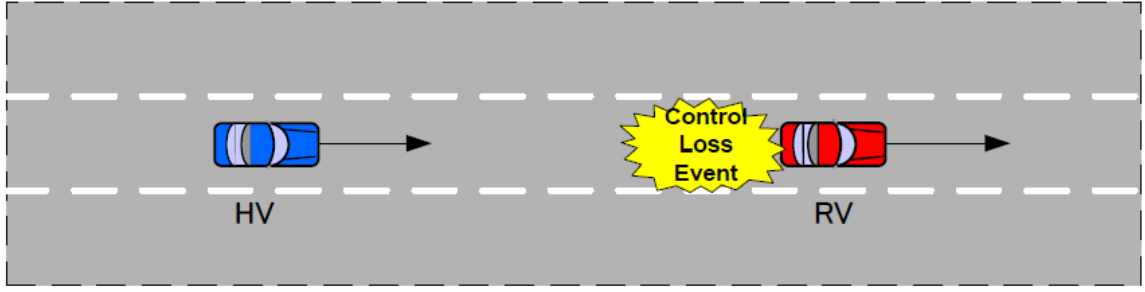


Figure 2.6: CLW-RV same direction of travel [3]

2.1.7 TCU General Stack

The general TCU stack is depicted in Figure 2.7, illustrating the V2X communication encompassing point-to-point singlecast, multicast and broadcast functionalities. From the bottom to the top, the stack comprises the physical layer, data link layer, network layer, transport layer, presentation layer and application layer.

In the physical layer (PHY), it encompasses vehicle interface and sensors access along with 4G/5G Uu and PC5 sidelink air-interface for V2X communication. This includes the Resource Blocks (RBs) with different subcarriers and time slots.

The Media Access Control (MAC) layer incorporates date filters, carrier/resource selection, logical channel priority sourcing as well as RBs distribution strategies.

Within the Radio Link Control (RLC) layer lies Service Data Units (SDU) distribution and reorganization methods for singlecast, multicast and broadcast scenarios. It facilitates Protocol Data Unit (PDU) transmission and receiving in the

Unacknowledged Mode (UM). In case of Sidelink Broadcast Control Channel (SBCCH) is in sidelink transmission, RLC supports Transparent Mode (TM).

In the layer of Packet Data Convergence Protocol (PDCP), it delivers the upper layer PDUs and duplicate the lower layer SDUs at PDCP and re-establish procedure for RLC Acknowledged Mode (AM).

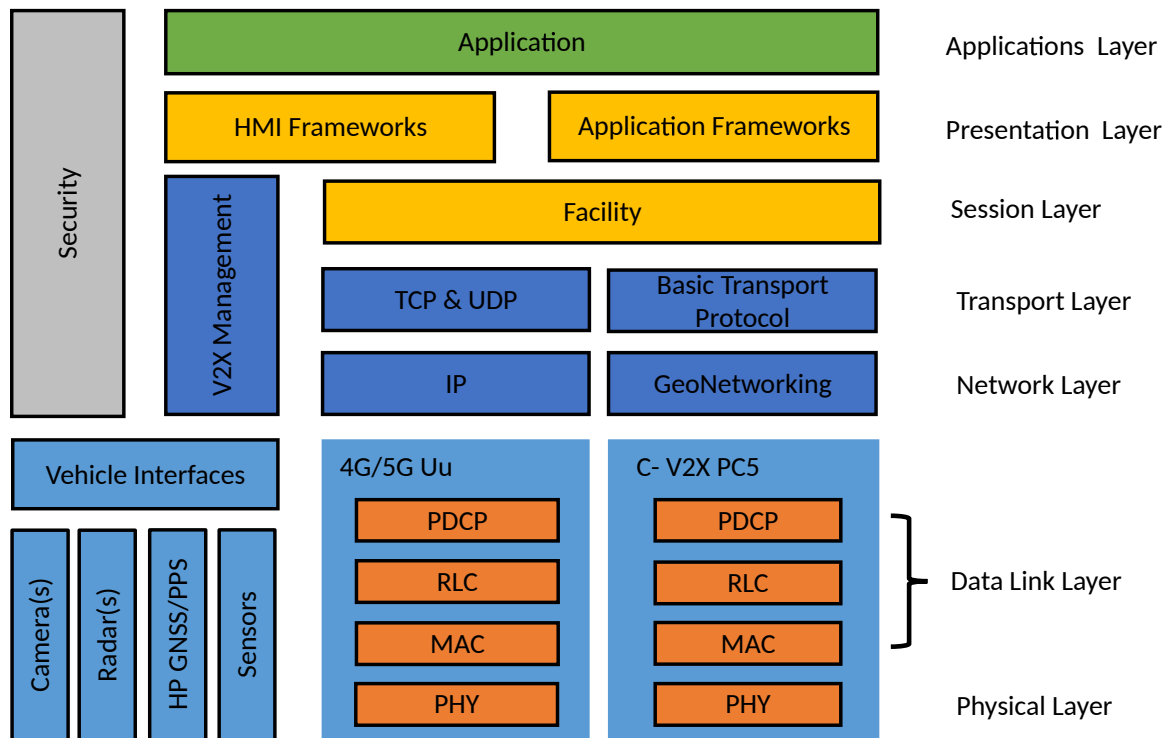


Figure 2.7: TCU general stack

2.2 Physical Layer Configurations

The publication of Band 47 frequency allocation by the Intelligent Transport System (ITS), Department of Transportation (DOT), and Ministry of Industry and Information Technology (MIIT) resulted in the worldwide release of a 70 MHz bandwidth for V2X services without requiring a license. Within this frequency spectrum, the low-end range of 5855-5875 MHz is designated for non-safety applications, while the high-end frequency band is allocated to road-vehicle safety services. Both the USA and China utilize Channel 183 (5905-5925 MHz) for C-V2X services. In China, V2V services operate within the range of 5905-5915 MHz, while V2I services are assigned to frequencies between 5915-5925 MHz [12] [13] [14] [15] [16] [17]. The basic C-V2X physical configurations are list in Table 2.1

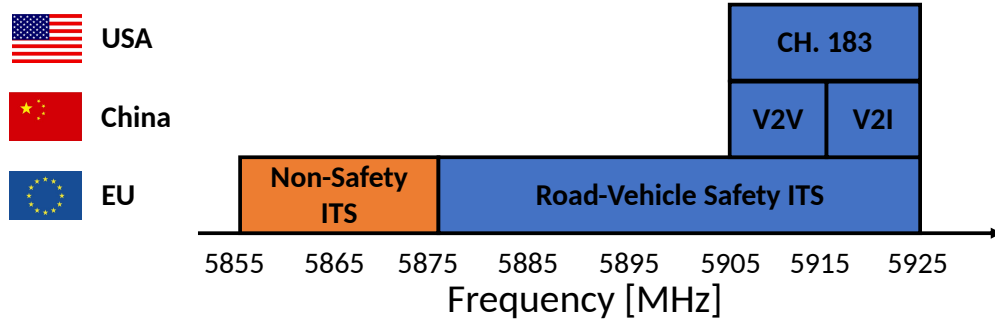


Figure 2.8: C-V2X frequency allocation

The Global Navigation Satellite System (GNSS) comprises the United States Global Positioning System (GPS), Russia's GLONASS, the European Union's Galileo, and China's Beidou. To simplify terminology, "L1" represents GPS Link 1

Table 2.1: C-V2X physical layer configurations

CH 183 Parameter	20 MHz	10 MHz
Sample Duration: T_S	0.05 us	0.1 us
Channel Spacing: $\Delta_F = 1/T_S$	20 MHz	10 MHz
Subcarrier Spacing: $\Delta_f = \Delta_F/64$	312.5 KHz	156.25 KHz
Orthogonality Duration: $T = 1/\Delta_f$	3.2 us	6.4 us
Cyclic Prefix (Guard Interval): $T_{GI} = T_S/4$	0.8 us	1.6 us
OFDM Symbol Duration: $T_{SYM} = T + T_{GI}$	4 us	8 us
MCS (QPSK, 1/2)	12 Mbit/s	6 Mbit/s
MCS (16-QAM, 1/2)	24 Mbit/s	12 Mbit/s

(1574.42 - 1576.42 MHz), Glonass G1 (1593 - 1607 MHz), and Beidou B1 (1559 - 1561 MHz). Similarly, "L2" denotes the GPS Link 2 band (1215 - 1237 MHz), while "L5" refers to the GPS Link 5 band (1164 - 1189 MHz). The frequency spectrum is shown on Figure 2.9 and the Table 2.2 presents a comprehensive overview of major civil system configurations.

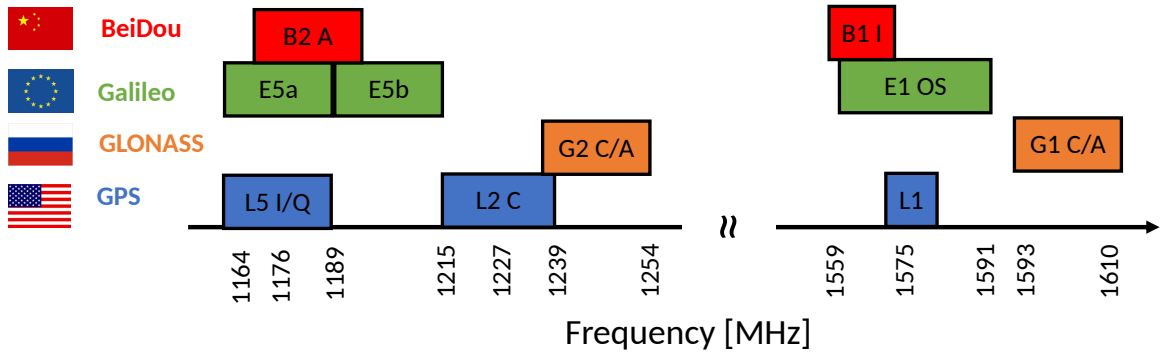


Figure 2.9: GNSS frequency allocation

Table 2.2: GNSS physical layer configurations

GNSS System	GPS L1 C/A	GPS L2C	GPS L5 I/Q	GLONASS G1 C/A	GLONASS G2 C/A	Galileo E1 OS	Galileo E5 OS	BeiDou B1I	BeiDou B2A
Center Freq. [MHz]	1575.42	1227.60	1176.45	1601.71	1245.78	1575.42	1191.795	1561.098	1176.45
Access Technology	CDMA	CDMA	CDMA	FDMA	FDMA	CDMA	CDMA	CDMA	CDMA
Modulation	BPSK (1)	BPSK (0.511)	BPSK (10)	BPSK (0.511)	BPSK (0.511)	CBOC (6,1,1/11)	AltBOC (15,10)	BPSK (4)	BPSK (10)
Code Freq. [MHz]	1.023	0.511	10.23	0.511	0.511	1.023	10.23	2.046	10.23
Code Family	Gold	M-seq.	M-seq.	M-seq.	M-seq.	Random	M-seq.	Gold	Gold
Min. Rx Power [dBw]	-158.5	-164.5	-157.9	-161	-167	-157	-155	-163	-156
Elevation Mask [°]	5	5	5	5	5	10	10	5	5

2.3 Wireless Propagation Channel

The performance of wireless communication is primarily constrained by the wireless propagation channel. The propagation path between the Transmitter (Tx) and Receiver (Rx) is highly complicated, ranging from the simplest scenario of free space line-of-sight (LOS) path loss to various complex terrain objects. Unlike cable communication, wireless communications exhibit extreme randomness and lack robustness and predictability. Particularly, the movement of mobile stations introduces dynamic fluctuations in signal levels. Modeling wireless channels poses a significant challenge in designing wireless systems, often relying on statistical methods for solutions.

The wireless propagation channel encompasses both large-scale and small-scale fading phenomena. Large-scale fading modeling estimates average signal strength over long distances (several hundred meters to several kilometers) between Tx and Rx, while small-scale fading modeling describes signal strength over short distances (a few electrical wavelengths)[18][19].

The large-scale fading includes:

- Free space path loss
- Reflection
- Transmission
- Diffraction
- Scattering

The small-scale fading includes:

- Multipath delay spread

- Flat fading
- Frequency selective fading
- Doppler spread
 - Fast fading
 - Slow fading

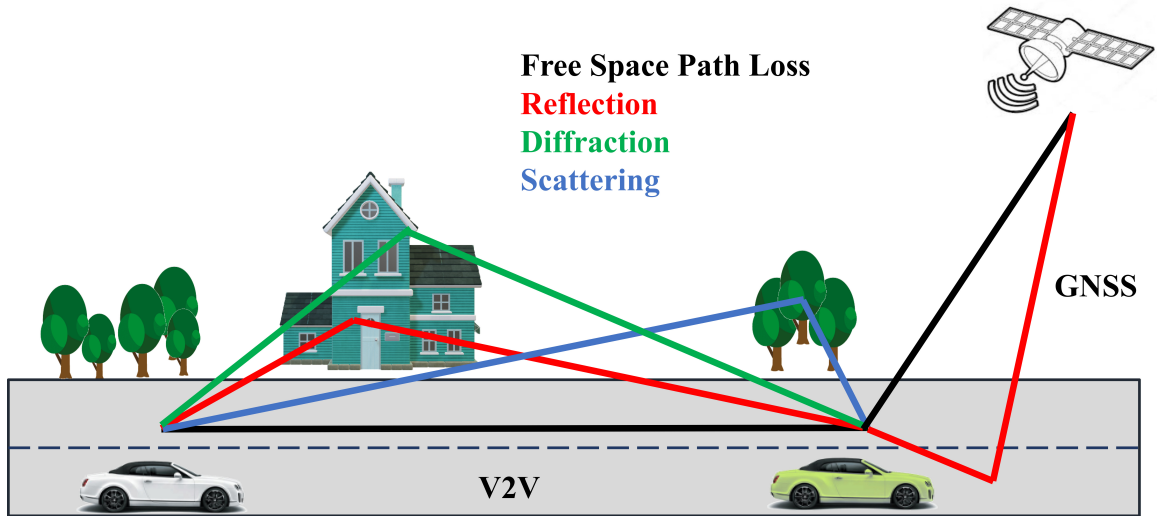


Figure 2.10: Wireless propagation channel

2.3.1 Large-Scale Fading

2.3.1.1 Free Space Path Loss A free space path loss model is shown in Figure 2.11. The basic transmission signal channel characteristic, defined as the transmit power P_t and received power P_r ratio in Eq.(2.1).

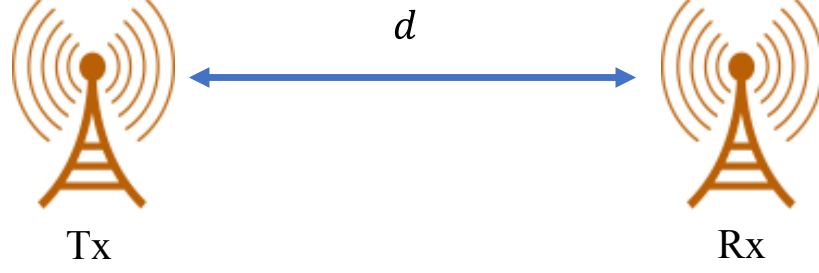


Figure 2.11: Free space path loss

$$PL(dB) = 10 \log \frac{P_r}{P_t} = -10 \log \left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2} \right] \quad (2.1)$$

where G_t is transmitter antenna's gain, G_r is receiver antenna's gain, λ is the wavelength, and d is the distance between transmitter and receiver.

2.3.1.2 Reflection and Transmission When the EM wave incidents on the boundary of two different dielectric material's surface, a part of the energy of wave will reflect, the other part of the energy will transmit. In general, EM wave is a kind of polarization wave, thus the electric field ($E - field$), magnetic field ($H - field$) and propagation direction are perpendicular to each other.

If the electric field strength of the incident wave is parallel to the plane of incidence wave, this wave is defined as parallel polarization wave (Figure 2.12 a); If the electric field strength of the incident wave is perpendicular to the plane of incidence wave, this wave is defined as orthogonal polarization wave (Figure 2.12 b).

(1). Incident wave reflection in parallel polarization:

The E-field of incident wave, reflected wave and transmission wave are

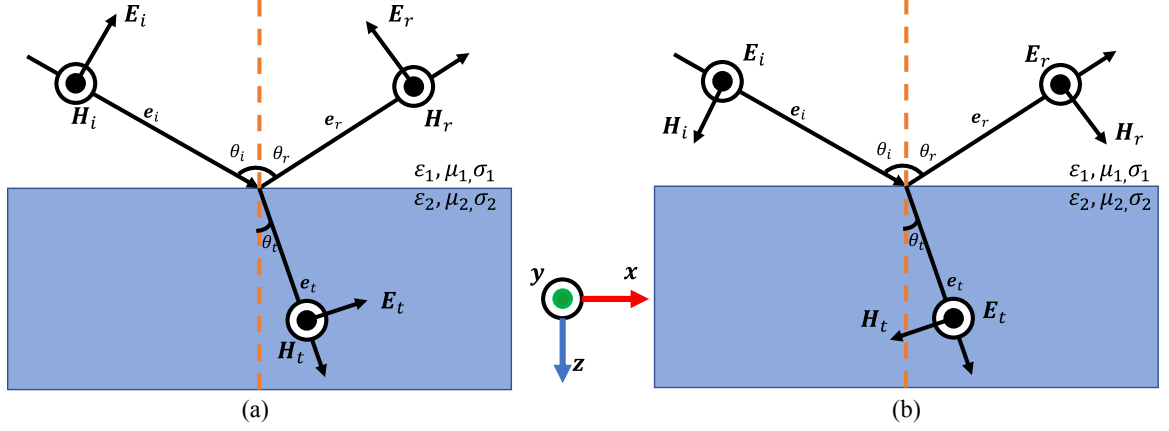


Figure 2.12: Incident wave reflection;
(a) Parallel polarization;
(b) Orthogonal polarization

$$\dot{\mathbf{E}}_i = (\mathbf{e}_x \cos \theta_i - \mathbf{e}_z \sin \theta_i) \dot{E}_i e^{-jk_1(x \sin \theta_i + z \cos \theta_i)} \quad (2.2a)$$

$$\dot{\mathbf{E}}_r = (-\mathbf{e}_x \cos \theta_r - \mathbf{e}_z \sin \theta_r) \dot{E}_r e^{-jk_1(x \sin \theta_r - z \cos \theta_r)} \quad (2.2b)$$

$$\dot{\mathbf{E}}_t = (\mathbf{e}_x \cos \theta_t - \mathbf{e}_z \sin \theta_t) \dot{E}_t e^{-jk_2(x \sin \theta_t + z \cos \theta_t)} \quad (2.2c)$$

On the boundary ($z = 0$), the electric field and magnetic field should meet the boundary condition:

$$\dot{E}_i \cos \theta_i - \dot{E}_r \cos \theta_r = \dot{E}_t \cos \theta_t \quad (2.3a)$$

$$\frac{\dot{E}_i}{\eta_1} + \frac{\dot{E}_r}{\eta_1} = \frac{\dot{E}_t}{\eta_2} \quad (2.3b)$$

where the η_i is the different material intrinsic impedance ($i = 1, 2$), and is given by $\sqrt{\mu_i/\varepsilon_i}$, the ration of electric to magnetic field for a uniform plane wave in the particular medium.

The incident wave reflection with parallel polarization in reflection factor

$$\Gamma_{\parallel} = \frac{\dot{E}_r}{\dot{E}_i} = \frac{\eta_1 \cos \theta_i - \eta_2 \cos \theta_t}{\eta_1 \cos \theta_i + \eta_2 \cos \theta_t} \quad (2.4a)$$

The incident wave reflection with parallel polarization in transmission factor

$$T_{\parallel} = \frac{\dot{E}_t}{\dot{E}_i} = \frac{2\eta_2 \cos \theta_i}{\eta_1 \cos \theta_i + \eta_2 \cos \theta_t} \quad (2.4b)$$

(2). Incident wave reflection in orthogonal polarization:

The E-field of incident wave, reflected wave and transmission wave are

$$\dot{\mathbf{E}}_i = \mathbf{e}_y \dot{E}_i e^{-jk_1(x \sin \theta_i + z \cos \theta_i)} \quad (2.5a)$$

$$\dot{\mathbf{E}}_r = \mathbf{e}_y \dot{E}_r e^{-jk_1(x \sin \theta_r - z \cos \theta_r)} \quad (2.5b)$$

$$\dot{\mathbf{E}}_t = \mathbf{e}_y \dot{E}_t e^{-jk_2(x \sin \theta_t + z \cos \theta_t)} \quad (2.5c)$$

On the boundary ($z = 0$), the E-field and H-field should meet the boundary condition:

$$\dot{E}_i + \dot{E}_r = \dot{E}_t \quad (2.6a)$$

$$-\frac{\dot{E}_i \cos \theta_i}{\eta_1} + \frac{\dot{E}_r \cos \theta_i}{\eta_1} = -\frac{\dot{E}_t \cos \theta_t}{\eta_2} \quad (2.6b)$$

The incident wave reflection with orthogonal polarization in reflection factor

$$\Gamma_{\perp} = \frac{\dot{E}_r}{\dot{E}_i} = \frac{\eta_2 \cos \theta_i - \eta_1 \cos \theta_t}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_t} \quad (2.7a)$$

The incident wave reflection with orthogonal polarization in transmission factor

$$T_{\perp} = \frac{\dot{E}_t}{\dot{E}_i} = \frac{2\eta_2 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_t} \quad (2.7b)$$

2.3.1.3 Two-Ray Ground Reflection Model In the most wireless communication scenarios, using the free space LOS path loss model is not accurate. At two-ray ground reflection model is based on the geometrical optics propagation modeling with direct and ground reflected path [19],

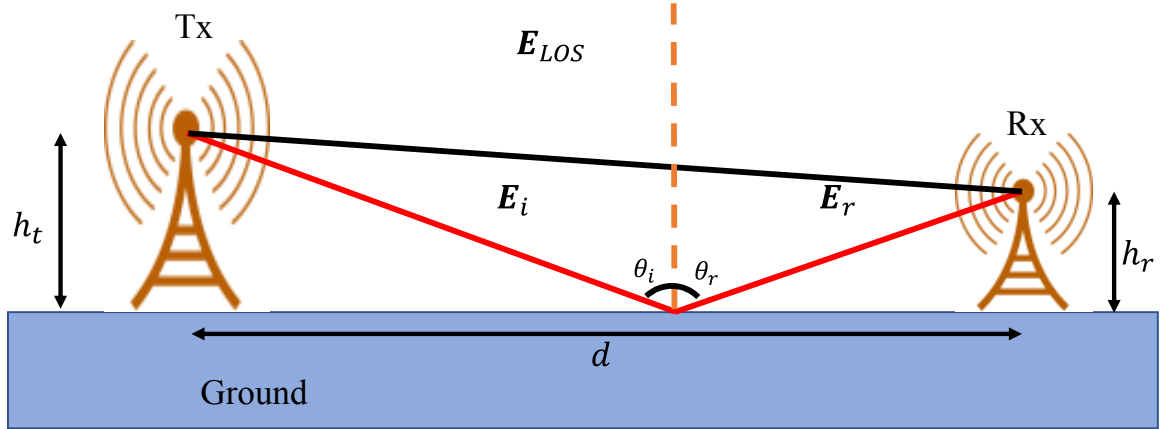


Figure 2.13: Two-ray ground reflection model

$$E_{LOS} = \frac{E_0 d_0}{d_{LOS}} \cos \left(\omega_c \left(t - \frac{d_{LOS}}{c} \right) \right) \quad (2.8)$$

where E_0 is the electric-field (in unit of v/m) from the distance d_0 of reference transmission antenna, and ω_c is carrier angular frequency.

$$d_{LOS} = \sqrt{(h_t - h_r)^2 + d^2} \quad (2.9)$$

where h_t is the height of transmission antenna, and h_r is the height of receiving antenna, d is the distance between transmitter and receiver,

$$E_r = E_g = (\Gamma) \frac{E_0 d_0}{d_r} \cos \left(\omega_c \left(t - \frac{d_r}{c} \right) \right) \quad (2.10)$$

where E_g is ground reflection electric-field, Γ refers to ideal electric surface reflection factor, and here is $\Gamma = -1$,

$$d_r = \sqrt{(h_t + h_r)^2 + d^2} \quad (2.11)$$

the total E-field is

$$\begin{aligned} E_{\text{total}}(d, t) &= E_{LOS} + E_g \\ &= \frac{E_0 d_0}{d_{LOS}} \cos \left(\omega_c \left(t - \frac{d_{LOS}}{c} \right) \right) + (-1) \frac{E_0 d_0}{d_r} \cos \left(\omega_c \left(t - \frac{d_r}{c} \right) \right) \end{aligned} \quad (2.12)$$

when $d \gg h_t + h_r$

$$E_{TOT}(d) \approx \frac{2E_0 d_0}{d} \times \frac{2\pi h_t h_r}{\lambda d} \quad (2.13)$$

the receiver power is,

$$P_r = P_t G_t G_r \frac{h_t^2 h_r^2}{d^4} \quad (2.14)$$

The first Fresnel zone is defined as an ellipsoid whose foci are the locations of the transmit and receive antennas. The distance from either antenna to a point on the ellipsoid and back to the other antenna is $\lambda/2$ greater than the direct path distance [20]. The distance d_f at which the first Fresnel zone becomes obstructed, is given by

$$d_f = \frac{1}{\lambda} \sqrt{(a^2 - b^2)^2 - 2(a^2 + b^2) \left(\frac{\lambda}{2}\right)^2 + \left(\frac{\lambda}{2}\right)^4} \quad (2.15)$$

where $a = h_t + h_r$ and $b = h_t - h_r$.

2.3.1.4 Transmission through a Dielectric Substrate When the EM wave transmit through a dielectric substrate, such as windshield glass, the incident wave is hit on a point of surface (initial hits), part of the waves is transmitted, other parts are reflected then re-transmitted [21]. As shown in Figure 2.14

2.3.1.5 Diffraction When a large obstacle is encountered in the propagation path, the EM wave will travel forward around the obstacle, a phenomenon called diffraction. The signal quality is affected is not only related to the height of the obstacle, and the distance between Tx and Rx, but also related to the frequency of the EM wave. The wave has weaker diffraction capabilities as its higher frequency and shorter wavelength. Diffraction losses are commonly modeled by knife edge diffraction method based on the Fresnel-Kirchoff diffraction parameter.

2.3.1.6 Scattering In the real world wireless communication scenario, the received signal strength is stronger than the reflection and diffraction models, while the EM wave is facing the rough surface, the energy of the reflected signal spreads in all the directions due to scattering, such as lamp posts and trees. Thus it will provide extra

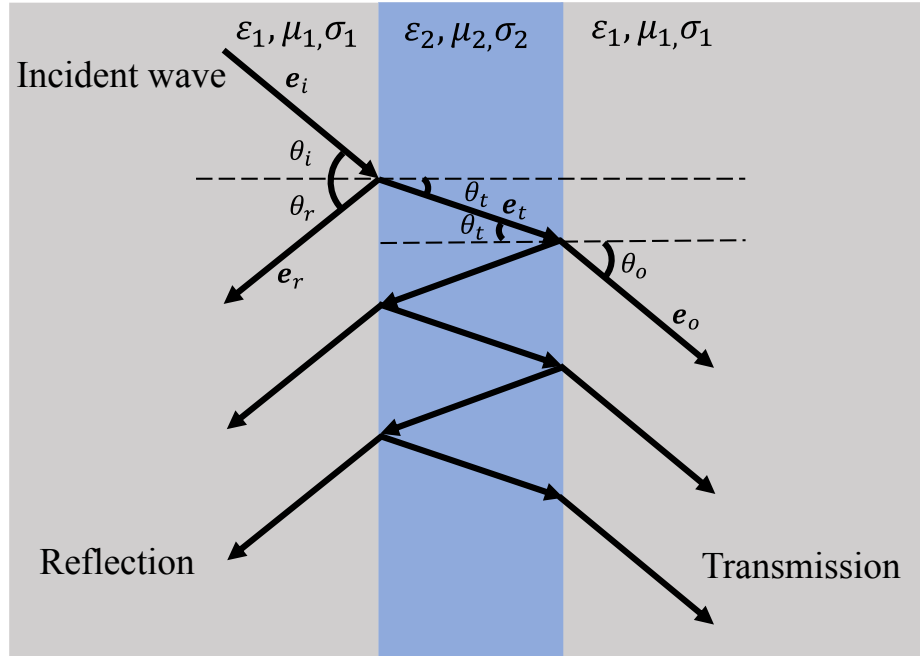


Figure 2.14: Transmission through a dielectric substrate

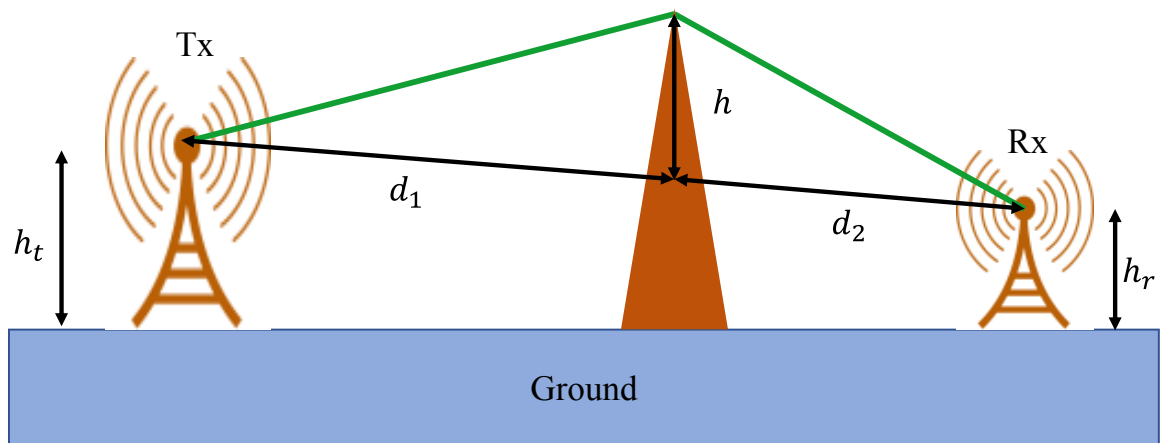


Figure 2.15: Diffraction as knife edge model

energy to the received signal strength. When the incident wavelength is much longer than the smooth surface's dimension, it modeled as reflective surfaces, while the roughness of surfaces often occurs different propagation effects, as shown in Figure 2.16.

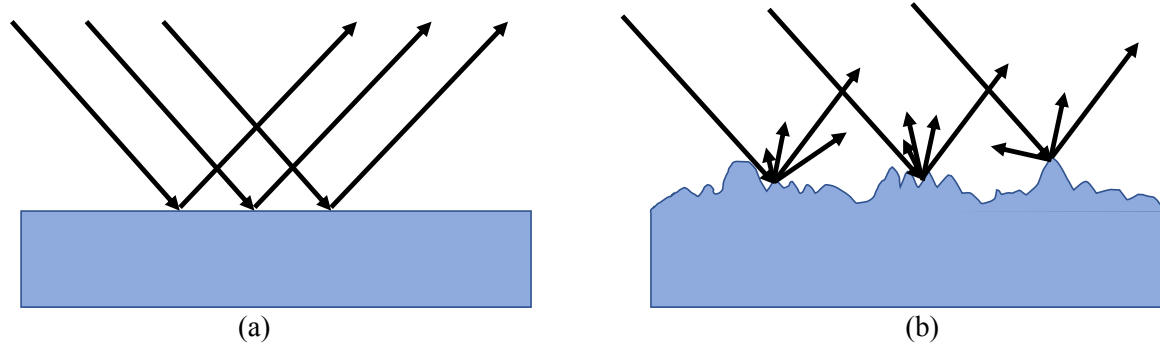


Figure 2.16: Scattering;
(a) Smooth surface;
(b) Rough surface

2.3.2 Small-Scale Fading

The small-scale fading occurs when the ray propagation has reflection, diffraction and transmission. The signal level in a small-scale path has a significant change, which could simultaneously bring an impact on the receiving power sensitivity of wireless communication.

2.3.2.1 Multipath Delay Spread A common method to evaluate the multipath propagation is using time dispersion parameter from a power delay profile, such as mean excess delay and Root-Mean-Square (RMS) delay spread.

The mean excess delay is the first moment of the power delay profile and is defined to be [19],

$$\bar{\tau} = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)} \quad (2.16)$$

a_k is the k arrival multipath signal amplitude, τ is the excess delay bins, and $P(\tau_k)\tau_k$ is the k received power delay profile.

The Root-Mean-Square (RMS) Delay Spread is defined as the square root of the second central moment of the power delay profile

$$\sigma_\tau = \sqrt{\overline{\tau^2} - (\bar{\tau})^2} \quad (2.17)$$

where

$$\overline{\tau^2} = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k) \tau_k^2}{\sum_k P(\tau_k)}$$

The coherence bandwidth could be set as correlation function above 0.9 [22]

$$B_c \approx \frac{1}{50\sigma_\tau} \quad (2.18)$$

- Flat Fading: If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of transmitted signal $B_c \gg B_s$, as shown on Figure 2.17(a).
- Frequency Selective Fading: If the channel possesses a constant-gain and linear phase response over a bandwidth that is smaller than the bandwidth of transmitted signal $B_c \ll B_s$, as shown on Figure 2.17(b).

2.3.2.2 Doppler Spread Doppler Spread and coherence time are parameters which describe the time varying of the channel. *Doppler Spread B_D is a measure of the*

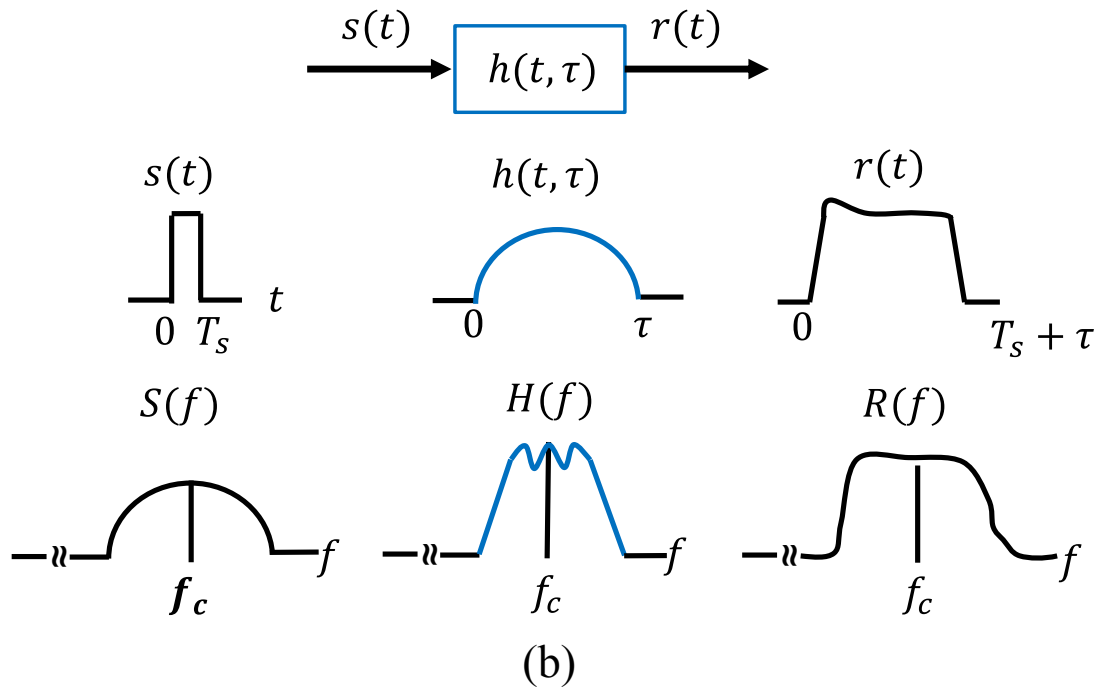
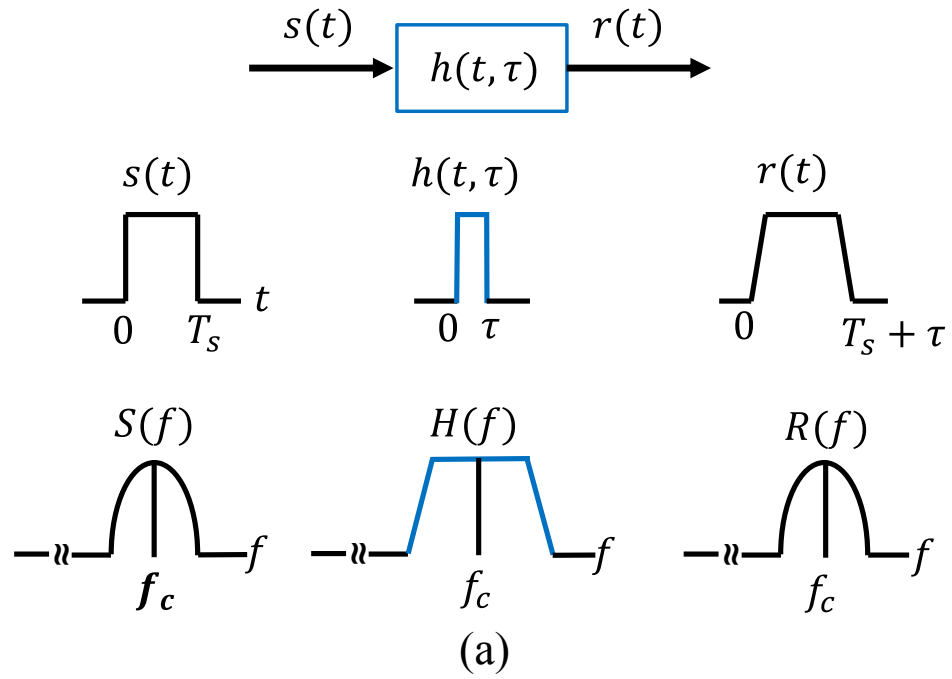


Figure 2.17: Multipath channel characteristics;
(a) Flat fading $B_S \ll B_C$ and $T_S \gg \sigma_\tau$;
(b) Frequency selective fading $B_S > B_C$ and $T_S < \sigma_\tau$

spectral broadening caused by the time rate of change of the mobile radio channel and is defined as the range of frequencies over which the received Doppler Spectrum is essentially non-zero[19]. Coherence time T_C is the time domain of Doppler Spread and is used to characterize the time varying nature of the frequency dispersiveness of the channel in the time domain. The Doppler Spread and coherence time are inversely proportional to one another . That is

$$T_C \approx \frac{1}{f_D} \quad (2.19)$$

where $f_D = \frac{\lambda}{v}$, λ is wave length and v is the relative speed of the rover.

If the coherence time is defined as the time over which the time correlation function is above 0.5, then the coherence time is [23]

$$T_C \approx \frac{9}{16\pi f_D} \quad (2.20)$$

2.3.2.3 Small-Scale Fading Summary The summary of the above section discussed the small-scale fading in wireless communication in Table 2.3. Depending on the relation between the signal parameters (such as bandwidth, symbol period) and the channel parameters (such as RMS delay spread and Doppler spread), different transmitted signals will undergo different type of fading. The time dispersion and frequency dispersion mechanisms in a mobile radio channel lead to four possible distinct effects, which are manifested depending on the nature of the transmitted signal, the channel, and the velocity. While multipath delay spread leads to *flat fading* and *frequency selective fading*, it depends on the relationship of B_S and B_C . Doppler spread leads to *fast fading* and *slow fading*, which is depended on the B_S and B_C [19].

Table 2.3: Small-Scale Fading Summary

Multipath Delay Spread (Time Dispersion)	Doppler Spread (Frequency Dispersion)
Flat Fading: $B_S \ll B_C$ and $T_S \gg \sigma_\tau$	Fast Fading: $T_S > T_C$ and $B_S \ll B_D$
Frequency Selective Fading: $B_S > B_C$ and $T_S < \sigma_\tau$	Slow Fading: $T_S \ll T_C$ and $B_S \gg B_D$

where B_S : symbol bandwidth,

T_S : symbol period,

B_C : coherence bandwidth,

σ_τ : RMS delay spread;

B_D : Doppler spread bandwidth,

T_C : Coherence time.

2.4 GNSS Measurement Models

The principle of satellite navigation is the satellite-user range measurements based on the times of transmission and receipt of signal, which includes two major methods: code Eq.(2.21) and carrier phase Eq.(2.22) measurement as illustrated in Figure 2.18 of [4]. The carrier phase measurement adds the carrier wave and integer ambiguity, the rest of equation part is the same as the code pseudorange. The reason the carrier phase could significantly improve the position accuracy and precision is that measurement resolution is much finer than code measurement. The L1 C/A code chipping rate is 1.023 MHz, the corresponding wavelength is around 300 m, while the carrier phase of L1 is 1575.42 MHz, and the corresponding

wavelength is around 0.19 m. By analogy with the wavelength as a measuring tape, carrier phase measurement has a fine tick mark [4].

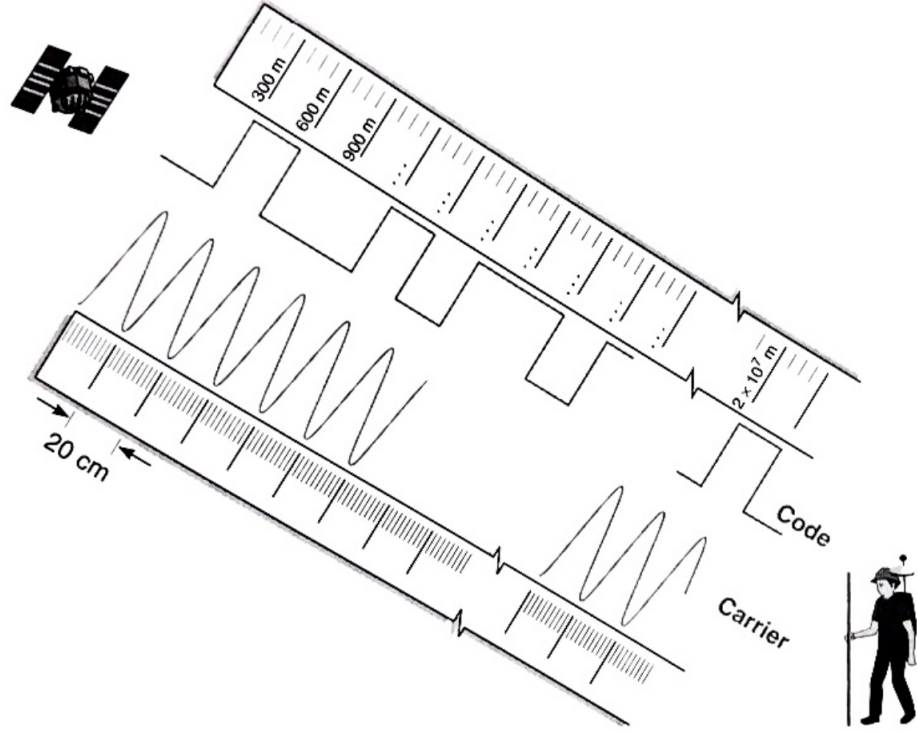


Figure 2.18: A conceptual representation of the code and carrier phase measurement with a tape measurement [4]

On the other hand, the code pseudorange period is known as a digital signal, thus ρ is an easy acquisition; while the carrier phase has difficulty finding the transmission wave period of integer ambiguity N ,

$$\rho = d + I_\rho + T_\rho + c[\delta t_u - \delta t^s] + \varepsilon_\rho \quad (2.21)$$

$$\varphi = \frac{1}{\lambda}[d + I_\rho + T_\rho] + \frac{c}{\lambda}(\delta t_u - \delta t^s) + N + \varepsilon_\varphi \quad (2.22)$$

where ρ represents the code pseudorange, d denotes the true range distance between the user and satellite, I_ρ is ionospheric pseudorange delay, T_ρ is tropospheric delay, c is the speed of the light in vacuum, λ is transmission signal wavelength (unit: meter), I_ρ and I_φ are equal in magnitude but opposite in sign ($I_\rho = -I_\varphi$), N is the period integer ambiguity, ε is the measurement error.

Real-Time Kinematic (RTK) is a technology that enables high accuracy and precision positioning by resolving various unknowns as follows;

- 1) Single difference eliminates the satellite clock biases $\delta t^{(k)}$, ionospheric delay I and tropospheric delay T ;
- 2) Double difference eliminates clock bias in reference δt_r and user δt_u ;
- 3) Triple difference eliminates the integer ambiguity N . The principle of satellite navigation by using RTK carrier phase measurement is shown in Figure 2.19.

2.4.1 Single Difference

Single-difference is using one reference station, as shown in Figure 2.20, φ_r : Eq.(2.23a) and φ_u : Eq.(2.23b) to measure a same satellite (k) at the Eq.(2.23b – 2.23a). This is an idealized carrier phase measurement setup.

$$\varphi_r^{(k)} = \frac{1}{\lambda}[d_r^{(k)} + I_r^{(k)} + T_r^{(k)}] + \frac{c}{\lambda}(\delta t_r - \delta t^{(k)}) + N_r^{(k)} + \varepsilon_r^{(k)} \quad (2.23a)$$

$$\varphi_u^{(k)} = \frac{1}{\lambda}[d_u^{(k)} + I_u^{(k)} + T_u^{(k)}] + \frac{c}{\lambda}(\delta t_u - \delta t^{(k)}) + N_u^{(k)} + \varepsilon_u^{(k)} \quad (2.23b)$$

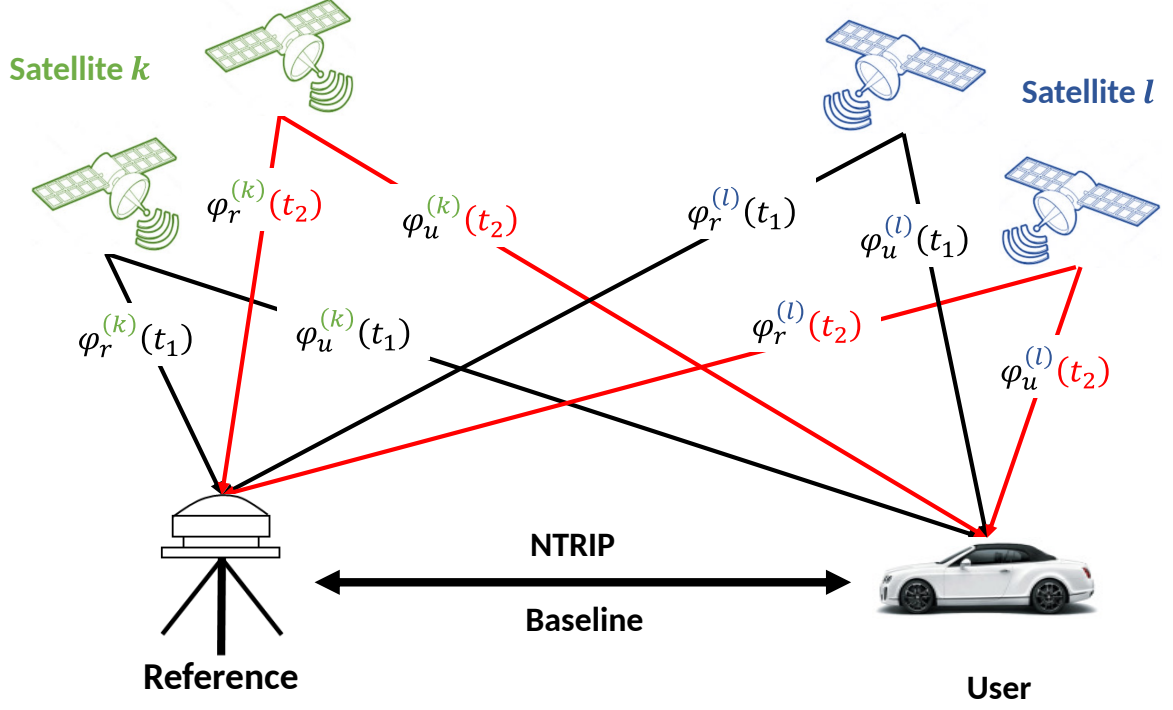


Figure 2.19: The principle of satellite navigation by using carrier phase measurement: including single, double and triple difference.

$$\begin{aligned}
 \Delta\varphi_{ur}^{(k)} \stackrel{3.1b-3.1a}{\implies} \varphi_u^{(k)} - \varphi_r^{(k)} &= \frac{1}{\lambda} \left[\left(d_u^{(k)} - d_r^{(k)} \right) + \left(I_u^{(k)} - I_r^{(k)} \right) + \left(T_u^{(k)} - T_r^{(k)} \right) \right] \\
 &\quad + \frac{c}{\lambda} (\delta t_u - \delta t_r) + \left(N_u^{(k)} - N_r^{(k)} \right) + \left(\varepsilon_u^{(k)} - \varepsilon_r^{(k)} \right)
 \end{aligned} \tag{2.24}$$

If the reference station is not far away from the user rover (less than 20 km), the ionospheric delay I and tropospheric delay T have strong correlation, their delay could be eliminated as well

$$\Delta\varphi_{ur}^{(k)} = \frac{1}{\lambda} d_{ur}^{(k)} + \frac{c}{\lambda} (\delta t_{ur}) + N_{ur}^{(k)} + \varepsilon_{ur}^{(k)} \tag{2.25}$$

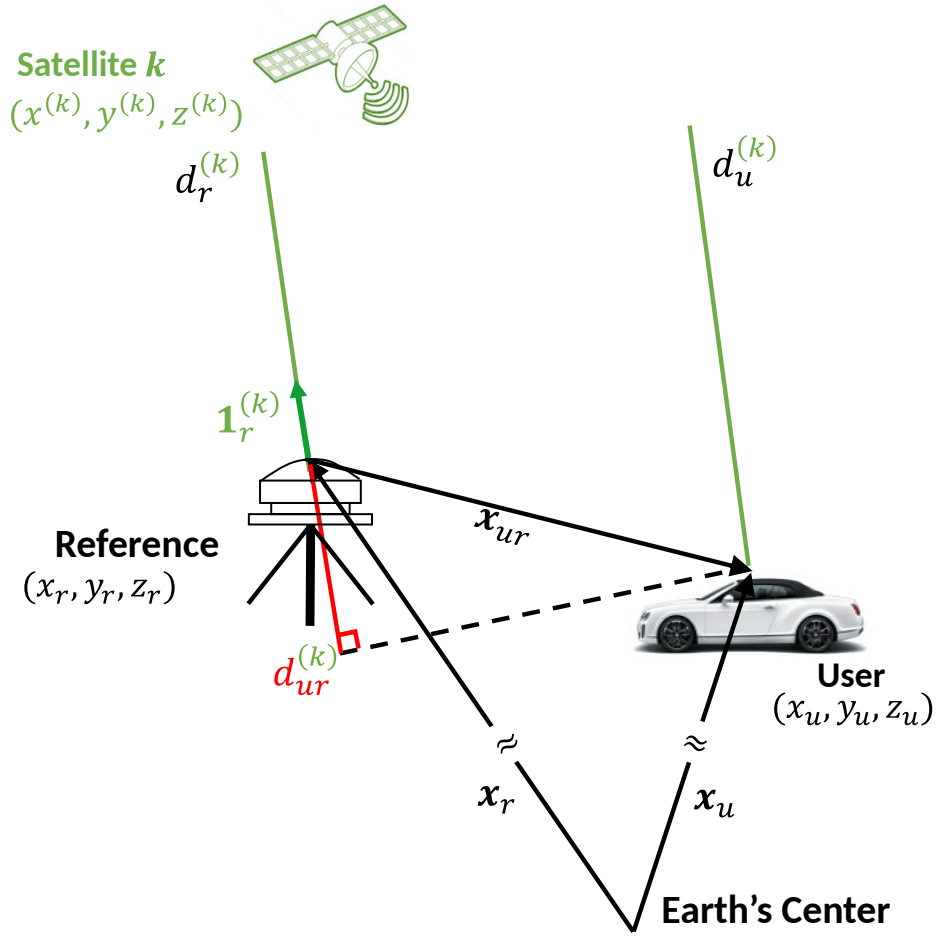


Figure 2.20: Geometry of the single-difference measurement

Solving single difference ambiguity N from Eq.(2.25) by linear model. In Earth Center Earth Fixed (ECEF) coordinating system

$$d_{ur}^{(k)} = d_u^{(k)} - d_r^{(k)} \quad (2.26)$$

where

$$d_u^{(k)} = \sqrt{\left(x^{(k)} - x_u\right)^2 + \left(y^{(k)} - y_u\right)^2 + \left(z^{(k)} - z_u\right)^2} = \left\|\mathbf{x}^{(k)} - \mathbf{x}_u\right\| \quad (2.27a)$$

$$d_r^{(k)} = \sqrt{\left(x^{(k)} - x_r\right)^2 + \left(y^{(k)} - y_r\right)^2 + \left(z^{(k)} - z_r\right)^2} = \left\|\mathbf{x}^{(k)} - \mathbf{x}_r\right\| \quad (2.27b)$$

K from Eq.(2.26) could be solved by linear model for position estimation. This idea is referred to *Newton – Raphson method*, which start with a rough estimate of the user position, then refine the estimation iteratively.

Let $\mathbf{x}^{(k)}$ is satellite k position, $\mathbf{x}_0 = (x_0, y_0, z_0)$ is the first guess of the user position, $d_0^{(k)}$ is corresponding approximation based on the first trail $\mathbf{x}_0$,

$$d_0^{(k)} = \left\|\mathbf{x}^{(k)} - \mathbf{x}_0\right\| \quad (2.28)$$

the true position $\mathbf{x}_u = (x_u, y_u, z_u)$ would be represented as

$$\mathbf{x}_u = \mathbf{x}_0 + \delta\mathbf{x} \quad (2.29)$$

where $\delta\mathbf{x}$ in unknown corrections to be applied to the initial estimates.

$$\delta d_u^{(k)} = d_u^{(k)} - d_0^{(k)} = \left\|\mathbf{x}^{(k)} - \mathbf{x}_0 - \delta\mathbf{x}\right\| - \left\|\mathbf{x}^{(k)} - \mathbf{x}_0\right\| \approx -\frac{\left(\mathbf{x}^{(k)} - \mathbf{x}_0\right)}{\left\|\mathbf{x}^{(k)} - \mathbf{x}_0\right\|} = -\mathbf{1}_0^{(k)}\delta\mathbf{x} \quad (2.30)$$

$\mathbf{1}_0^{(k)}$ is the estimated line-of-sight unit vector directed from the initial estimate of the user position to satellite k

$$\mathbf{1}_0^{(k)} = \frac{1}{\left\|\mathbf{x}^{(k)} - \mathbf{x}_0\right\|} \left(x^{(k)} - x_0, y^{(k)} - y_0, z^{(k)} - z_0\right)^T \quad (2.31)$$

the iteration is stopped by met correction criteria.

For the single difference, it takes the reference station position

$\mathbf{x}_r = (x_r, y_r, z_r)$, taking Eq.(2.30) into Eq.(2.26)

$$d_{ur}^{(k)} = d_u^{(k)} - d_r^{(k)} = -\mathbf{1}_r^{(k)} \cdot \mathbf{x}_{ur} \quad (2.32)$$

the Eq.(2.25) can be written as

$$\Delta\varphi_{ur}^{(k)} = \frac{1}{\lambda}d_{ur}^{(k)} + \frac{c}{\lambda}(\delta t_{ur}) + N_{ur}^{(k)} + \varepsilon_{ur}^{(k)} = \frac{1}{\lambda} \left(-\mathbf{1}_r^{(k)} \right) \cdot \mathbf{x}_{ur} + \frac{c}{\lambda}(\delta t_{ur}) + N_{ur}^{(k)} + \varepsilon_{ur}^{(k)} \quad (2.33a)$$

in vector-matrix notation as below,

$$\Delta\varphi_{ur}^{(k)} = \frac{1}{\lambda} \begin{bmatrix} \left(-\mathbf{1}_r^{(1)} \right)^T \\ \left(-\mathbf{1}_r^{(2)} \right)^T \\ \vdots \\ \left(-\mathbf{1}_r^{(k)} \right)^T \end{bmatrix} \cdot \mathbf{x}_{ur} + \frac{c}{\lambda}(\delta t_{ur}) \begin{bmatrix} 1 \\ 1 \\ \vdots \\ 1 \end{bmatrix} + \begin{bmatrix} N_{ur}^{(1)} \\ N_{ur}^{(2)} \\ \vdots \\ N_{ur}^{(k)} \end{bmatrix} + \varepsilon_{ur}^{(k)} \quad (2.33b)$$

2.4.2 Double Difference

Double-difference is using one reference station to measure two different satellites (k and l) at a same epoch (t_1). The geometry of the double-difference measurement setup is shown in Figure 2.21.

$$\Delta\varphi_{ur}^{(k)} = \frac{1}{\lambda}d_{ur}^{(k)} + \frac{c}{\lambda}(\delta t_{ur}) + N_{ur}^{(k)} + \varepsilon_{ur}^{(k)} \quad (2.34a)$$

$$\Delta\varphi_{ur}^{(l)} = \frac{1}{\lambda}d_{ur}^{(l)} + \frac{c}{\lambda}(\delta t_{ur}) + N_{ur}^{(l)} + \varepsilon_{ur}^{(l)} \quad (2.34b)$$

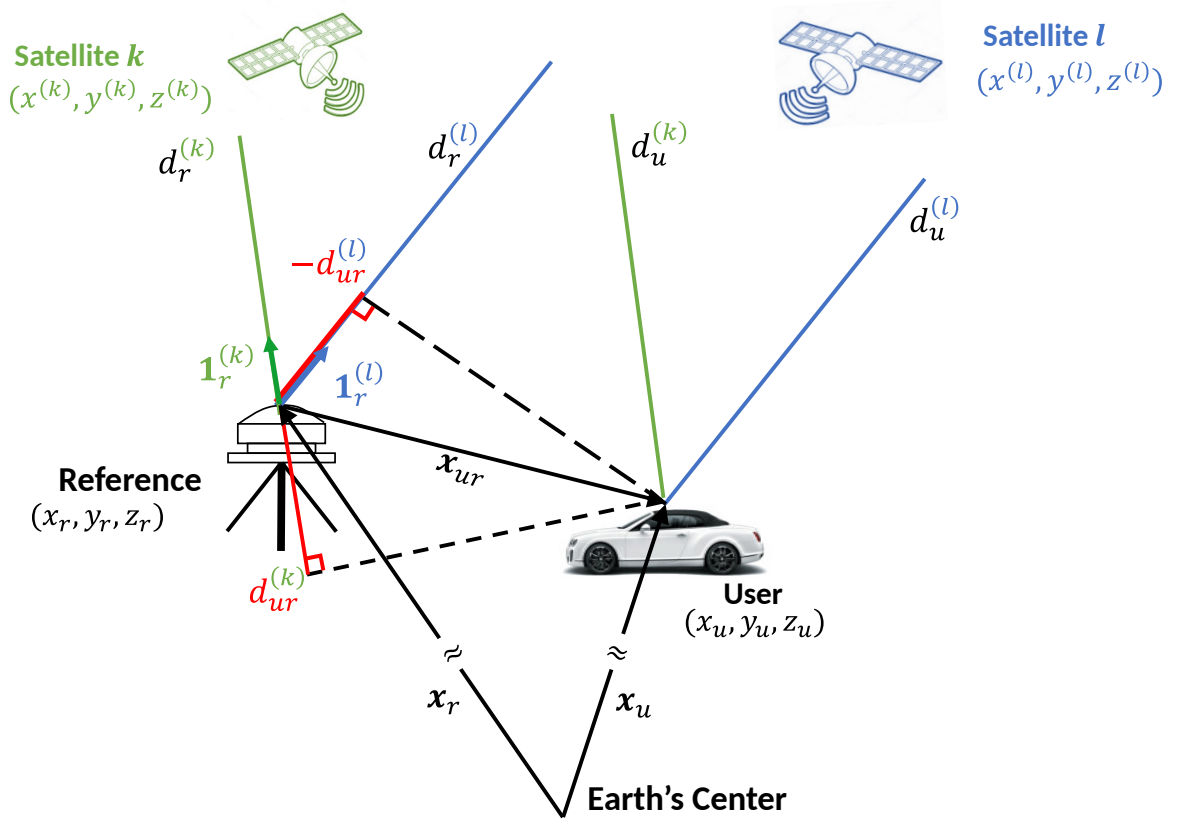


Figure 2.21: Geometry of the double-difference measurement

The reference δt_r and user δt_u clock bias will be eliminated by Eq.(2.34b – 2.34a).

$$\Delta\varphi_{ur}^{(kl)} \xrightarrow{2.34b-2.34a} \Delta\varphi_{ur}^{(k)} - \Delta\varphi_{ur}^{(l)} = \frac{1}{\lambda} d_{ur}^{(kl)} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)} \quad (2.35)$$

solving double difference ambiguity N from Eq.(2.35) by linear model as well, rewriting Eq.(2.35)

$$\Delta\varphi_{ur}^{(kl)} = \frac{1}{\lambda} d_{ur}^{(kl)} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)} \quad (2.36)$$

by using the same linear model for position estimation method

$$d_{ur}^{(kl)} = \left(d_u^{(k)} - d_r^{(k)}\right) - \left(d_u^{(l)} - d_r^{(l)}\right) = -\left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)}\right) \cdot \mathbf{x}_{ur} \quad (2.37)$$

$\mathbf{1}_r^{(k)}$ and $\mathbf{1}_r^{(l)}$ are the unit vector pointing to satellite k and l from the reference station respectively. With J satellites in the sky view, there are $(J - 1)$ double differences, normally selected the highest satellite number $\mathbf{1}$ as the reference to avoid multipath at the Figure 2.21 epoch, substituting Eq.(2.37) into Eq.(2.36)

$$\Delta\varphi_{ur}^{(kl)} = \frac{1}{\lambda}d_{ur}^{(kl)} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)} \quad (2.38a)$$

in vector matrix format

$$\begin{bmatrix} \Delta\varphi_{ur}^{(21)} \\ \Delta\varphi_{ur}^{(31)} \\ \vdots \\ \Delta\varphi_{ur}^{(J1)} \end{bmatrix} = \frac{1}{\lambda} \begin{bmatrix} -\left(\mathbf{1}_r^{(2)} - \mathbf{1}_r^{(1)}\right)^T \\ -\left(\mathbf{1}_r^{(3)} - \mathbf{1}_r^{(2)}\right)^T \\ \vdots \\ -\left(\mathbf{1}_r^{(J)} - \mathbf{1}_r^{(1)}\right)^T \end{bmatrix} \cdot \mathbf{x}_{ur} + \begin{bmatrix} N_{ur}^{(21)} \\ N_{ur}^{(31)} \\ \vdots \\ N_{ur}^{(J1)} \end{bmatrix} + \begin{bmatrix} \varepsilon_{ur}^{(21)} \\ \varepsilon_{ur}^{(31)} \\ \vdots \\ \varepsilon_{ur}^{(J1)} \end{bmatrix} \quad (2.38b)$$

Solving Ambiguities N

Now we need to solve the ambiguities $N_{ur}^{(kl)}$ in Eq.(2.38a), by using the same manner of linear model. The reference station $\mathbf{x}_r$ is known, the user position is $\mathbf{x}_u$ is unknown, the relative position reference and user position need to find out

$$\mathbf{x}_{ur} = \mathbf{x}_u - \mathbf{x}_r \quad (2.39)$$

The first guest position is $\mathbf{x}_0 = 0$,

$$\mathbf{x}_{ur} = \mathbf{x}_0 + \delta\mathbf{x} \quad (2.40)$$

where $\delta(x)$ is the unknown corrections to be applied to the initial estimates,

$$\begin{aligned} d_{ur}^{(kl)} &= - \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right) \cdot \mathbf{x}_{ur} \\ &= - \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right) \cdot \mathbf{x}_0 - \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right) \cdot \delta \mathbf{x} = d_0^{(kl)} - \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right) \cdot \delta \mathbf{x} \end{aligned} \quad (2.41)$$

substituting this expression for $d_{ur}^{(kl)}$ into Eq. (2.38a),

$$\Delta \varphi_{ur}^{(kl)} = \frac{1}{\lambda} d_0^{(kl)} - \frac{1}{\lambda} \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right) \cdot \delta \mathbf{x} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)} \quad (2.42)$$

let set $y_{ur}^{(kl)} = \Delta \varphi_{ur}^{(kl)} - \frac{1}{\lambda} d_0^{(kl)}$ and $\mathbf{g}_{ur}^{(kl)} = -\frac{1}{\lambda} \left(\mathbf{1}_r^{(k)} - \mathbf{1}_r^{(l)} \right)$

$$y_{ur}^{(kl)} = \mathbf{g}_{ur}^{(kl)} \cdot \delta \mathbf{x} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)} \quad (2.43)$$

Now we consider at measurement epoch t_i

$$y_{ur}^{(kl)}(t_i) = \mathbf{g}_{ur}^{(kl)}(t_i) \cdot \delta \mathbf{x} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)}(t_i) \quad (2.44)$$

in all the satellite measurement

$$\mathbf{y}_{ur}(t_i) = \mathbf{G}_{ur}(t_i) \cdot \delta \mathbf{x} + \mathbf{N}_{ur} + \boldsymbol{\varepsilon}_{ur}(t_i) \quad (2.45)$$

where $\mathbf{y}_{ur}(t_i)$ is the difference between the observed and computed the carrier phase double differences for the initial position estimates. $\mathbf{G}_{ur}(i)$ is the observation user-reference geometry matrix. $\mathbf{N}_{ur}$ is the integer ambiguities to be estimated, in the epoch t_{i+1}

$$y_{ur}^{(kl)}(t_{i+1}) = \mathbf{g}_{ur}^{(kl)}(t_{i+1}) \cdot \delta \mathbf{x} + N_{ur}^{(kl)} + \varepsilon_{ur}^{(kl)}(t_{i+1}) \quad (2.46)$$

combining epoch t_i and $t_i + 1$, we could get a continually epochs

$$\begin{bmatrix} y_{ur}^{(kl)}(t_i) \\ y_{ur}^{(kl)}(t_{i+1}) \end{bmatrix} = \begin{bmatrix} \mathbf{G}_{ur}(t_i) \\ \mathbf{G}_{ur}(t_{i+1}) \end{bmatrix} \cdot \delta \mathbf{x} + \begin{bmatrix} \mathbf{I} \\ \mathbf{I} \end{bmatrix} \mathbf{N}_{ur} + \begin{bmatrix} \boldsymbol{\varepsilon}_{ur}(t_i) \\ \boldsymbol{\varepsilon}_{ur}(t_{i+1}) \end{bmatrix} \quad (2.47)$$

In general, the linear system of equation is

$$\mathbf{y}_{ur} = \mathbf{G}_{ur} \cdot \delta \mathbf{x} + \mathbf{A} \mathbf{N}_{ur} + \boldsymbol{\varepsilon}_{ur} \quad (2.48)$$

where

$\mathbf{y}_{ur}$: vector of observed and computed carrier phase double difference

$\mathbf{G}_{ur}$: geometry matrix for the baseline component

$\delta \mathbf{x}$: unknown vector 3-baseline components

$\mathbf{A}$: design matrix for the ambiguities

$\mathbf{N}_{ur}$: unknown vector of double difference ambiguities

$\boldsymbol{\varepsilon}_{ur}$:vectors include measurement noise

by using the least-squares criterion, we look for the minimize the cost function

through a weight matrix Q_y

$$c_y(\delta \mathbf{x}, \mathbf{N}_{ur}) = \min \|\mathbf{y}_{ur} - \mathbf{G}_{ur} \cdot \delta \mathbf{x} - \mathbf{A} \mathbf{N}_{ur}\|_{Q_y}^2 \quad (2.49)$$

when solutions meet the integer least-squares problems, estimator position $\delta \hat{\mathbf{x}}$ and

ambiguities $\widehat{\mathbf{N}}_{ur}$ are the *fixedleast - squares* solutions;

when solutions have unconstrained least-squares problems, estimator position $\delta \hat{\mathbf{x}}$

and ambiguities $\widehat{\mathbf{N}}_{ur}$ are the *floatleast - squares* solution;

where

$$\begin{aligned}
& \| \mathbf{y}_{ur} - \mathbf{G}_{ur} \cdot \delta \mathbf{x} - \mathbf{A} \mathbf{N}_{ur} \|_{Q_y}^2 \\
& = (\mathbf{y}_{ur} - \mathbf{G}_{ur} \cdot \delta \mathbf{x} - \mathbf{A} \mathbf{N}_{ur})^T \cdot \mathbf{Q}_y^{-1} \cdot (\mathbf{y}_{ur} - \mathbf{G}_{ur} \cdot \delta \mathbf{x} - \mathbf{A} \mathbf{N}_{ur})
\end{aligned} \tag{2.50}$$

$\mathbf{Q}_y$ is the variance-covariance matrix of the double difference observation.

The above equation could be solved by using Least-squares Ambiguity Decorrelation Adjustment (LAMBDA) method from [24][25][26].

The first step is to obtain float solutions

$$\text{Float solution} = \begin{bmatrix} \hat{\mathbf{x}} \\ \widehat{\mathbf{N}_{ur}} \end{bmatrix} \tag{2.51}$$

$$\text{Cov } Q_y \begin{bmatrix} \delta \hat{\mathbf{x}} \\ \widehat{\mathbf{N}_{ur}} \end{bmatrix} = \begin{bmatrix} \sum \delta \hat{\mathbf{x}} & \sum_{\delta \hat{\mathbf{x}}, \widehat{\mathbf{N}_{ur}}} \\ \sum_{\delta \hat{\mathbf{x}}, \widehat{\mathbf{N}_{ur}}}^T & \sum \widehat{\mathbf{N}_{ur}} \end{bmatrix} \tag{2.52}$$

where $\delta \hat{\mathbf{x}}$ and $\widehat{\mathbf{N}_{ur}}$ are the float solutions; $\sum \delta \hat{\mathbf{x}}$ and $\sum \widehat{\mathbf{N}_{ur}}$ are the corresponding covariance matrices; $\sum_{\delta \hat{\mathbf{x}}, \widehat{\mathbf{N}_{ur}}}$ is the cross-correlations between the two.

The second step is to find integer vector $\widetilde{\mathbf{N}_{ur}}$ which minimizes the cost function

$$\widetilde{\mathbf{N}_{ur}} = \arg \min_{N \in \mathbf{Z}} \left\| \mathbf{N}_{ur} - \widehat{\mathbf{N}_{ur}} \right\|_{Q_N}^2 = \arg \min_{N \in \mathbf{Z}} \left(\mathbf{N}_{ur} - \widehat{\mathbf{N}_{ur}} \right)^T \cdot \mathbf{Q}_N^{-1} \cdot \left(\mathbf{N}_{ur} - \widehat{\mathbf{N}_{ur}} \right) \tag{2.53}$$

where the integer constraints $N \subset \mathbf{Z}$, Z-transformation leaves the integer nature of integer vectors invariant. This is a nonstandard least-square problem, due to the integer constraint $N \subset \mathbf{Z}$ [25]. The residuals are the difference between $\mathbf{N}_{ur}$ and $\widehat{\mathbf{N}_{ur}}$, when the estimated values of $\delta \hat{\mathbf{x}}$ and $\widehat{\mathbf{N}_{ur}}$ are consider in

$$\text{Residuals} = \mathbf{N}_{ur} - \widehat{\mathbf{N}_{ur}} \quad (2.54)$$

The integer vector solution is validated by a “Ratio-Test”, the ratio factor R is defined as the ratio of the weighted sum of the squared residuals by the second-best solution $\check{\mathbf{N}}_2$ to one by the best $\check{\mathbf{N}}$,

$$R = \frac{(\check{\mathbf{N}}_2 - \hat{\mathbf{N}})^T \cdot \mathbf{Q}_{\check{\mathbf{N}}}^{-1} \cdot (\check{\mathbf{N}}_2 - \hat{\mathbf{N}})}{(\check{\mathbf{N}} - \hat{\mathbf{N}})^T \cdot \mathbf{Q}_{\check{\mathbf{N}}}^{-1} \cdot (\check{\mathbf{N}} - \hat{\mathbf{N}})} > R_{\text{thres}} \quad (2.55)$$

where $\check{\mathbf{N}}$ is the best solution, $\check{\mathbf{N}}_2$ is the second-best solution, R_{thres} is the validation threshold, which are used to check the reliability of the solution.

The third step is to search “fixed” solution $\delta\mathbf{x}$ from Eq.(2.48) and fixing the integer ambiguities to N found in second step. The threshold R_{thres} defines the ellipses searching area, if the R_{thres} is too small, the best solution is out of searching area. However, the R_{thres} is too big, the searching area is huge and searching efficiency get low.

2.4.3 Triple Difference

Triple-difference is using different stations to measure two different satellites (k and l) at two different epochs (t_1 and t_2). The integer ambiguity N is eliminated

$$\Delta\varphi_{ur}^{(kl)}(t_1) = \Delta\varphi_{ur}^{(kl)}(t_1) + N_{ur}^{(kl)}(\mathbf{t}_0) + \varepsilon_{ur}^{(kl)}(t_1) \quad (2.56a)$$

$$\Delta\varphi_{ur}^{(kl)}(t_2) = \Delta\varphi_{ur}^{(kl)}(t_2) + N_{ur}^{(kl)}(\mathbf{t}_0) + \varepsilon_{ur}^{(kl)}(t_2) \quad (2.56b)$$

$$\Delta\varphi_{ur}^{(kl)}(t_1, t_2) \xrightarrow{2.56a-2.56b} \Delta\varphi_{ur}^{(kl)}(t_1) - \Delta\varphi_{ur}^{(kl)}(t_2) - \Delta\varepsilon_{ur}^{(kl)}(t_1, t_2) \quad (2.57)$$

After triple difference, the equation involves only geometric ranges and carrier phase measurement error Eq.(2.57). However, it is lost geometrical leverage. The position estimates obtained from Eq.(2.57) tend to be less accurate than double difference Eq.(2.38a). The principle of triple difference is shown in Figure 2.19.

The Real Time Kinematic (RTK) is a technology could achieve high accuracy and precision position, which could resolve the above Eq.(2.22) unknowns. By using the RTK methodology:

- 1) Single-difference eliminates the satellite clock biases $\delta t^{(k)}$, ionospheric delay I and tropospheric delay T ;
- 2) Double-difference eliminates clock bias in reference δt_r and user δt_u ;
- 3) Triple-difference eliminates the integer ambiguity N .

2.4.4 Precise Point Positioning (PPP) Static

PPP is a positioning technique provides removed GNSS system error to user receiver. International GNSS Service (IGS) broadcasts PPP information of GNSS satellite clock (clock error < 0.1 ns) and orbit (ephemeris error < 5 cm) corrections, which are generated from a network of global reference stations. Once the corrections are calculated, they are delivered to the end user via Network Transport of RTCM via Internet Protocol (NTRIP). These corrections are used by the receiver, resulting in decimetre-level or better positioning with no base station required [27].

RTK- Double Difference system provides a correction data to eliminate major error, I and T in Eq.(2.22) based on the reference station except for the receiver PCV terms with different antennas, normally one base station covers 10-50

km, while the weather condition is not change significantly near the base station. However, the correction data is not efficiently when the user is moving far away the base station.

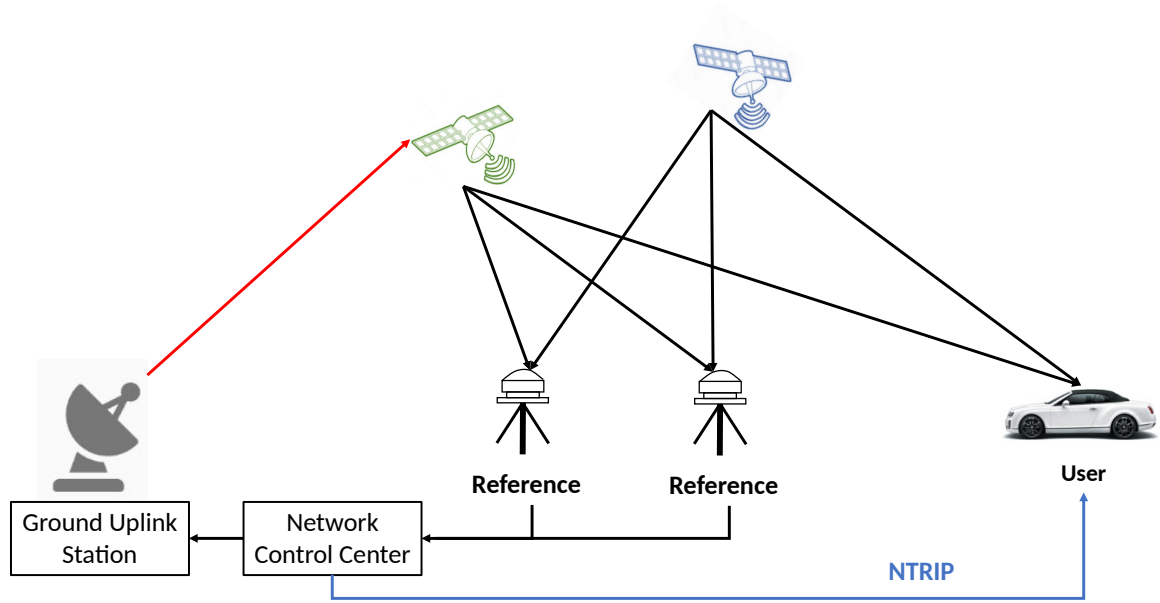


Figure 2.22: The principle of satellite navigation by using Precise Point Positioning

In the PPP mode, the estimation process is similar to the previous RTK method. However, the PPP baseline employs undifferenced, multi-frequency band and ionosphere-free (IF) observations, antenna phase calibration and the precise satellite orbit and clock product.

1) *Undifferenced Mode for PPP*

The IF combination is combined dual-frequency GNSS observation to eliminate nearly all of the ionospheric propagation, the pseudorange ρ_{IF} derive from

Eq.(2.21) and carrier phase φ_{IF} derive from Eq.(2.22) without the I ionospheric delay ϕ .

$$\rho_{IF} = d + T_\rho + c [\delta t_u - \delta t^s] + \varepsilon_\rho \quad (2.58)$$

$$\varphi_{IF} = \frac{1}{\lambda} [d + T_\varphi] + \frac{c}{\lambda} (\delta t_u - \delta t^s) + N + \varepsilon_\varphi \quad (2.59)$$

2) Receiver Antenna Phase Center Model

$$\varepsilon_\varphi = \varepsilon_{phase} + \varepsilon_{modeling} + \varepsilon_{noise} \quad (2.60)$$

the term carrier phase measurement error ε includes ε_{phase} antenna phase center motion error, $\varepsilon_{modeling}$ modeling error and ε_{noise} uncorrelated noise. ε_{phase} antenna phase center motion error includes [26]

$$\varepsilon_{phase} = \varepsilon_{r,PCV} + \varepsilon_{r,PCO} + \varepsilon_{PCV}^s + \varepsilon_{PCO}^s + \varepsilon_{disp} + \varepsilon_{pw} \quad (2.61)$$

$\varepsilon_{r,PCV}$ and $\varepsilon_{r,PCO}$ are the receiver antenna phase center variation and offset;

ε_{PCV}^s and ε_{PCO}^s are the satellite antenna phase center variation and offset;

ε_{disp} satellite antenna phase center model;

ε_{pw} is the satellite antenna phase wind-up effect can vary with the direction of arrival (azimuth and elevation) of a signal and such variation can range under centimeters or even millimeters.

The main purpose of next chapter is to determine receiver antenna phase center PCV $\varepsilon_{r,PCV}$ and PCO $\varepsilon_{r,PCO}$, and evaluate these two errors on high precision positioning impact.

3) Satellite Antenna Phase Center Model

The satellite antenna phase center PCV ε_{PCV}^s and PCO ε_{PCO}^s are based on solutions of several IGS analysis centers, the latest release version is *igs20.atx*, which is available on [28].

4) *Site Displacement by Earth Tides*

Site displacement by earth tides ε_{disp} refers to the vertical or horizontal movement of the Earth's surface due to the gravitational forces exerted by the Moon and the Sun. Earth tides are similar to ocean tides, but instead of affecting only the water, they also cause the solid Earth to deform. The displacement caused by Earth tides can be measured using various geodetic techniques, such as GPS [29], VLBI (Very Long Baseline Interferometry), or satellite altimetry. These techniques can detect even subtle changes in the Earth's surface position over time.

5) *Phase Windup Correction*

Satellite antenna phase windup correction ε_{phase} is a technique used to compensate for the phase errors that occur in the antenna system due to the rotation of the satellite and the movement of the antenna on the ground. These phase errors can introduce distortions in the received signal and affect the accuracy of the data transmission [30].

2.4.5 Multi-Frequency Band Benefit

In the range fundamental code Eq.(2.21) and carrier phase Eq.(2.22) measurement, the ionospheric delay can contribute a large portion of the positional error. This is due to slant delay, when the vertical component of a circularly polarized electromagnetic wave is influenced by the angle of incidence through the ionosphere. The diversity of signal frequencies could significantly overcome the ionospheric delay could be eliminated in a condition by using the single difference method, another approach is multi-frequency band (ex. L1, L2 and L5) receiver can

estimate the I_ρ and I_φ from the measurement without a reference base station. Based on this idea, the dual-band and even triple-band GNSS receiver are well-known and implemented in the automotive industry. In the next section, a multi-band GNSS antenna system will be discussed in detail.

By the physics discover, ionized gas is a dispersive medium for radio waves. The refractive index depends on a radio wave frequency in [4].

It takes an example as dual-frequency L1/L2 band. The estimate of pseudorange measurement ionospheric group delay at L1 is

$$I_\rho(L1) = \frac{f_{L2}^2}{f_{L1}^2 - f_{L2}^2} (\rho_{L2} - \rho_{L1}) \quad (2.62)$$

The estimate of carrier phase measurement ionospheric phase advance at L1 is

$$I_\varphi(L1) = -\frac{f_{L2}^2}{f_{L1}^2 - f_{L2}^2} [\lambda_{L1} (\varphi_{L1} - N_{L1}) - \lambda_{L2} (\varphi_{L2} - N_{L2})] \quad (2.63)$$

The Ionospheric error of I_ρ and I_φ can be both estimated, however I_ρ Eq.(2.62) is unambiguous, while I_φ Eq.(2.63) involves integer ambiguities. A practiced way to solve integer ambiguities is using code measurement to estimate integers [4].

2.4.6 Root-Mean-Square (RMS) Positioning Error

To estimate the position, it is necessary to analyze the RMS position error. Based on the linear system of equation Eq.(2.49) and the cost function Eq.(2.50), it can be concluded that the carrier phase measurement errors have a zero mean, which implies that both the estimated position $\delta\hat{\mathbf{x}}$ and ambiguities $\widehat{\mathbf{N}}_{ur}$ also have a zero mean,

$$E(\hat{\mathbf{x}} - \mathbf{x}) = 0 \quad (2.64)$$

$$E\left(\widehat{\mathbf{N}}_{ur} - \mathbf{N}_{ur}\right) = 0 \quad (2.65)$$

the covariance matrix is calculated in Eq.(2.52), if the components are uncorrelated and each has identical variance σ^2 ,

$$\text{Cov } Q_y \begin{bmatrix} \delta \hat{\mathbf{x}} \\ \widehat{\mathbf{N}}_{ur} \end{bmatrix} = \begin{bmatrix} \Sigma_{\delta \hat{\mathbf{x}}} & \Sigma_{\delta \hat{\mathbf{x}}, \widehat{\mathbf{N}}_{ur}} \\ \Sigma_{\delta \hat{\mathbf{x}}, \widehat{\mathbf{N}}_{ur}}^T & \Sigma_{\widehat{\mathbf{N}}_{ur}} \end{bmatrix} = \sigma^2 \left(\mathbf{G}_{ur}^T \mathbf{G}_{ur} \right)^{-1} \quad (2.66)$$

let σ_x^2 , σ_y^2 and σ_z^2 denote the variances of the 3-baseline x , y and z position error and σ_N^2 be the variance of the double difference ambiguities estimate, set

$$\mathbf{H} = \left(\mathbf{G}^T \mathbf{G} \right)^{-1} \quad (2.67)$$

denoting the i th entry on the diagonal of matrix $\mathbf{H}$ as H_i ,

$$\sigma_x^2 = \sigma^2 H_{11}; \sigma_y^2 = \sigma^2 H_{22}; \sigma_z^2 = \sigma^2 H_{33}; \sigma_N^2 = \sigma^2 H_{44}; \quad (2.68)$$

$$2D \text{ RMS position error} = \sqrt{\sigma_x^2 + \sigma_y^2} = \sigma \sqrt{H_{11} + H_{22}} \quad (2.69)$$

$$3D \text{ RMS position error} = \sqrt{\sigma_x^2 + \sigma_y^2 + \sigma_z^2} = \sigma \sqrt{H_{11} + H_{22} + H_{33}} \quad (2.70)$$

The above equation is based on the Earth-Centered Earth-Fixed system (ECEF) coordinates. In practice, we use the local East-North-Up (ENU) coordinate system to evaluate the position error by rewriting Eq.(2.31).

$$\mathbf{1}^{(k)} = \begin{pmatrix} \cos el^{(k)} \sin az^{(k)} & \cos el^{(k)} \cos az^{(k)} & \sin el^{(k)} \end{pmatrix}^T \quad (2.71)$$

the Eq.(2.67) is in local ENU coordinate

$$\tilde{\mathbf{H}} = \left(\tilde{\mathbf{G}}^T \tilde{\mathbf{G}} \right)^{-1} \quad (2.72)$$

and

$$\sigma_E^2 = \sigma^2 \tilde{H}_{11}; \sigma_N^2 = \sigma^2 \tilde{H}_{22}; \sigma_U^2 = \sigma^2 \tilde{H}_{33}; \sigma_N^2 = \sigma^2 \tilde{H}_{44} \quad (2.73)$$

$$2D \text{ RMS position error} = \sqrt{\sigma_x^2 + \sigma_y^2} = \sqrt{\sigma_E^2 + \sigma_N^2} \quad (2.74)$$

$$3D \text{ RMS position error} = \sqrt{\sigma_x^2 + \sigma_y^2 + \sigma_z^2} = \sqrt{\sigma_E^2 + \sigma_N^2 + \sigma_U^2} \quad (2.75)$$

CHAPTER THREE

GNSS ANTENNA PHASE CENTER CHARACTERIZING

Over the past three decades, Global Navigation Satellite Systems (GNSS) have undergone extensive development. The positional accuracy for civilian users has significantly improved from ± 5 meters to ± 1 meter during this period. However, with the increasing demand for autonomous vehicles, a positional accuracy of ± 1 m will no longer suffice. To address this issue in low-cost mass production vehicles, various potential methods can be employed to enhance position accuracy, including multi-system aliasing, multi-band signal utilization, and carrier phase correction [31]. In the 1990s, conceptual calibration of antenna phase characteristics was introduced through both absolute [32] and relative [33] methods. Subsequently, high-precision GNSS technology enabled studies on anechoic chamber-based absolute method antenna phase information [34] [35]. More recently, there has been research focusing on characterizing phase center variations for GNSS signal processing based on chamber measurements [36]. However, these studies primarily utilize professional survey-grade antennas in chambers and are not feasible for widespread civil applications. Consequently, a comparative study was conducted to characterize antenna phase centers using both high-cost and low-cost antennas [37].

3.1 The GNSS Antenna

As mentioned in the previous chapter, the receiver antenna phase center PCV $\varepsilon_{r,PCV}$ and PCO $\varepsilon_{r,PCO}$ in Eq.(2.61) are influenced by the characteristics of the receiver's antenna. Therefore, it is crucial to ensure a well-designed antenna system. The implementation of a multi-band receiver enhances the estimation accuracy of ionospheric delay and thus automotive Original Equipment

Table 3.1: OEMs multi band GNSS strategy

	L1 (GNSS) 1559-1607 MHz	L2(GPS) 1215-1239 MHz	L5(GPS) 1164-1189 MHz
OEM1	Yes		Yes
OEM2	Yes		Yes
OEM3	Yes	Yes	
OEM4	Yes	Yes	Yes

Manufacturers (OEMs) are actively promoting its adoption with an emphasis on optimizing antenna performance including phase variation.

From the Table 3.1 shown above, the multi-band GNSS antenna strategy varies in different OEMs. OEM1 and OEM2 are American based companies, while OEM3 (with additional L-band correction service 1525-1550 MHz) and OEM4 are European manufactures. To ensure compatibility with various OEM requirements, a L1/L2/L5 multi-frequency band antenna is selected for this research study.

A common antenna system consists of two parts, as shown in Figure 3.1, the passive antenna section and the active Front End Module (FEM) section, which includes pre-filter and Low Noise Amplifier (LNA). This chapter primarily focuses on the design and evaluation of the passive section, including antenna design and ceramic patch selection with consideration for passive gain, Axial Ratio (AR), PCV and PCO. The performance results of PCV and PCO will be used to evaluate the antenna's performance in the following stages;

- 1) a standalone antenna is measured on a circular 250 mm diameter ground plane;
- 2) with the ceramic patch antenna integrated into a compact shark-fin product on the same ground plane;

- 3) with that same shark-fin product mounted on a vehicle roof;
- 4) on-vehicle positioning accuracy analysis with and without PCV correction.

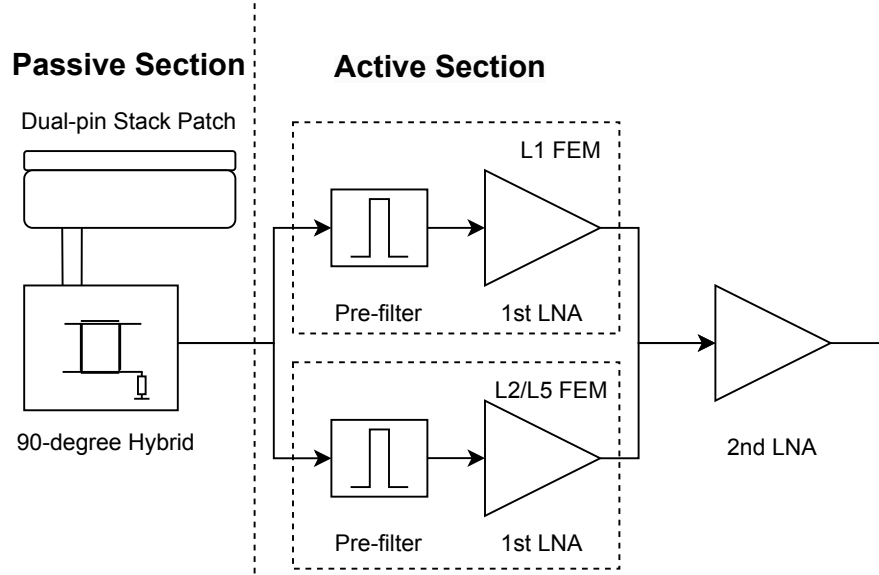


Figure 3.1: Antenna system diagram

3.1.1 Passive Section Design

Antenna selection and design constitute the initial crucial step in system design. The passive antenna's performance is determined by the receiving system's gain, such as Carrier Noise ratio (CN0) on pseudorange ρ from Eq.(2.21).

Additionally, the antenna performance impacts Dilution of Precision (DOP), which specifies the effect that navigation satellite geometry on positional measurement precision. On designing a product for the automotive industry, the cost control and

mass production quality are both important topics not to be avoided. Thus, a good balance of antenna performance and price needs to drive the design process.

The passive antenna, being the primary receiver of satellite signals, plays a crucial role in determining system performance. Therefore, careful selection of the antenna type is important. Considering the requirements for mass production, reliability, and cost-effectiveness, the square ceramic patch antenna appears to be an excellent initial choice due to its compact size and simple assembly process. Moreover, square patch antennas generally exhibit a symmetric surface current distribution which can potentially lead to better Phase Center Variation (PCV). Consequently, this study focuses on designing a square ceramic patch antenna in subsequent sections.

In order to find the patch antenna field radiated performance, it can be evaluated magnetic current density (M_1 and M_2) and electric field (E_1 and E_2) by patch slots, as slot1 and slot2 examples are demonstrated in Figure 3.2. The theoretical calculation is based on the perfect boundary condition. On the following section, the field result could get influence by the boundary condition changes. Furthermore, the later antenna far-field phase result will explain the degradation. The perfect boundary condition:

Electrical wall: patch antenna top and bottom layer

$$\mathbf{E}_x(z' = 0 \text{ and } z' = h) = 0 \quad (3.1a)$$

$$\mathbf{E}_y(z' = 0 \text{ and } z' = h) = 0 \quad (3.1b)$$

Magnetic wall: from top to bottom layer

$$\mathbf{H}_z(0 < z < h) = 0 \quad (3.1c)$$

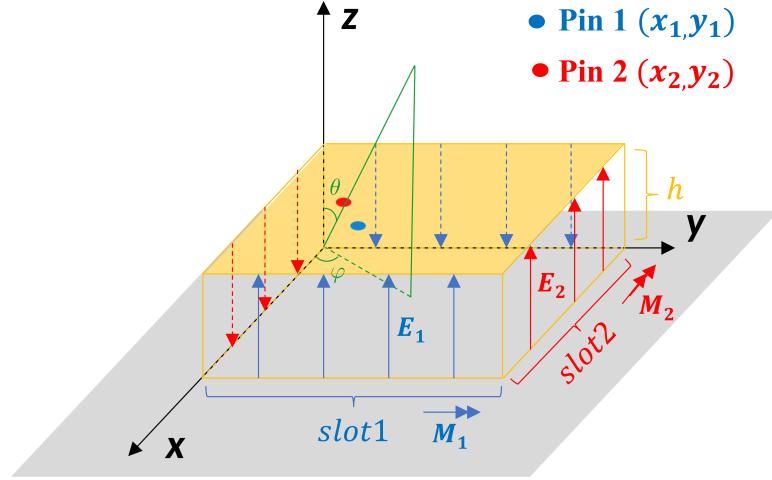


Figure 3.2: Rectangular 2-pin patch with electrical radiating slots E and equivalent magnetic current densities M ;

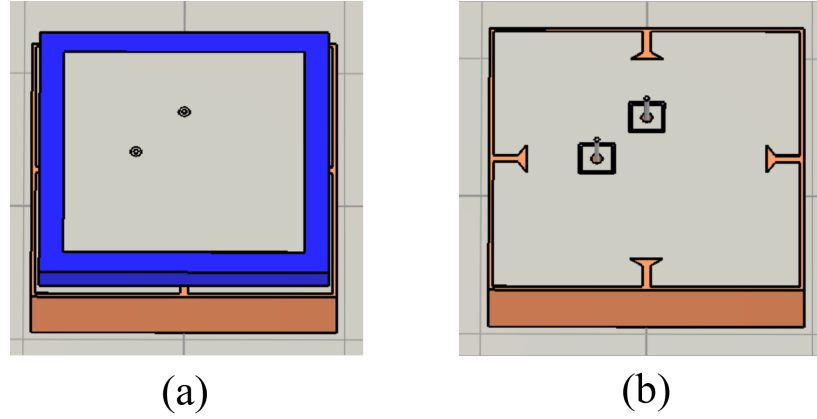


Figure 3.3: Square patch antenna simulation model;
(a) top layer of L1 band patch: $36 \times 36 \times 3mm$ (LWH), DK 10;
(b) bottom layer of L2/L5 band patch: $38 \times 38 \times 7mm$ (LWH), DK 13.

The antenna is fed by two orthogonal of phase pins (Pin1 TM_{100}^Z and Pin2 TM_{010}^Z modes) on the top layer, which is a pair of orthogonal degenerate modes. Out of phase port, is generated by a 90-degree hybrid coupler (Figure 3.6) with right-hand polarization in this case [38][39].

The multi-band GNSS system could utilize multiple frequency bands to reduce the major ionospheric error in Eq.(2.62 and 2.63). The challenge is to design an antenna system that can efficiently and affordably mitigate ionospheric error in order to increase GNSS positional accuracy. Considering each frequency phase center and the receiver module architecture, a stacked patch will have less phase center variation and cable bias than two separate patches. The above discussion is a general antenna selection guideline and it may have other applicable approaches.

A Right Hand Circularly Polarized (RHCP) ceramic patch antenna is commonly used in modern industry products, because of its physical stability, low profile, manufacturability and affordable cost. A typical patch antenna is designed and simulated by a Commercial-Off-The-Shelf (COTS) three-dimensional, full wave electromagnetics (EM) field Time Domain solver of CST[®] based off the method of time domain finite integration technique. As shown in Figure 3.3, the top layer (blue) is designed to receive the L1 band signal. The top L1 stacked patch ($LWH : 36 \times 36 \times 3mm$) material has a ceramic dielectric constant (DK) 10. The bottom layer (orange) is designed to receive L2/L5 band signal, as shown in Figure 3.3 (c), the bottom L2/L5 stacked patch ($LWH : 38 \times 38 \times 7mm$) material is DK 13, and the total height after assembly including adhesive tape is 11 mm. In order to fulfill vehicle styling, a compact shark-fin antenna design is widely implemented. Thus, the patch antenna's dimension is limited by the shark-fin antenna housing [40].

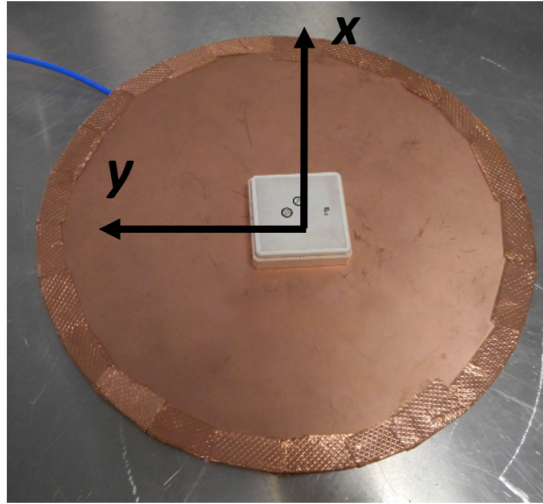


Figure 3.4: Real patch antenna EVB on 250mm circle ground plane

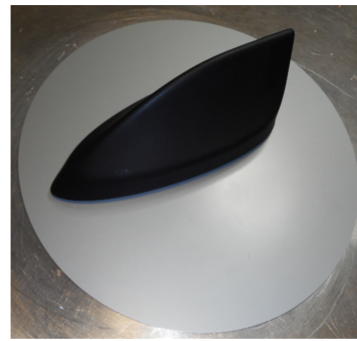
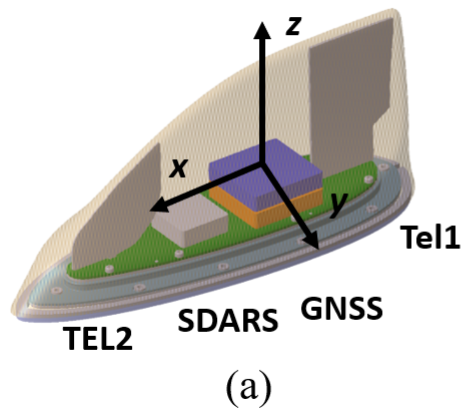


Figure 3.5: CAD model of shark-fin antenna;
 (a) Isometric view of shark-fin antenna;
 (b) Shark-fin antenna on 250mm circle ground plane

A few engineering samples are fabricated to validate the design, and these samples are mounted on a ground plane. Figure 3.4 depicts a standalone real patch antenna mounted on an evaluation board (EVB), which is positioned on a 250 mm circular ground plane. Figure 3.5(a) illustrates a production-intent shark-fin antenna with an isometric view in CAD software. The proposed L1/L2/L5 triple-band GNSS antenna is highlighted using the same color scheme as shown in Figure 3.3. The shark-fin radome also accommodates a Satellite Digital Audio Radio Services (SDARS) patch antenna located at the front of GNSS patch, along with two cellular radiators (Tel1 and Tel2) situated at rear/front side of the shark-fin antenna. Figure 3.5(b) presents an actual shark-fin antenna mounted on a similar 250 mm circular ground plane.

The antenna passive section (Figure 3.1) is including the patch antenna and 90-degree hybrid coupler, the measurement result (Figure 3.6) at coupler output port (black) have a wide band matching performance to cover L1, L2 and L5. In the practice, return loss bandwidth should less than -10 dB.

The simulation results of surface current densities on the evaluation board are demonstrated in Figure 3.7. In this quasi-perfect boundary condition, the majority of current densities exhibit a symmetric distribution along the patch center line (y -axis). When the port is excited at L1 1575 MHz with a 90-degree phase signal, the maximum current density is approximately 38.0144 dB (A/m) beneath the patch's ground plane.

Meanwhile, a simplified CST simulation model of a shark-fin antenna without delicate mechanical structures is presented in Figure 3.8. As mentioned above, the small ground size of the GNSS antenna would limit the electrical wall condition Eq.(3.1a) and Eq.(3.1b), the antenna surrounding elements would break the GNSS patch antenna magnetic wall boundary condition Eq.(3.1c). The

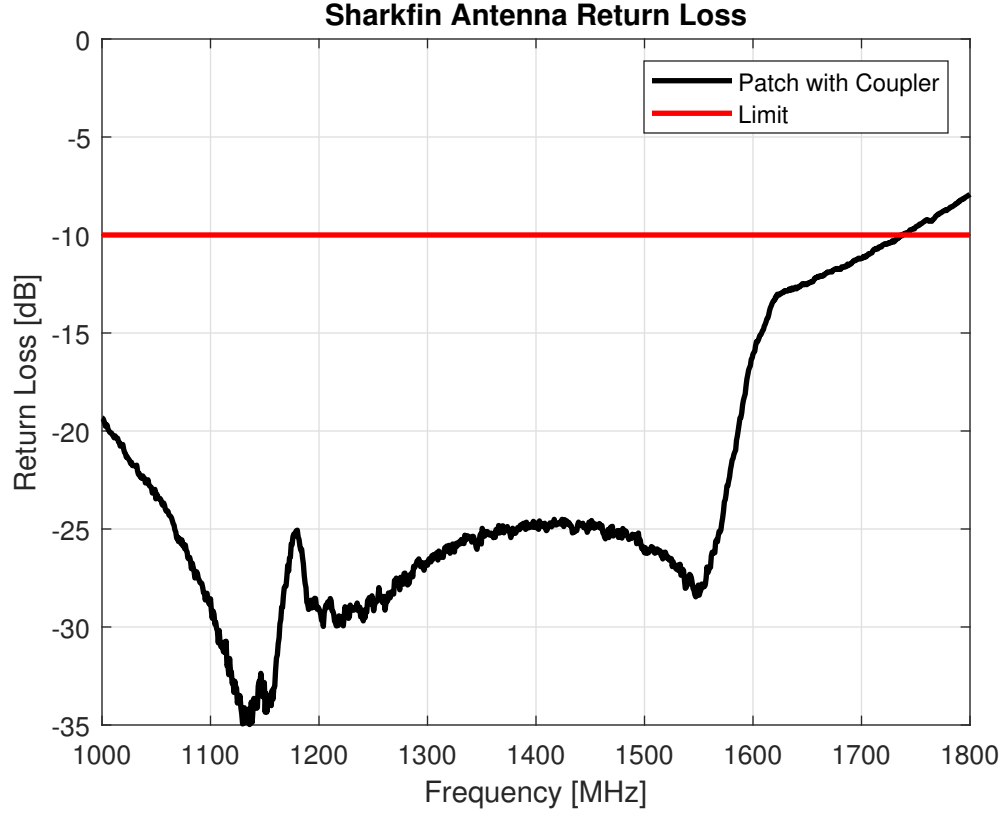


Figure 3.6: Patch antenna return loss

simulation results show that both Tel1 element and SDARS patch have an influence on surface current densities. As indicated by markers at Tel1 bottom point, coupling current densities can reach up to 7.74162 dB (A/m). Due to the height of the Tel1 element, there is a coupling current distribution on the vertical direction. This kind of impact would involve the GNSS antenna performance at low elevation angle until the patch antenna have an open sky view.

Similar coupling current densities are observed on SDARS patch antenna surface as well as for L2 and L5 frequency bands according to simulation results

shown here. Considering that L1 band (1575.42 MHz) is closer to cellular Band 3 (1710-1880 MHz), Tel1 element has more impact on L1 band than L5 band. Real measurement result will be presented in subsequent subsections.

Comparing Figure 3.7 and Figure 3.8, it is evident that the GNSS antenna in close proximity to the surrounding elements exhibits an isolation issue, as indicated by the analysis of surface current densities. The coupling current on the adjacent element breaks the boundary condition of the GNSS antenna, leading to further degradation in its far-field performance. During the design stage, several strategies can be employed to mitigate this influence

- A certain distance between the elements, Tel1 radiator with GNSS patch's distance; SDARS patch with GNSS patch's distances;
- Telephone antenna resonate structure with GNSS Frequency notch out;
- SDARS patch rotation in certain angle;
- Anti-resonate frequency tuning away GNSS in-band frequency.

The methods are not limited to the aforementioned ones; they should encompass diverse approaches. By utilizing EM software for simulating E-field / H-field / current densities, a visual method is provided to guide design work. Additionally, it is crucial to have a bench sample for manual tuning. However, the co-existing problem has a mutual impact, as nearby GNSS antennas can also experience performance degradation. The golden rule of a better decoupling solution is having further physical distance.

3.1.2 Indoor Antenna Range Setup

The indoor far-field antenna range is provided by Continental Advanced Antenna Inc, USA. The distance between transmission antenna and Antenna Under

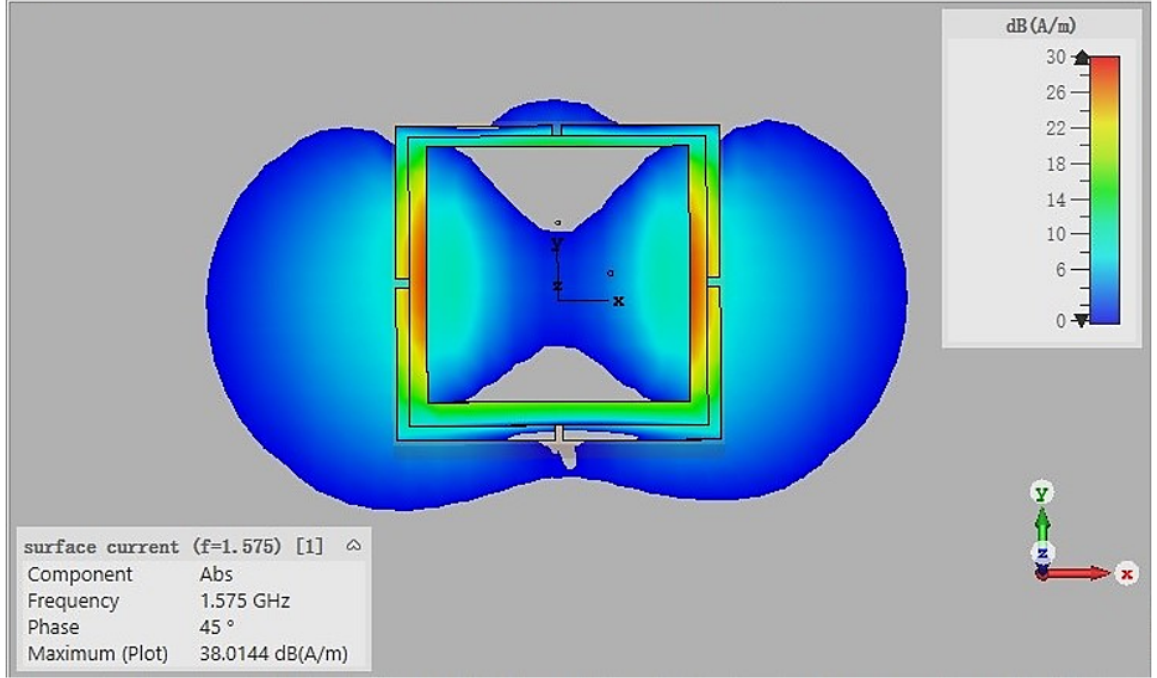


Figure 3.7: Simulation result of surface current densities at EVB 1575 MHz

Test (AUT) is 9.14 m, which satisfies the far-field requirement in L1/L2/L5 frequency bands. The antenna test range's setup is demonstrated in Figure 3.9(a). The indoor test range is set for a spherical coordinate system. The configuration of the upper hemisphere is such that

- the angle of antenna elevation ($90 - \theta$): 0 to 90° , with a step of 1° ;
- the angle of antenna azimuth (ϕ): 0 to 360° , with a step of 5° ;
- the initial rotation center: the Antenna Reference Point (ARP) $\theta = 0^\circ$ or elevation = 90° , $\phi = 0^\circ$, PCV = 0 mm;
- the patch antenna is mounted on an offset arm, which ensures the AUT is in the rotation center, as shown in Figure 3.9(b).

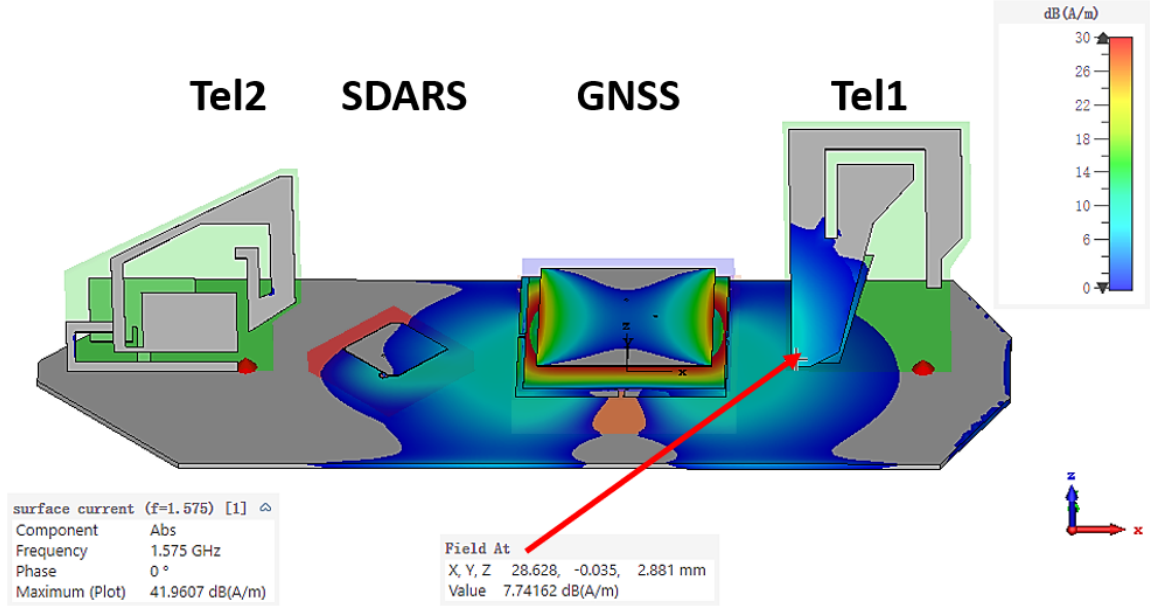
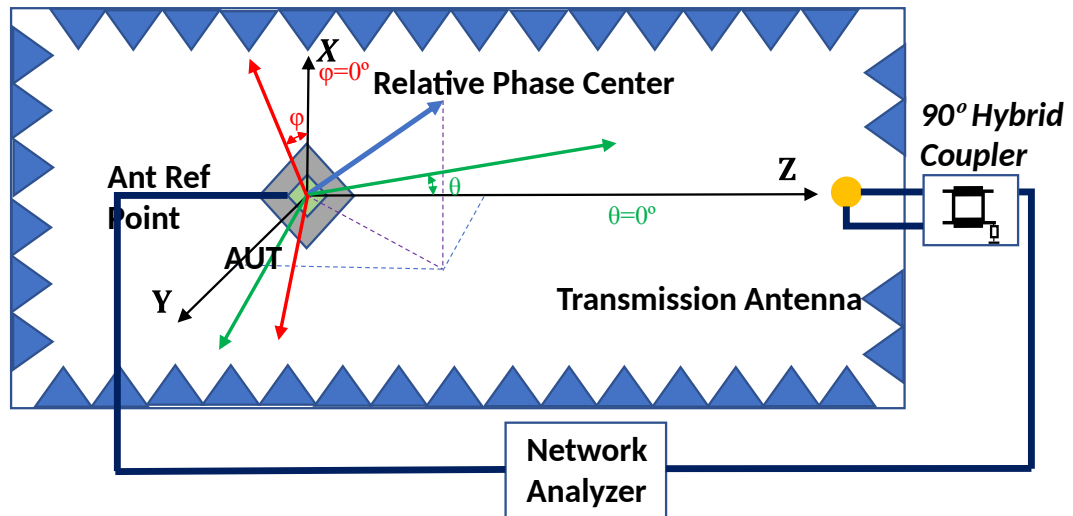
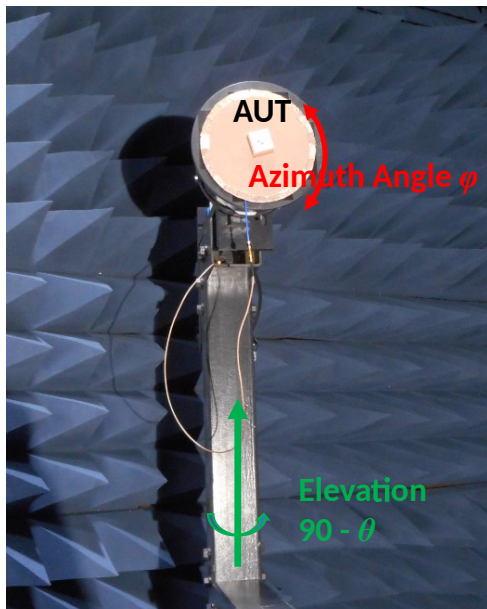


Figure 3.8: Simulation result of surface current densities at sharkfin antenna 1575 MHz

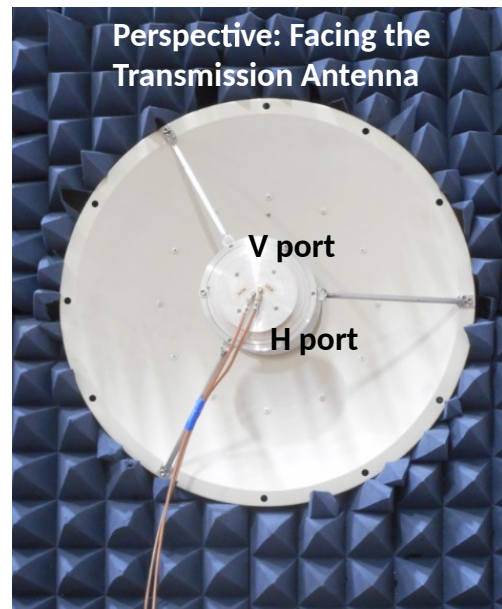
The vertical and horizontal electrical plane radiation patterns are generated by the two-port cross-dipole parabolic reflector antenna depicted in Figure 3.9(c). The incident signal is transmitted through the vertical (V) and horizontal (H) port cables to vector network analyzer. Subsequently, post-processing is performed to determine the output phase delay based on the polarization chirality. In the following section, we analyze both Right-Hand Circular Polarization (RHCP) and Left-Hand Circular Polarization (LHCP). This directional parabolic reflector antenna excites a narrow beam with a half-power beamwidth of 18° , effectively minimizing radiation reflections from within the anechoic chamber walls.



(a)



(b)



(c)

Figure 3.9: Indoor antenna test range setup;
 (a) Diagrammatic drawing;
 (b) AUT and turntable;
 (c) Transmission antenna

3.1.3 Antenna Gain and Axial Ratio

Traditionally, the performance of GNSS antennas is assessed based on their Linear Average Gain (LAG) and Axial Ratio (AR) over the elevation angle at the center frequencies of L1, L2 and L5 band. These results are depicted in Figure 3.10, Figure 3.11 and Figure 3.12 respectively. The LAG represents the antenna gain (dBi) across all azimuth angles (0 to 360°), while the axial ratio (dB) is calculated similarly.

In the Figure 3.10 left, along the horizontal axis, the elevation angles from 0° to 90° are divided into increments of 6°. The dashed curve represents the average axial ratio in unit decibel (dB), with values read from the left vertical axis. On the other hand, the solid curve represents the LAG in dBi, with values are read from the right vertical axis. The legend [Simulation EVB 1575] represents the electromagnetic simulation result of a standalone patch antenna on a 250 mm circular ground plane at 1575 MHz, as depicted in Figure 3.3(b). The legend [EVB 1575] denotes an evaluation board standalone patch antenna on a 250 mm circular ground plane at 1575 MHz within an indoor antenna range, as shown in Figure 3.3(a). The legend [Sharkfin 1575] signifies the compact multi-function shark-fin antenna on a 250 mm circular ground plane at 1575 MHz within an indoor antenna range (Figure 3.5). The remaining steps can be performed similarly.

The simulation results for the LAG show good correlation with the real antenna sample in triple band. The LAG ranges from -9 dBi to +5 dBi over an elevation range of 0° to 90°. Within the tested range, the real-world measurements do not deviate from the simulation results by more than 2 dBi.

In the L1 frequency band shown in Figure 3.10 on the left, the simulation and EVB results match, while the shark-fin results exhibit performance degradation

at lower elevation angles ($< 60^\circ$). It is clearly shown on the right side of the figure, where the shark-fin GNSS antenna's omni-directivity is further degradation, especially on the front orientation. Referring back to the simulation result of current density in Figure 3.8, it can be seen that the GNSS L1 band is influenced by coupling surface current from nearby Telephone 1 frequency Band 3 (1710 - 1880 MHz). Due to this coupling, disturbances occur in patch antenna current densities and result in a null radiation pattern in the rear direction. As a result, circularly polarized fields are degraded, which explains why there is a 4 dB axial ratio at an elevation angle of 60° on the left side.

The real measurement results show that when the GNSS antenna is assembled with other elements in a compact multi-functional package, the electrical performance is affected. The simulation result shows that the antenna performance nearly as expected in a perfect electrical wall boundary condition Eq.(3.1a) and Eq.(3.1b), with good agreement in the theory calculation circular polarization, such as axial ratio. However, the two telephone antennas (Tel1 and Tel2) could be considered as directors or reflectors for the patch antenna, which breaks the perfect magnetic wall boundary condition Eq.(3.1c). As a result, there will be deviations between the simulated shark-fin GNSS antenna's far-field radiation pattern and phase performance compared to both simulation and EVB antenna's result. The SDARS antenna has a similar ceramic construction but is smaller in size physically. The shown results represent the best tuning of each antenna while maximizing the physical distance between each element; however, interference between these the co-existing system is inevitable. To minimize this influence and prioritize key performance metrics within a desired range, an axial ratio below 3 dB at approximately 60° elevation angle is targeted.

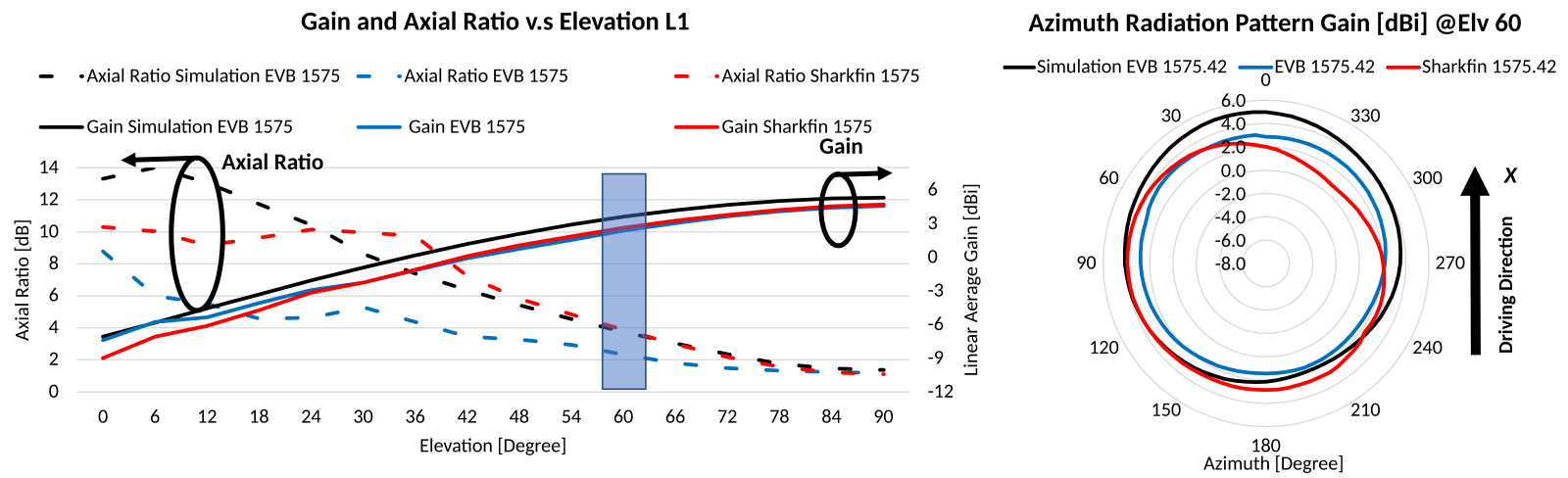


Figure 3.10: Linear average gain and axial ratio over elevation at L1 band

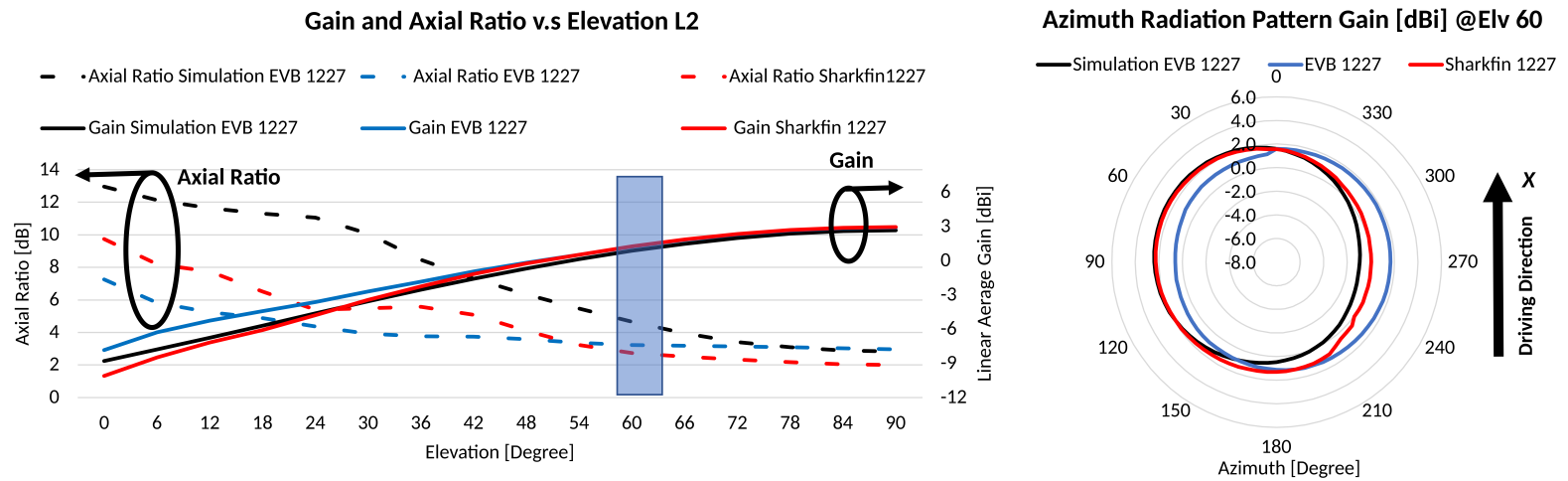
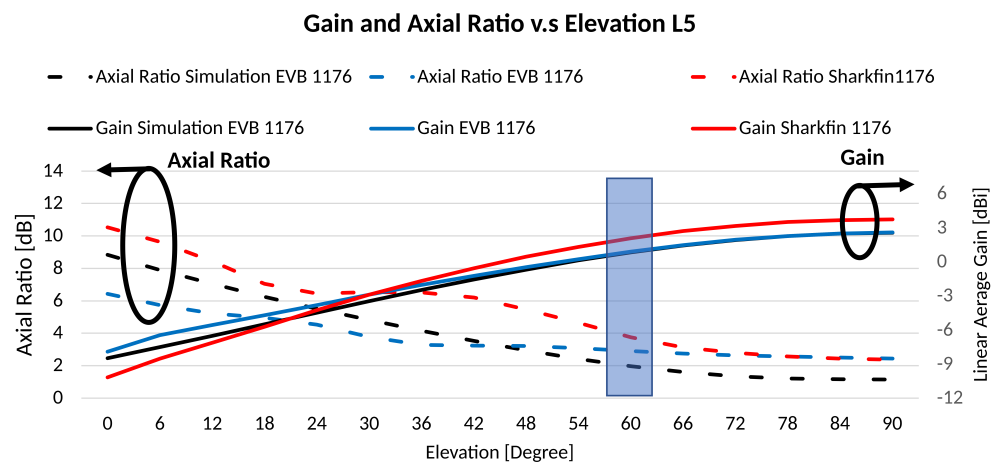


Figure 3.11: Linear average gain and axial ratio over elevation at L2 band



Azimuth Radiation Pattern Gain [dBi] @Elv 60

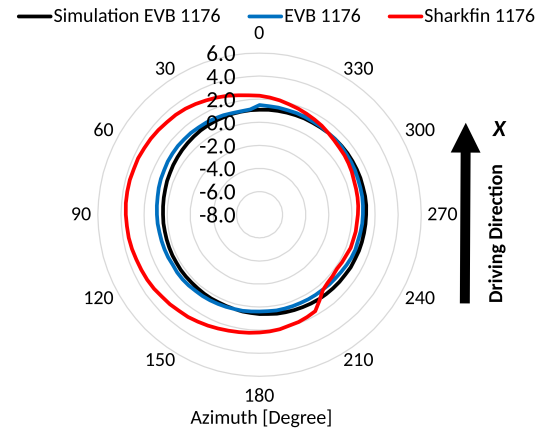


Figure 3.12: Linear average gain and axial ratio over elevation at L5 band

3.1.4 The Up / Down Ratio

The GNSS Antenna is mounted on a vehicle, which is positioned on the ground. In this scenario, the ground acts as an electrical conductor wall. When an incident circular polarization signal reflects off the ground plane, the vertical polarization of an electrical wave remains unchanged, while the horizontal polarization of the same electrical wave becomes inverted, as shown in Figure 3.13. This phenomenon is known as *Image Theory* according to [41]. The reflected LHCP wave results from the reflection of the incident RHCP wave when it exceeds the *Brewster Angle*. For high-quality GNSS antennas, this cross-polarization can introduce significant errors in both pseudorange measurement as well as carrier phase measurement.

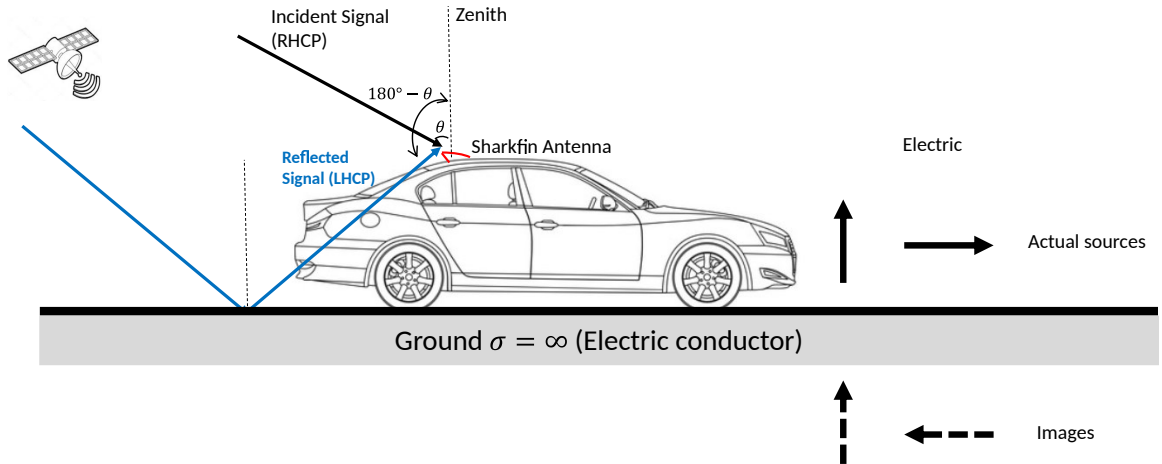


Figure 3.13: Incident signal reflection

The Up / Down gain ratio, which is averaged over azimuth angels, is used to evaluate the multipath ground effect in [39] and [42]. The equation below is employed to determine the Up / Down gain ratio:

$$\frac{Up}{Down}Ratio = \frac{G_{RHCP}(\theta)}{G_{LHCP}(180^\circ - \theta)} \quad (3.2)$$

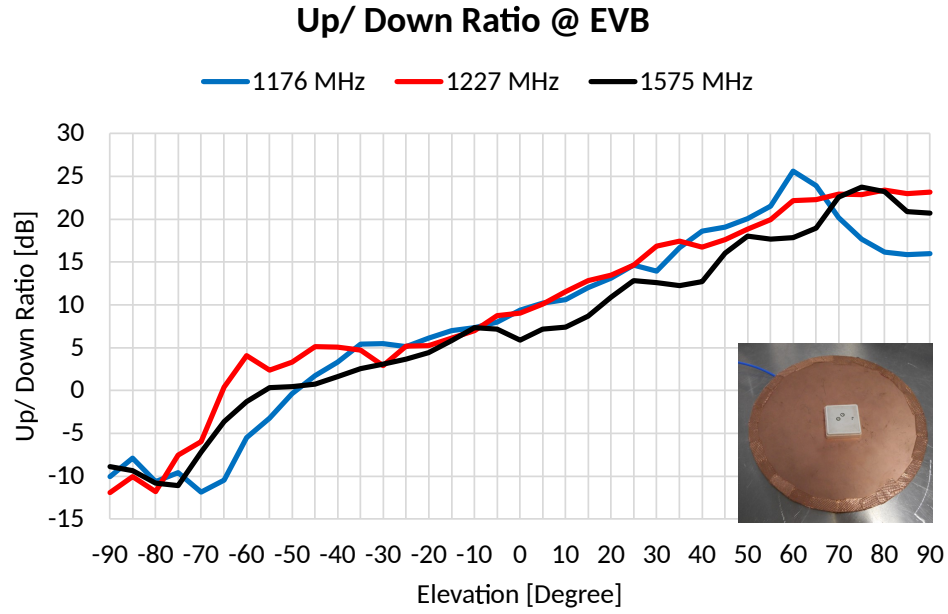


Figure 3.14: Up / Down ratio at EVB

Figure 3.14 and Figure 3.15 display the result of the Up / Down gain ratio at the center frequencies of the L1 band, L2 band and L5 band for a standalone patch on an EVB and in the compact multifunction shark-fin antenna respectively. The

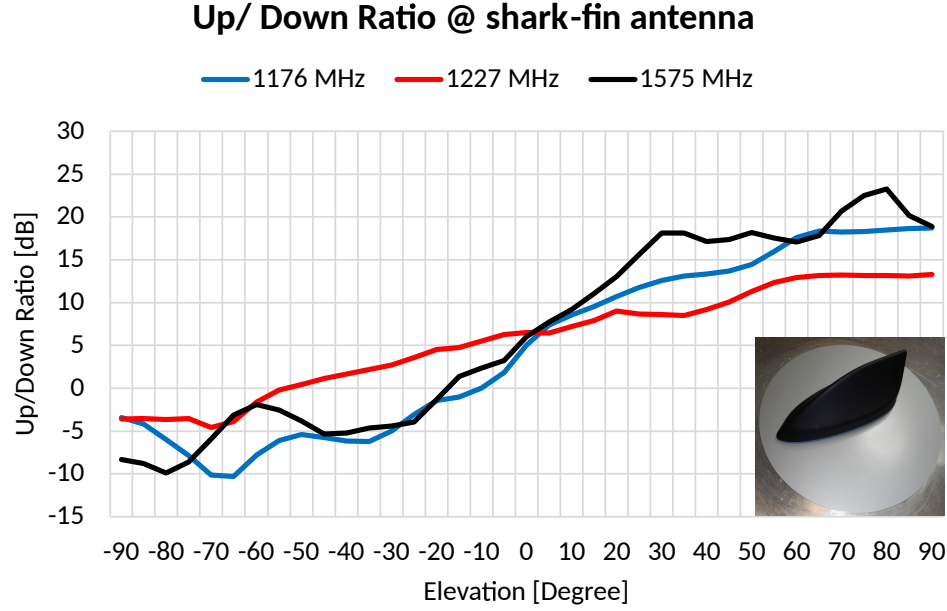


Figure 3.15: Up / Down ratio at shark-fin antenna

test results are set up in the same manner as shown in Figure 3.4. According to [43], dry soil ground has a Brewster Angle of 22° , which serves as the cut-off angle for reflected signals. The Up / Down gain ratio range is above 10 dB for elevation higher than 20° . This demonstrates that the antenna exhibits good cross polarization (x-pol) and can overcome multipath ground effects.

3.1.5 Outdoor Antenna Range Setup

The standalone patch and multi-function shark-fin antenna are tested in the indoor lab. After that, on-vehicle measurements are conducted at Oakland University's quasi-far-field antenna range, as shown in Figure 3.16. The shark-fin antenna is mounted on a sedan vehicle with a curved roof, ensuring proper ground

connection and accurately representing real-world performance on a curved ground plane [44]. A zoomed-in detail of the mounted antenna is provided in the bottom left corner. The vehicle is parked on an electronically controlled rotating steel turntable which served as an electric conductor. The transmitting antenna used for testing is a cross-dipole parabolic reflector antenna (same as indoor transmission antenna, Figure 3.9c) mounted on a white fiberglass arched arm. The height of the arm's rotational axis could be adjusted to match with the vehicle height, allowing for accurate positioning of the AUT at the center of the turntable. This setup enabled illumination of the car roof by the transmitting antenna in a spherical coordinate system. A comparison between post-processed data from indoor and outdoor test ranges will be presented in the next subsection.

3.1.6 Phase Center Variation and Phase Center Offset

In the antenna design theory, the phase center is defined as *the point from which electromagnetic radiation spreads spherically outward, with the phase of the signal being equal at any point on the sphere* [45], as demonstrated in Figure 3.17. If the antenna's phase contour is ideally spherical, the phase center is located at the physical center point of the antenna, also known as the Antenna Reference Point (ARP) [32] and [41]. However, in reality, an antenna's phase contour cannot be perfectly spherical. The actual far-field phase deviates from the mean ideal phase depending on signal directionality. This deviation from mean ideal phase is referred to as PCV. A lower PCV corresponds to a stable phase variation in phase $\varepsilon_{r,PCV}$ according to Eq.(2.61). Measurements of an antenna's received phase can be used to calculate its corresponding phase center by determining where each specific direction originates based on real measured signals' phases for different elevations and arc lengths. This calculation involves creating a triangle using two adjacent

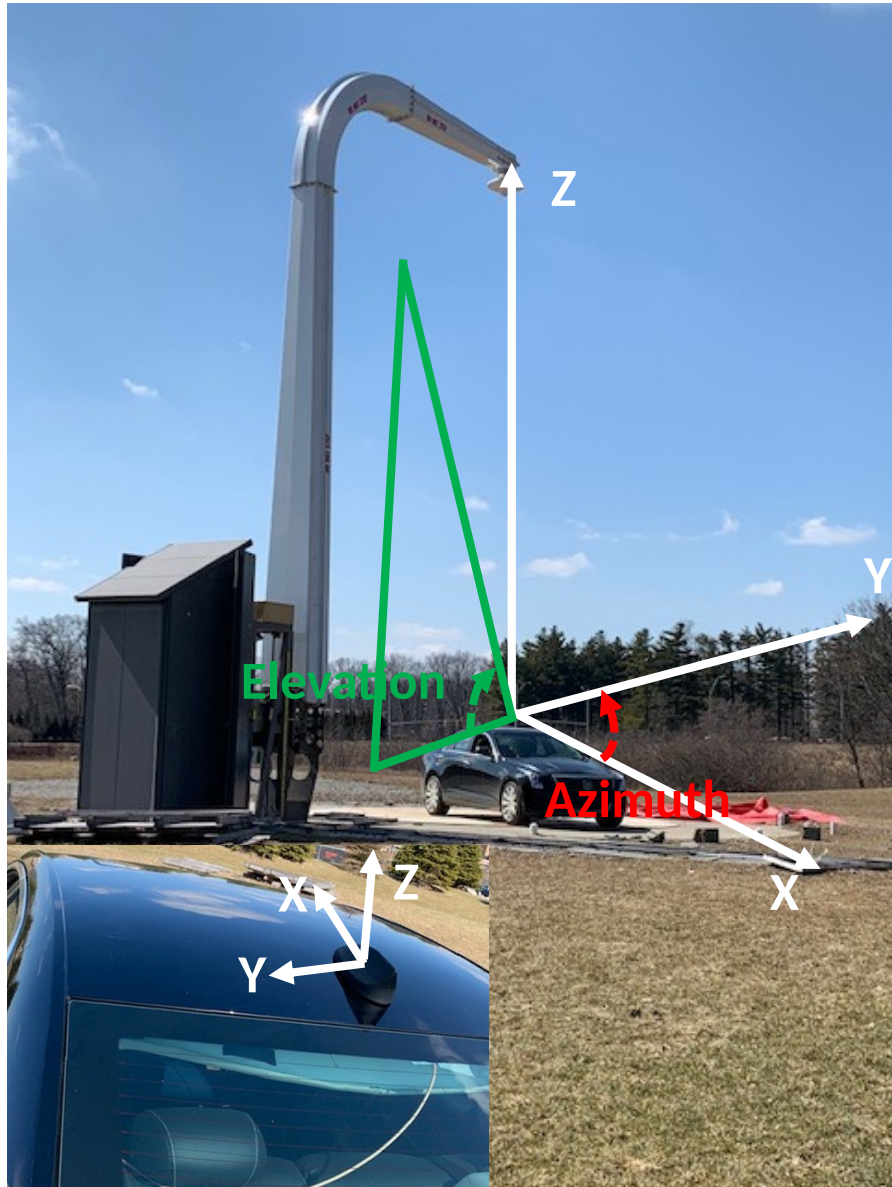


Figure 3.16: Oakland university outdoor antenna range

measurement points' angles and their measured phases as outer vertices and calculating the inner vertex (the real phase center) that represents where a signal originating from would produce those measured phases at these points. *The distance between the origin of this signal and the reference point is denoted as PCO.* The initial correction of PCO can only be done accurately if the distance to transmitted signals is precisely known. A lower PCO indicates a more concentrated AUT (Antenna Under Test) phase center position. Therefore, both PCV $\varepsilon_{r,PCV}$ and PCO $\varepsilon_{r,PCO}$ are important parameters for evaluating an antenna's performance in terms of its phased characteristics according to Eq.(2.61).

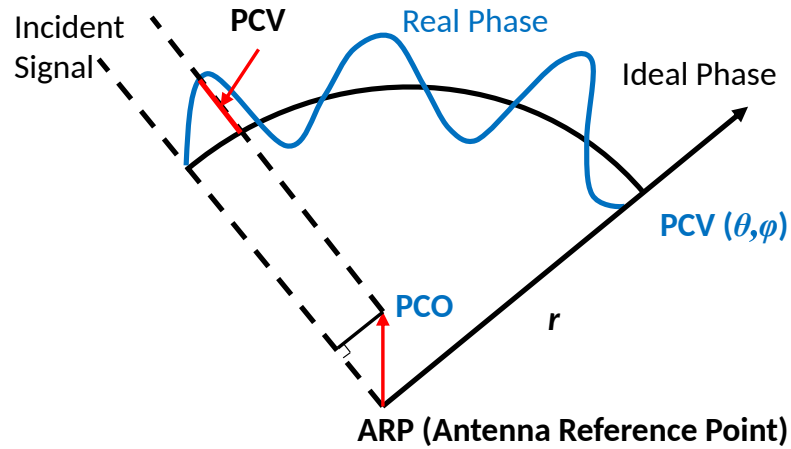


Figure 3.17: PCV and PCO model

Furthermore, the variation in frequency will cause changes in phase center due to the difference in wavelengths between Eq.(2.62) and Eq.(2.63). The determination of the phase center is presented for the L1 band (1575 MHz), L2

band (1227 MHz) and L5 band (1176 MHz) using a spherical phase method based on a least squares method [45], which requires an upper hemisphere far-field radiation pattern that includes phase information $Phase(\theta, \varphi)$ following spherical coordinate conventions. A far-field phase expansion can be expressed as

$$Phase(\theta, \varphi) = \frac{2\pi}{\lambda}(x \cdot \sin \theta \cos \varphi + y \cdot \sin \theta \sin \varphi + z \cdot \cos \theta) \quad (3.3)$$

Let the phase center origin be located at $PCO(x, y, z)$, which is the unknown parameter to be solved for, and λ represents the wavelength of the measured signal in millimeters [mm].

Rewriting the Eq.(3.3) in matrix format

$$Phase(\theta, \varphi) = \frac{2\pi}{\lambda} [\sin \theta \cos \varphi \quad \sin \theta \sin \varphi \quad \cos \theta] \cdot \begin{bmatrix} x \\ y \\ z \end{bmatrix} \quad (3.4)$$

$$\begin{bmatrix} x \\ y \\ z \end{bmatrix} = \frac{\lambda}{2\pi} [\sin \theta \cos \varphi \quad \sin \theta \sin \varphi \quad \cos \theta]^{-1} \cdot Phase(\theta, \varphi) \quad (3.5)$$

The received phase of the antenna from different Angles of Arrival (AOA) is extracted from the far-field measurement results. If an AOA phase correction could be applied to adjust the preliminary phase measurement result in advance, then the localization system could retrieve the predefined PCV result for distance calculation. This would address the antenna phase-center variation error $\varepsilon_{r,PCV}$ in Eq.(2.61) mentioned in [4], leading to improved carrier phase accuracy.

The range measures the received signal $Phase(\theta, \varphi)$ and uses this phase data to calculate the PCV based on wavelength. Furthermore, the final PCV result for a given angle can be expressed in millimeters [mm].

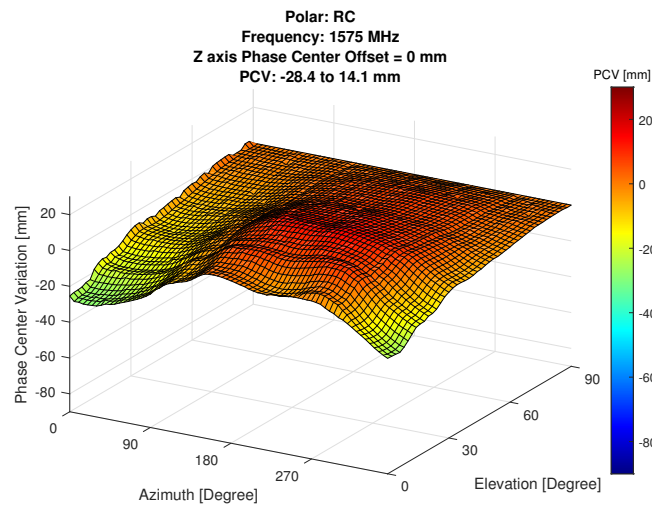
$$PCV = \frac{Phase(degree)}{360^\circ} \times \frac{c(speed\ of\ light)}{Frequency(Hz)} [mm] \quad (3.6)$$

Figure 3.18(a) shows the PCV $Phase(\theta, \varphi)$ at the center of the L1 Band (1575 MHz) on an evaluation board. In the result of PCV $Phase(\theta, \varphi)$, the horizontal axes represent elevation and azimuth angles, while the vertical axis represents PCV shown as a Jet colormap to indicate variations in millimeter. At zenith (Elevation = 90°), the phase variation is zero, which serves as the phase pattern ARP. As elevation decreases, the PCV changes from -28.4 to +14.1 mm across both elevation and azimuth angles. When placing the patch in a shark-fin antenna as illustrated in Figure 3.18(b), the total range of PCV enlarges to -35.3 to +25.2 mm. After the indoor range testing, the outdoor range is conducted on-vehicle at Oakland University (Figure 3.16). The same shark-fin antenna is placed on a test vehicle and its results are represented in Figure 3.18(c). The PCV results shown a similar trend as discussed for the indoor EVB and shark-fin antenna; however, there is an increase in variation from -74.6 to +7.6 mm.

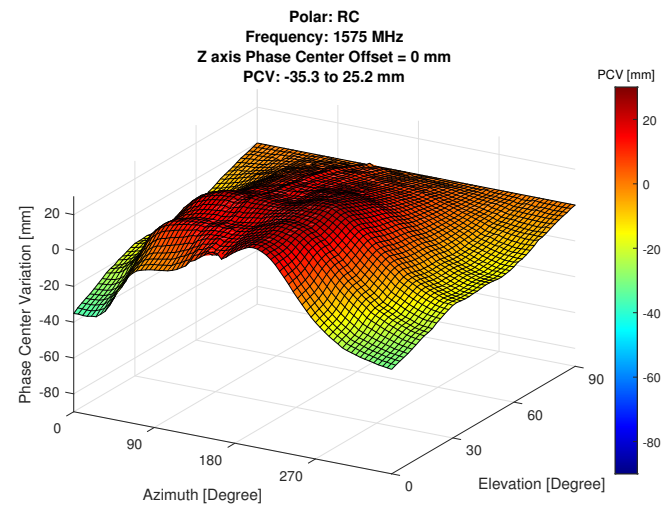
Moreover, the maximum phase variation occurs along the horizontal plane (Elevation = 0°) when the antenna is directed towards the front direction (Azimuth = 0°) and it is directed toward the rear direction (Azimuth = 180°). It is evident that the compact shark-fin antenna structure and on-vehicle scenarios have a large influence on the antenna's phase performance. The theory behind the phenomenon is that when the antenna is mounted on the roof, the ground connection dose not provide a perfect electrical wall boundary condition as described in Eq.(3.1a) and

Eq.(3.1b). Consequently, the patch's electrical field gets disrupted by the repulsive mechanism at its bottom of the patch, leading to a lack of dominance of charge concentration through attractive mechanism at its top layer [41]. Similar to what explained in subsection 3.1.1, this degradation in antenna phase performance can be attributed to field distribution issues. Finding a closed-form expression for finite ground with imperfect boundary conditions in mathematical works proves challenging; however, electromagnetic software can offer simulation results as guidelines. To accurately evaluate high precision automotive GNSS antenna phase center performance during design work, real vehicle testing becomes essential.

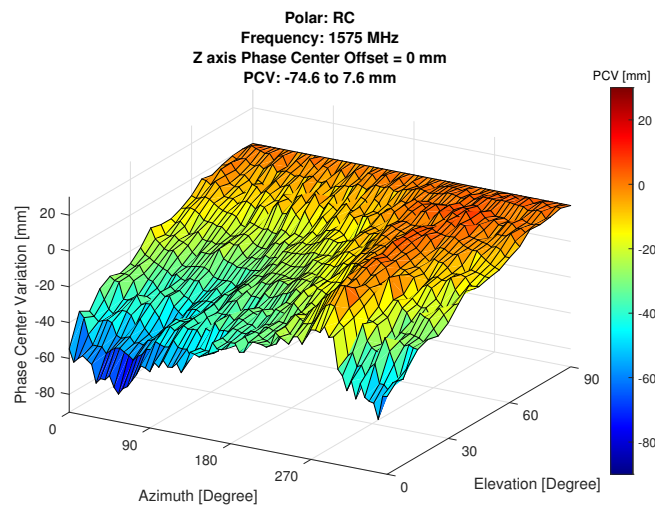
Figure 3.19(a), (b), (c) and Figure 3.20(a), (b), (c) show the center frequency of L2 (1227 MHz) and L5 (1176 MHz) PCV results for the GNSS triple-band patch on an EVB with a 250 mm circular ground plane, on a shark-fin antenna with a 250 mm circular ground plane, and on a shark-fin antenna mounted on a vehicle roof, respectively. The conclusion is almost identical to that of the L1 band's result. The EVB antenna exhibits the best PCV performance, while the on-vehicle testing points to the worst PCV in the real-world scenario. So far, $Phase(\theta, \varphi)$ determination has been based on far-field measurements conducted indoor range of Figure 3.18(a),(b), Figure 3.19(a),(b) and Figure 3.20(a),(b) and outdoor range of Figure 3.18(c), Figure 3.19(c) and Figure 3.20(c). In order to calculate the $PCO(x, y, z)$, one can use the pseudo-inverse formula.



(a) PCV at 1575 MHz @EVb;

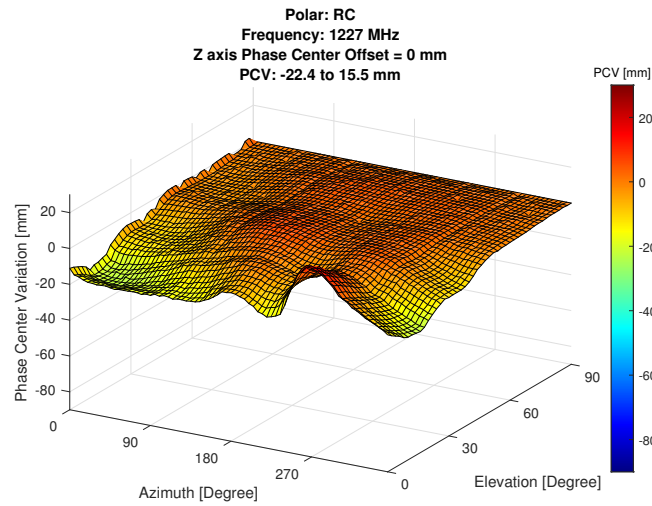


(b) PCV at 1575 MHz @Shark-fin antenna;

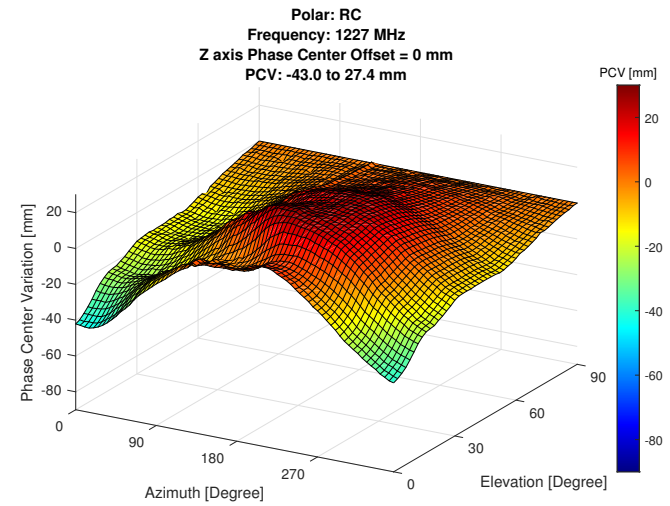


(c) PCV at 1575 MHz @On-vehicle;

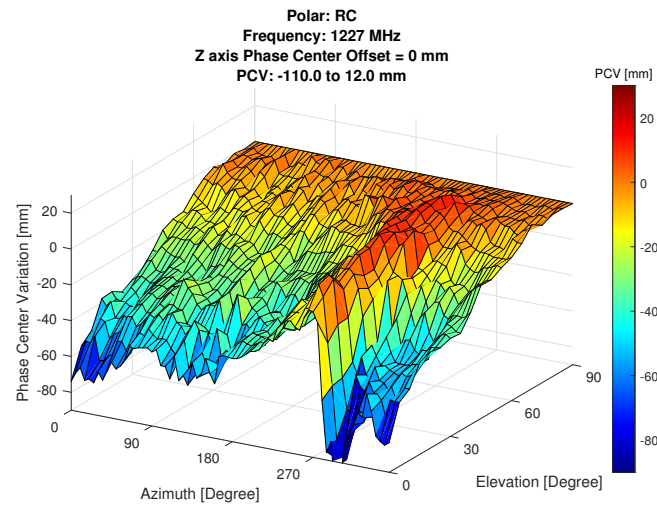
Figure 3.18: PCV at 1575 MHz in L1 band



(a) PCV at 1227 MHz @EVb;

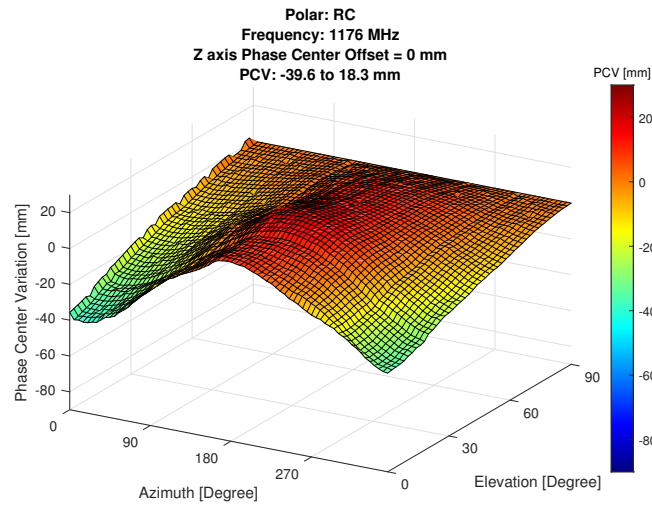


(b) PCV at 1227 MHz @Shark-fin antenna;

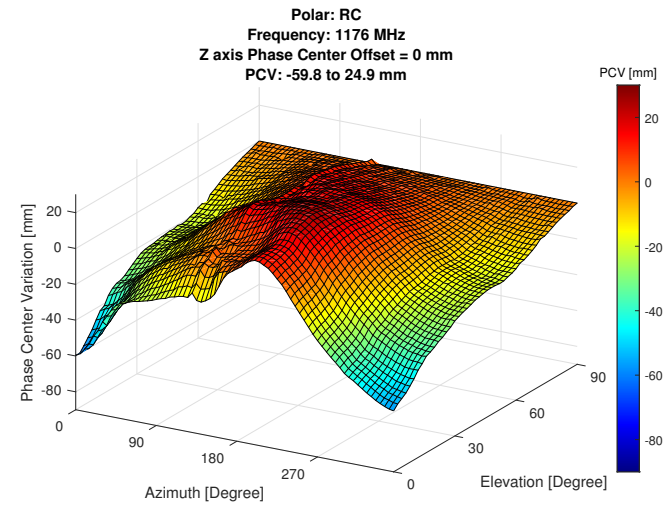


(c) PCV at 1227 MHz @On-vehicle;

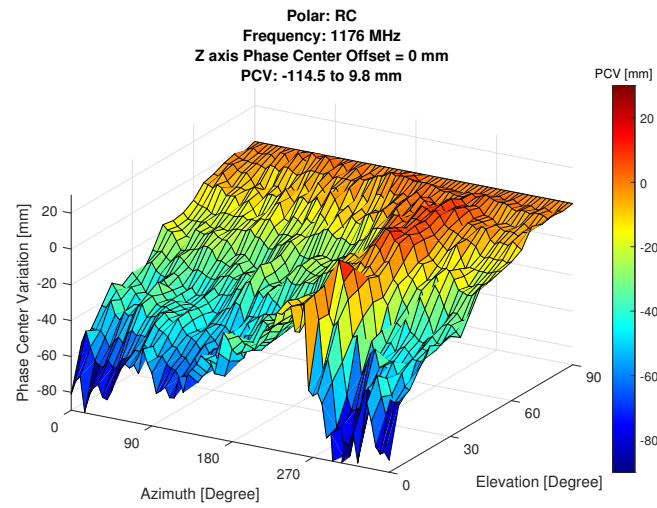
Figure 3.19: PCV at 1227 MHz in L2 band



(a) PCV at 1176 MHz @EVb;



(b) PCV at 1176 MHz @Shark-fin antenna;



(c) PCV at 1176 MHz @On-vehicle;

Figure 3.20: PCV at 1176 MHz in L5 band

$PCO(x, y, z)$ is represented by each point corresponding to an averaged azimuth ($\varphi_N : 0 \leq \varphi \leq 360^\circ$) over each elevation level $PCO_{\theta_{M(x,y,z)}}$. As an example in EVB, Figure 3.18 (a) shows the PCO result in 3D view in Figure 3.21. In this figure, the blue dots represent $PCO(x, y, z)$ at different elevations. Most of points are concentrated around the mean value, but at low elevation positions, the result are separated from the others. The reason for this is that when the antenna receives the incoming signal far below the cut-off Brewster angle, it is unable to receive most of the horizontal component of the circularly polarized waves and appears to be receiving vertically polarized waves instead. This further proves that poor PCO correlated with high axial ratio (as shown in Figure 3.10, Figure 3.11 and Figure 3.12). Moreover, there is low gain at these low elevations and the AR is not good enough to accurately distinguish correct polarization. In post-processing, data from low elevation would be excluded from GNSS receiver using an elevation mask as a well-known practice. The remaining test scenarios can be analyzed in a similar manner.

In automotive navigation, the focus is on the XY plane, also known as the horizontal plane. Figure 3.22(a) shows only on the XY cut view from the three-dimensional result in Figure 3.21. A common way to evaluate the PCO precision is through Circular Error Probable (CEP), *CEP50 is defined as the radius of a circle centered on the mean that is expected to include 50% of landing points.* CEP68 (1σ) and CEP95 (2σ) have a similar definitions. In the same example, EVB at L1 1575 MHz results in a two-dimensional view in Figure 3.22(a), CEP50 is 1.4 mm. The meaning behind this number is that without any phase correction, 50% of incident waves' phase centers will fall within a range of 1.9 mm on the XY plane (horizontal plane). Similar results were observed for CEP68 (1.7 mm) and CEP95 (2.6 mm).

$$PCO \begin{bmatrix} x \\ y \\ z \end{bmatrix} = \frac{\lambda}{2\pi} \cdot \begin{bmatrix} \sin \theta_1 \cos \varphi_1 & \sin \theta_1 \sin \varphi_1 & \cos \theta_1 \\ \sin \theta_1 \cos \varphi_2 & \sin \theta_1 \sin \varphi_2 & \cos \theta_1 \\ \vdots & \vdots & \vdots \\ \sin \theta_1 \cos \varphi_N & \sin \theta_1 \sin \varphi_N & \cos \theta_1 \\ \sin \theta_2 \cos \varphi_1 & \sin \theta_2 \sin \varphi_1 & \cos \theta_2 \\ \sin \theta_2 \cos \varphi_2 & \sin \theta_2 \sin \varphi_2 & \cos \theta_2 \\ \vdots & \vdots & \vdots \\ \sin \theta_M \cos \varphi_{N-1} & \sin \theta_M \sin \varphi_{N-1} & \cos \theta_M \\ \sin \theta_M \cos \varphi_N & \sin \theta_M \sin \varphi_N & \cos \theta_M \end{bmatrix}^{-1} \cdot \begin{bmatrix} \text{Phase } (\theta_1, \varphi_1) \\ \text{Phase } (\theta_1, \varphi_2) \\ \vdots \\ \text{Phase } (\theta_1, \varphi_N) \\ \text{Phase } (\theta_2, \varphi_1) \\ \text{Phase } (\theta_2, \varphi_2) \\ \vdots \\ \text{Phase } (\theta_M, \varphi_{N-1}) \\ \text{Phase } (\theta_M, \varphi_N) \end{bmatrix} \quad (3.7)$$

where $0 \leq \theta \leq 90^\circ$, $0 \leq \varphi \leq 360^\circ$.

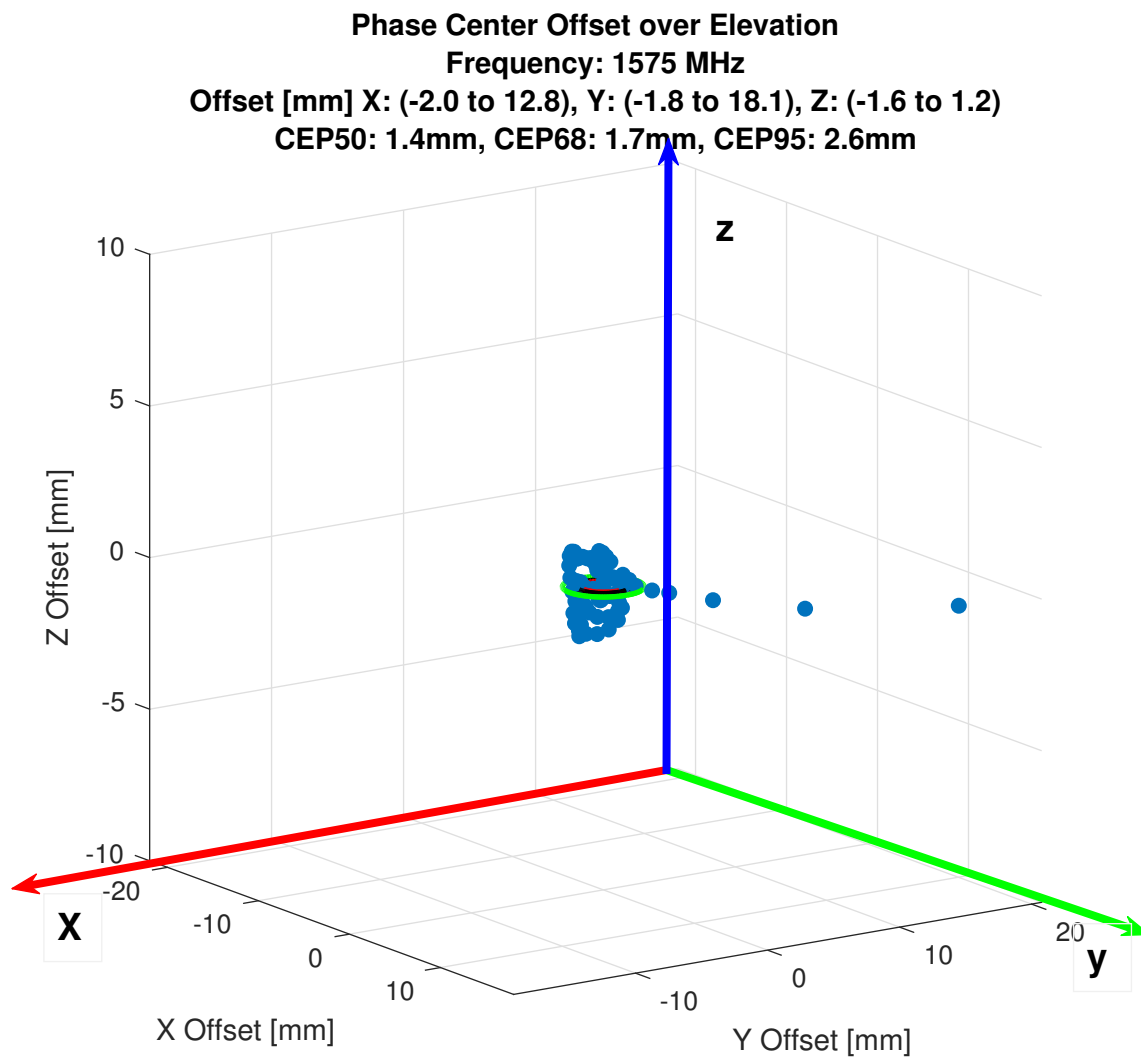
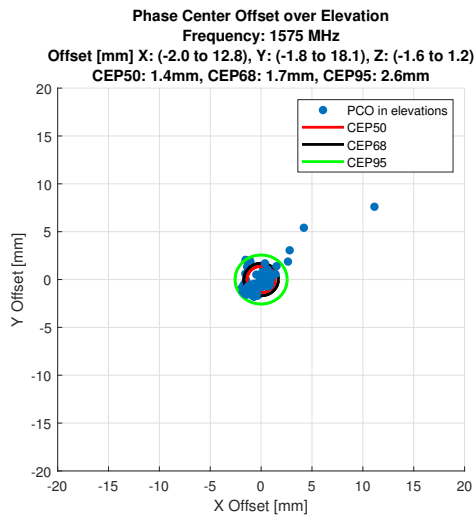
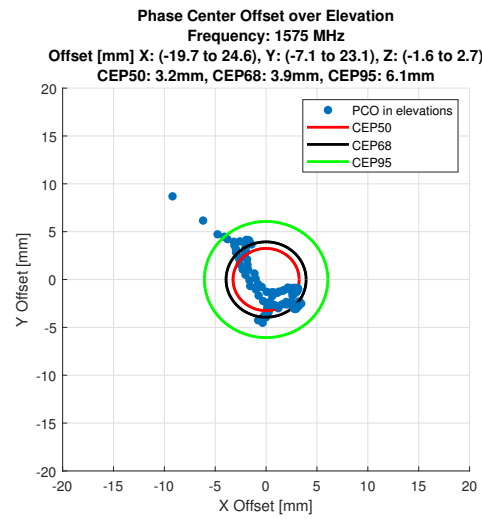


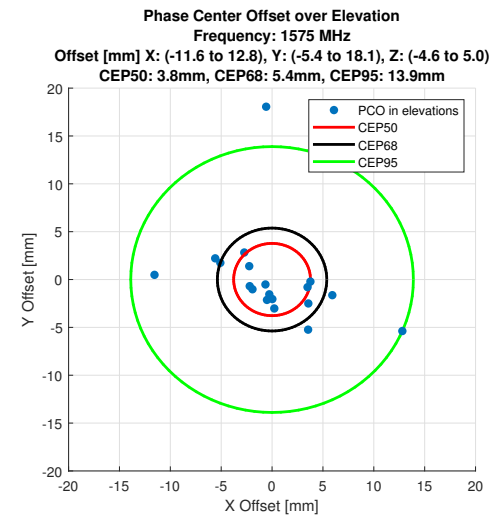
Figure 3.21: L1 band PCO in 3D view



(a) PCO at 1575 MHz @EVb;

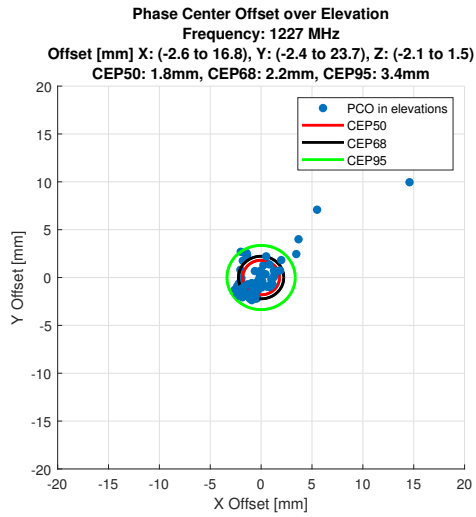


(b) PCO at 1575 MHz @Shark-fin;

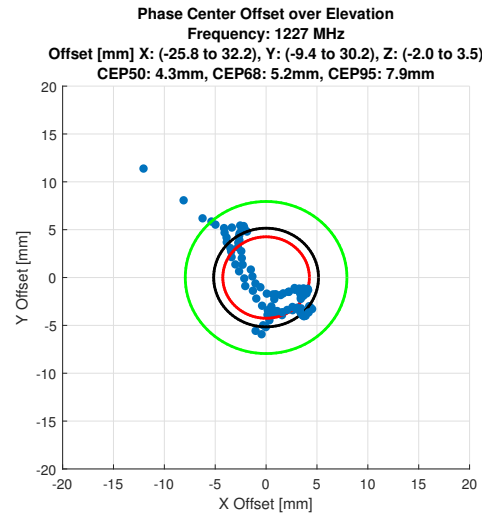


(c) PCO at 1575 MHz @On-vehicle;

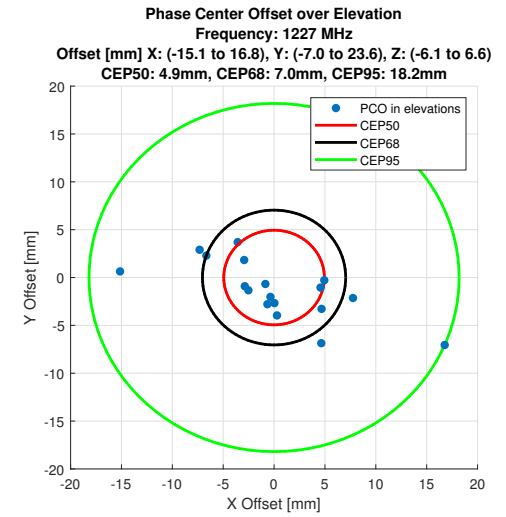
Figure 3.22: PCO at 1575 MHz in L1 band



(a) PCO at 1227 MHz @EVb;

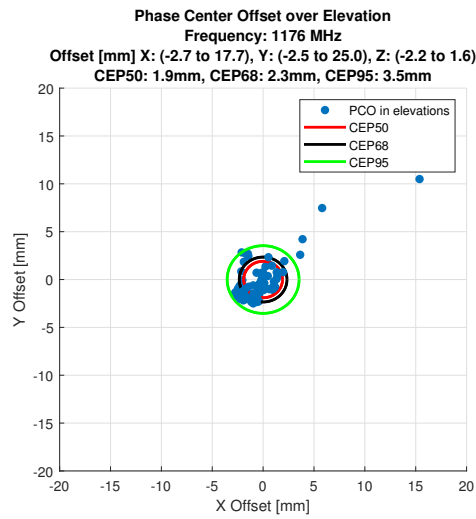


(b) PCO at 1227 MHz @Shark-fin;

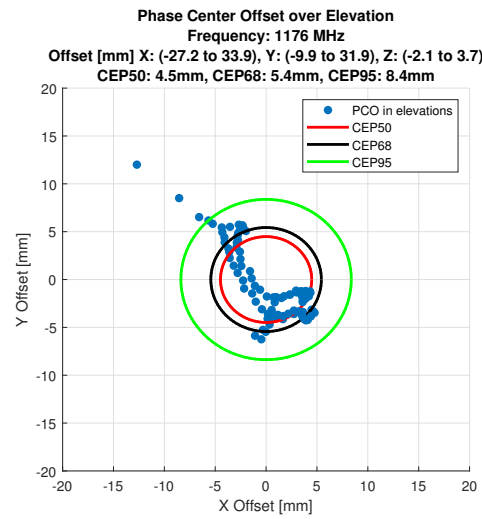


(c) PCO at 1227 MHz @On-vehicle;

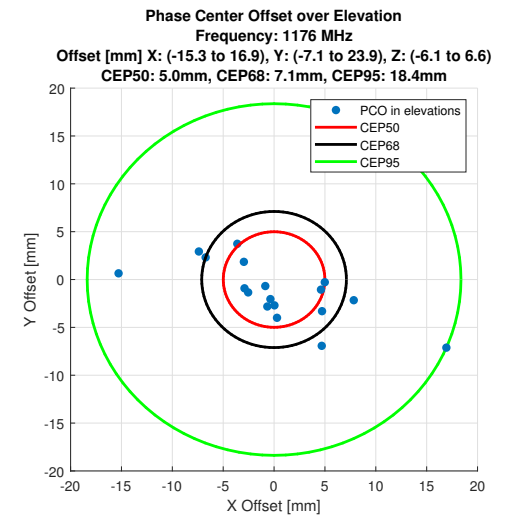
Figure 3.23: PCO at 1227 MHz in L2 band



(a) PCO at 1176 MHz @EVb;



(b) PCO at 1176 MHz @Shark-fin;



(c) PCO at 1176 MHz @On-vehicle;

Figure 3.24: PCO at 1176 MHz in L5 band

By using PCV shark-fin result from Figure 3.18(b), a similar two-dimensional XY plane can be calculated in Figure 3.22(b), which has larger values for PCO50 (3.2 mm), CEP86 (3.9 mm) and CEP95 (6.1 mm). Meanwhile, the PCO resulting from an antenna on-vehicle is presented in Figure 3.22(c). The on-vehicle result shows degradation (CEP50: 3.8 mm) compared to a standalone antenna on an EVB.

Similar results are found for L2 and L5 frequency band, as shown in Figures 3.23(a), (b) and (c) and Figure 3.24(a), (b) and (c) in the same three test scenarios. The PCV range and PCO's CEP summarized results are tabulated in the Table 3.2.

In general, the same patch antenna has the best performance of PCV and PCO on a 250 mm circular ground plane EVB. However, when integrated into a compact shark-fin housing or installed on a car roof, its phase performance degraded due to changes in magnetic wall boundary conditions caused by alterations to the antenna ground connection via an underneath PCB and shark-fin antenna chassis mounted with limited size fasteners that could disturb magnetic wall Eq.(3.1a) and Eq.(3.1b). Furthermore, complex elements surrounding the patch antenna slot inside its radome broke magnetic current density resulting in impacts on gain, axial ratio and phase performance of this type of antennas which has strong connections with their surrounding environments.

3.2 GNSS L1/L2 Testing with MDOT CORS Network

In the last section, it describes the radiation performance of a multi-band GNSS antenna (L1/L2/L5), including passive gain, axial ratio, phase center variation and phase center offset. In this section, a real field measurement is conducted to compare the GNSS code pseudorange and carrier phase calculation in

- L1 code pseudorange in ultra-short baseline;

Table 3.2: PCV and PCO' CEP

L1: 1575 MHz			
	EVB	Shark-fin	On-vehicle
PCV [mm]	-28.4 to 14.1	-35.3 to 25.2	-74.6 to 7.6
PCO			
CEP50 [mm]	1.4	3.2	3.8
CEP68 [mm]	1.7	3.9	5.4
CEP95 [mm]	2.6	6.1	13.9
L2: 1227 MHz			
	EVB	Shark-fin	On-vehicle
PCV [mm]	-22.4 to 15.5	-43.0 to 27.4	-110 to 12.0
PCO			
CEP50 [mm]	1.8	4.3	4.9
CEP68 [mm]	2.2	5.2	7.0
CEP95 [mm]	3.4	7.9	18.2
L5: 1176 MHz			
	EVB	Shark-fin	On-vehicle
PCV [mm]	-39.6 to 18.3	-59.8 to 24.9	-114.6 to 9.8
PCO			
CEP50 [mm]	1.9	4.5	5.0
CEP68 [mm]	2.3	5.4	7.1
CEP95 [mm]	3.5	8.4	18.4

- L1/L2 code pseudorange in ultra-short baseline;
- L1 carrier phase result with RTK CORS network correction in ultra-short / short baseline;
- L1/L2 carrier phase result with RTK CORS network correction in ultra-short / short baseline;
- L1/L2 PPP result

Table 3.3: CORS network status

	Site Status	Network Processing
Green	Connected - receive data	Fixed
Blue	Connected - receive data	Not Fixed
Red	Disconnected	Not Utilized

3.2.1 RTK MDOT CORS Network

A Continuously Operating Reference Stations (CORS) GNSS Network is established by the National Oceanic and Atmospheric Administration's (NOAA's) National Geodetic Survey (NGS) [46]. It provides a regional positioning enhanced service. In the late 1980s, the NGS recognized the potential contribution of GPS on civilian user in determine geodetic latitude, longitude and height [47]. In Michigan state, the CORS network is operated by Michigan Department of Transportation (MDOT). There are hundreds of reference stations spread across the northern and southern parts of state. The live status of these reference stations can be found on Site Status (<https://mdotcors.michigan.gov/sbc/User/SiteMap/SiteMap>) as shown in Figure 3.25. The color coding of the site network indicates explanations for each status listed in Table 3.3.

The closest observation base station, named METR-0017, is located in Auburn Hills by I-75 highway, in the Figure 3.26. Latitude: 42.68472° N ($42^{\circ}41'05''$ N), Longitude: 83.24278° W ($83^{\circ}14'34''$ W). The receiver manufacturer is Leica Geosystems AG, type LEICA GR50 and the antenna manufacturer type is LEIAR20 LEIM. An on-site photograph can be seen in Figure 3.27.

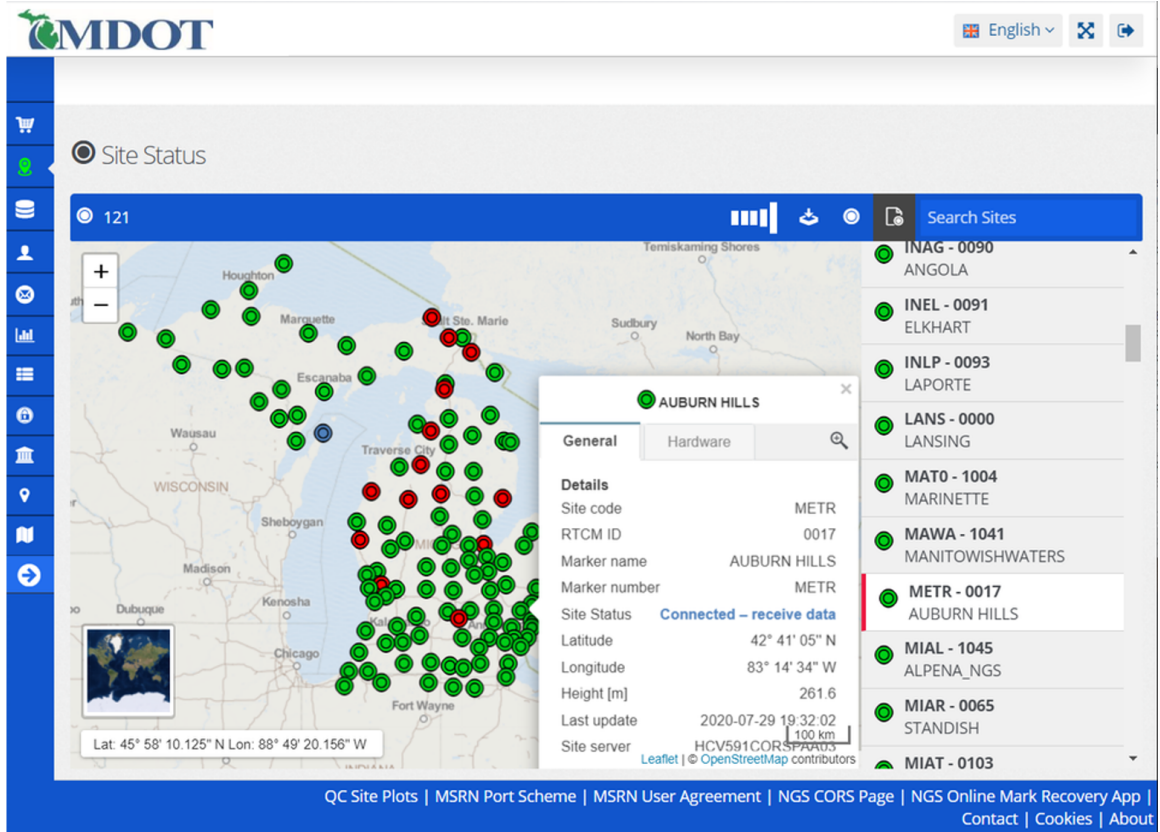


Figure 3.25: CORS reference stations are operated by MDOT

3.2.2 Ultra-short / short Baseline

The distance between the reference station and the user antenna is referred to as the baseline. The baseline distance is determined by the ionospheric delay I and tropospheric delay T in Eq.(2.21) and Eq.(2.22). In the field test, there are two baseline scenarios (ultra-short baseline of Figure 3.28 and short baseline Figure 3.29) and configurations listed in Table 3.4,

- ultra-short baseline: nearby MERT reference station is 40 meter (133 ft);

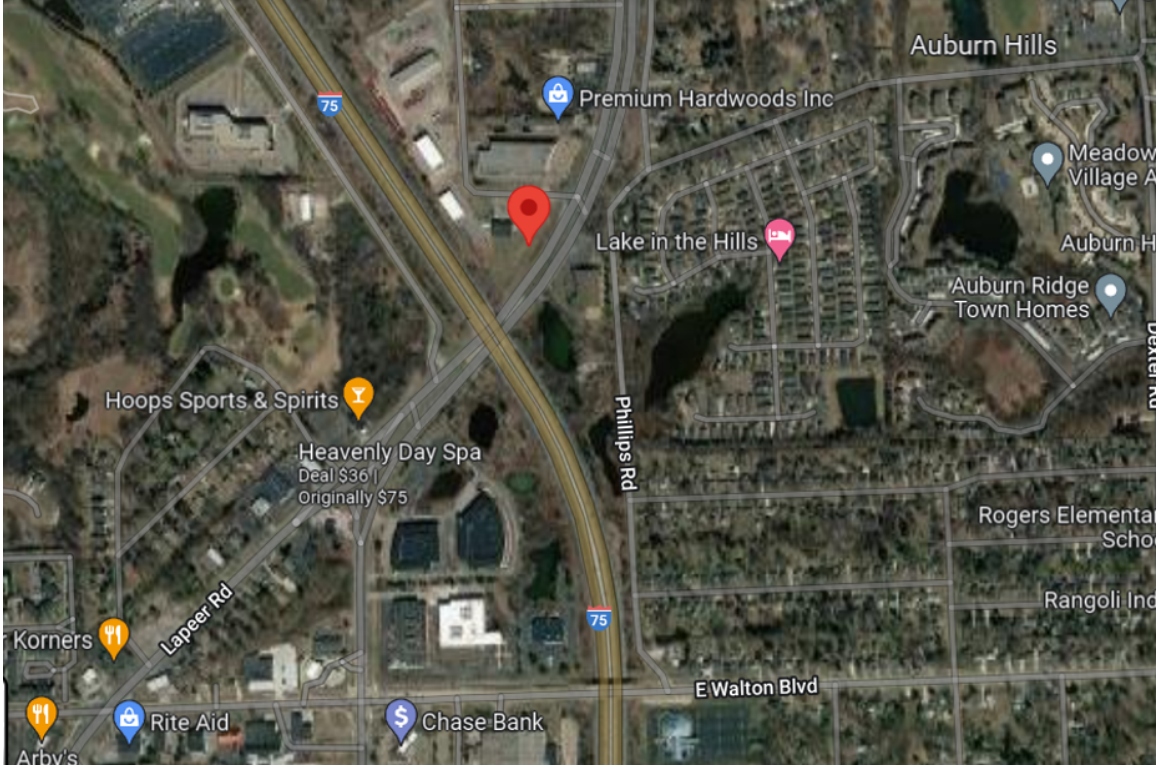


Figure 3.26: The nearest observation station is METR-0017 on Google Map

- short baseline: the direct distance between an open-field testing ground and MERT reference station is 9.35 km (5.8 miles).

3.2.3 RTKLIB

RTKLIB, an open-source GNSS program package for standard and precise positioning [26], consists of a portable program library and several Application Programs (APs). The software is written by researcher and professor Tomoji Takasu of Tokyo University of Marine Science and Technology. It supports standard and precise positioning algorithms with: GPS, GLONASS, Galileo, QZSS, Beidou and



Figure 3.27: Reference observation station setup with receiver and antenna

Table 3.4: Two test scenarios ultra-short and short baseline

	40 m Ultra-short Baseline	9.35 km Short Baseline
Reference position	42.68516° N, 83.242831°W	42.68516° N, 83.242831°W
Test date	2020/07/04	2020/08/08
Test GPS time	14:28:46 to 21:34:33	00:46:57 to 15:45:42
Weather	Cloudy	Clear sky
Sampling rates	1 sample / second	1 sample / second
Total sampling counts	25,439	52,297



Figure 3.28: Ultra-short baseline nearby MERT observation station

SBAS in various positioning modes for both real-time and post-processing. The main functions included in the RTKLIB are as follows,

- AP Launcher: RTKLAINCH
- Real-time Positioning: RTKNAVI
- Communication Server: STRSVR
- Post-processing Analysis: RTKPOST
- RINEX Converter: RTKCONV
- Plot Solutions and Observation Data: RTKPLOT
- Downloader of GNSS Data: RTKGET
- NTRIP Browser: SRCTBLBROWS

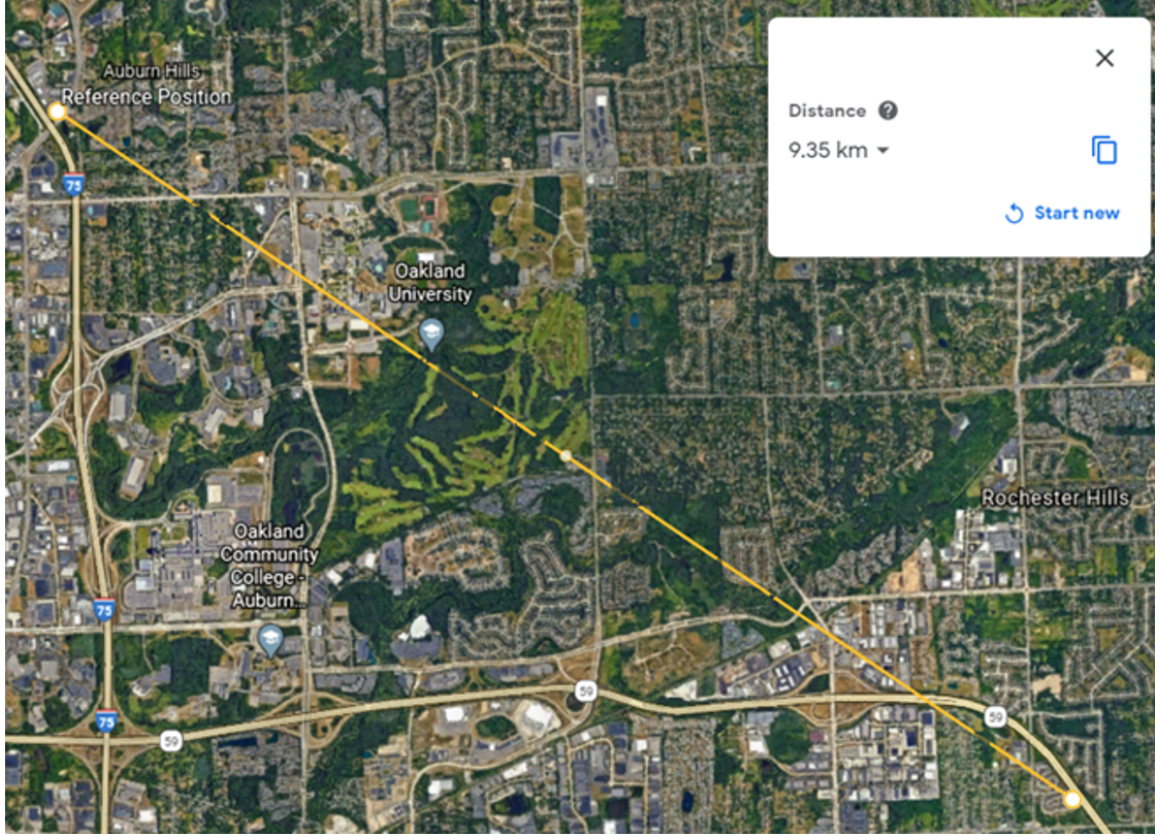


Figure 3.29: Short baseline, the direct distance between an open-field of testing ground and METR observation station

3.2.4 User Localization Devices Setup

The performance of a shark-fin antenna is introduced in the previous section 3.1 through component level testing in an antenna range, on-vehicle far-field measurement and field baseline measurement. The same antenna is mounted on the car roof (Figure 3.30) and the raw data is collected from u-blox[®] C100-F9K application board [48].



Figure 3.30: User shark-fin GNSS L1/L2/L5 band antenna setup

The u-blox C100-F9K evaluation kit includes an application board F9K with a reference L1/L2 magnetic mounted antenna [48]. The receiver setup is shown in Figure 3.31, and the application board requires vehicle reference points for Inertial Measurement Unit (IMU). In the setup, the application boards are fixed in position above the rear trunk using Hook & Loop tape.

- Working frequency bands L1 (1559-1610 MHz) / L2 (1215-1239 MHz).
- Supported protocols: UBX, NMEA 4.10 and RTCM 3.3
- Differential GNSS (DGNSS).
- Raw carrier-phase measurements and data output with UBX protocols, UBX-RXM-RAWX message.
- Automotive dead reckoning.



Figure 3.31: u-blox receivers setup on-vehicle

3.2.5 Static On-vehicle Testing Result

The testing results are based on the above u-blox C100-F9K application board, which exports raw data with UBX protocol. By using RTKLIB and its RINEX converter tool RTKCONV, the UBX protocol could be converted to standard RINEX v3.03 format. In order to distinguish the different scenarios, various settings are converted into RINEX configurations that include L1 single frequency band and L1/L2 dual frequency band observation file (.obs) and navigation file (.nav).

3.2.5.1 Code Pseudorange Measurement Result of L1 and L1/L2 in Ultra-short

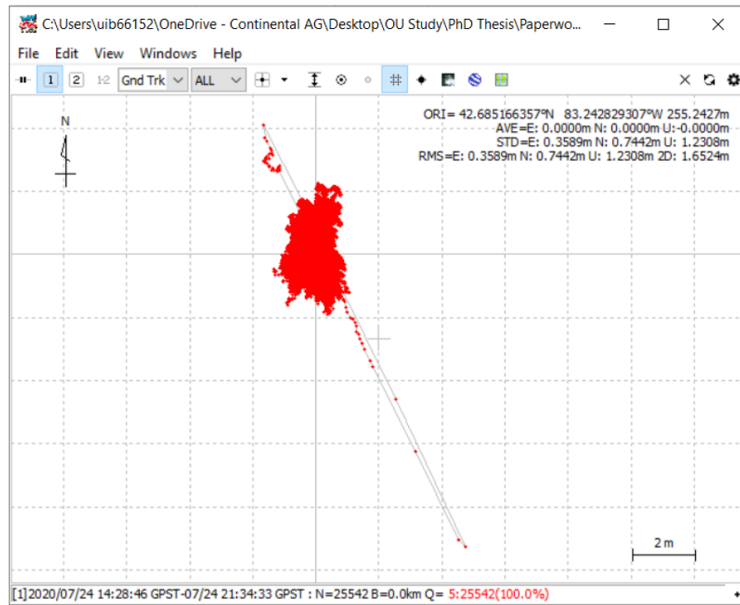
Baseline The single L1 code pseudorange result and the L1/L2 dual-band code differential GNSS pseudorange results are simultaneously converted from the same raw data during an ultra-short baseline testing with over 7 hours of observation

(Figure 3.28). The ground tracking (Gnd Trk) results are shown in Figure 3.32(a) and (b), respectively, with a scale of 2 m.

Comparing the results output, the L1 single-band analysis shows a 2D RMS of 1.6524 m, while the L1/L2 dual-band analysis shows a 2D RMS of 0.7149 m. From the single L1 result, there are sudden position changes during the observation. The position plot view (Figure 3.33) displays both Data1's L1 code single GNSS and Data2's L1/L2 code GDNSS together. For the 2D analysis, more focus is given to panel East-West (E-W) and panel North-South (N-S) directional results. There is a significance position change on Data1's code single GNSS at time 19:03, with E-W variants from -2 to +4 meters and N-S variants ranging from -6 to +4 meters. However, when using L1/L2 dual-band DGNSS, there is no major impact on any direction. This sudden change in position may be caused by environmental changes such as ionospheric scintillation and tropospheric differences; however, it has a greater impact on the single L1 band compared to the L1/L2 dual band which experiences much less impact due to its frequency dependence discussed in Fundamental Section Eq.(2.61). By utilizing the differential algorithm of using both L1/L2 dual frequency bands, this error can be dramatically reduced.

Post processing setup:

- Elevation mask angle: 10 °
- Receiver dynamics: OFF
- Earth tides correction: ON
- Ionosphere correction: Broadcast
- Troposphere correction: Saastamoinen model
- Satellite ephemeris clock: Broadcast ephemeris

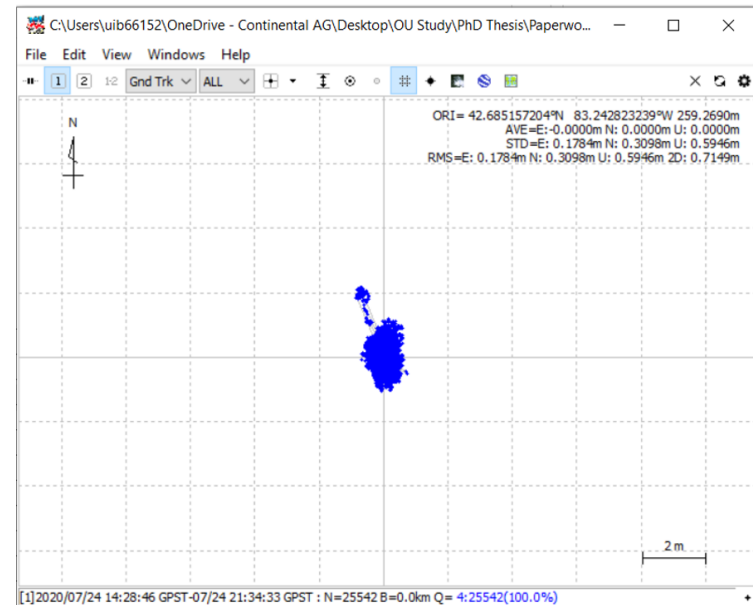


(a)

1D RMS = E:0.3589m, N:0.7442m, U: 1.2308m

2D RMS = 1.6524m

Q=5 Single: 25542 (100%)



(b)

1D RMS=E: 0.1784m, N: 0.3098m, U: 0.5946m

2D RMS = 0.7149m

Q=4 DGNSS: 25542 (100%)

Figure 3.32: (a) L1 code single GNSS, (b) L1/L2 code DGNSS measurement in Gnd Trk view

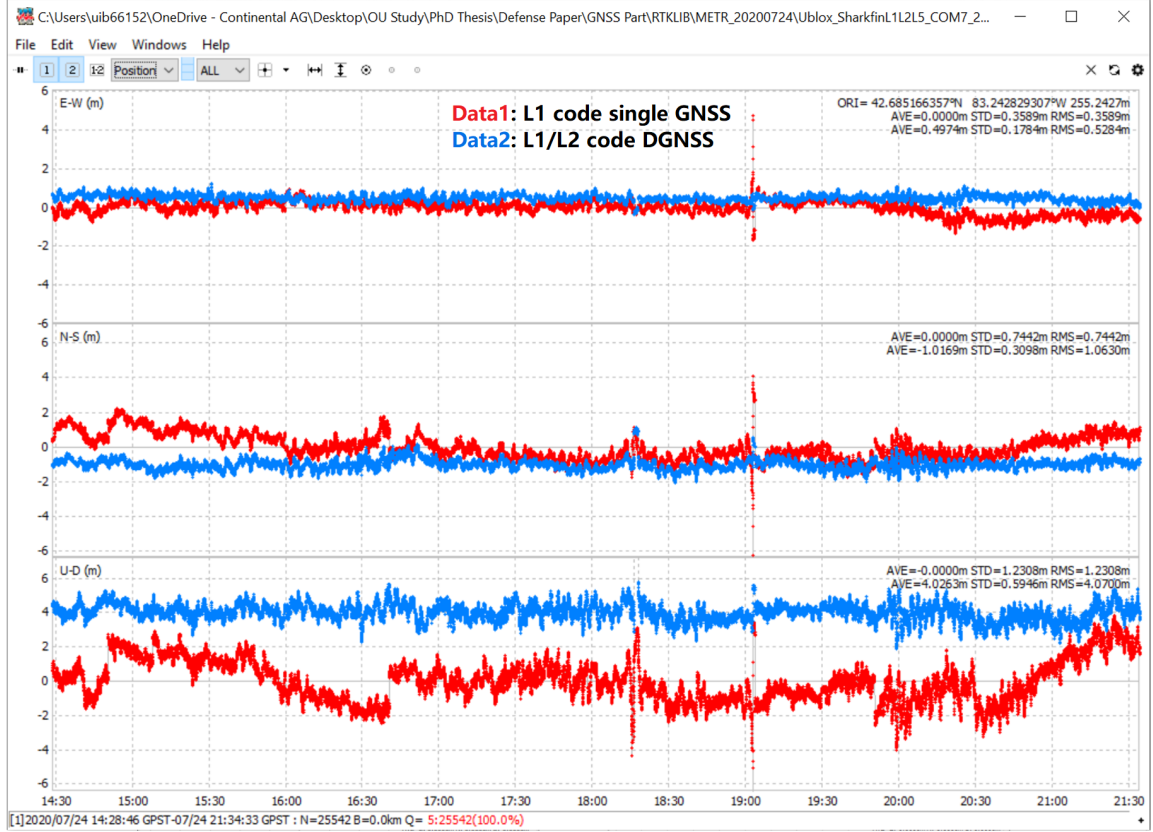


Figure 3.33: Position plot including Data1 L1 code single GNSS and Data2 L1/L2 code DGNSS

The above results shown in this section, the position calculation method is based on code pseudorange. From 2D RMS results, L1/L2 dual-band DGNSS method (2D RMS = 0.7149 m) accuracy improves comparing with L1 single band method (2D RMS = 1.6524 m). However, both results are in the meter-level range.

3.2.5.2 Carrier Phase Measurement result of L1 and L1/L2 in Ultra-short Baseline

The position accuracy in Section 2.4.2 can be significantly improved by the carrier phase method Eq.(2.22) compared to the code pseudorange method Eq.(2.21). The

results shown in Figure 3.34 include a green fixed solutions and orange float solutions, which represent the integer and unconstrained solutions of the least-squares problem Eq.(2.55). In this case, a validation threshold of $R_{thres} = 3$ is selected for the ratio-test.

Post-processing setup

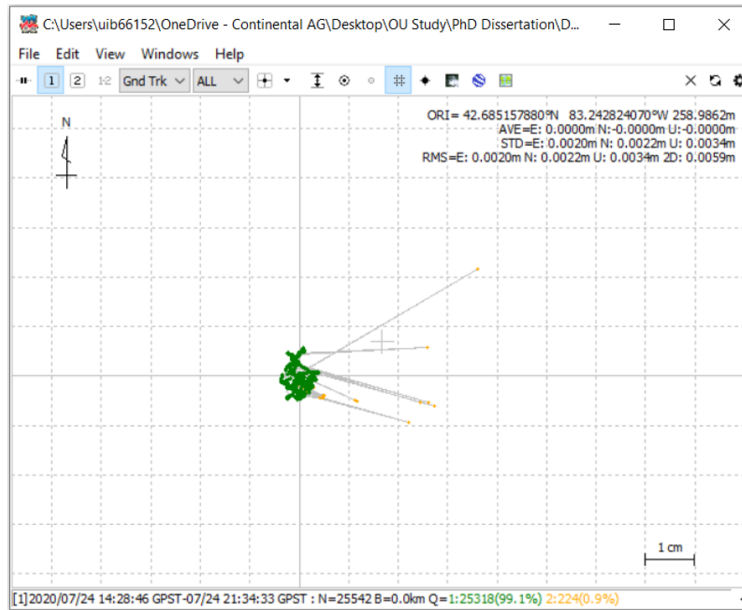
- Elevation mask angle: 10°
- Receiver dynamics: OFF
- Earth tides correction: ON
- Ionosphere correction: Broadcast
- Integer ambiguity resolution options: Fix and Hold
- Integer ambiguity validation option: 3

As shown in Figure 3.34(a), the scale is 1 cm. In the case of a single L1 frequency band, there are 224 float solutions, which account for 0.9% of the total solutions. In the result summary, the 2D RMS is 0.0059 m. Comparing it with the result of single L1 band code pseudorange (2D RMS = 1.6524 m), we can see that carrier phase accuracy has significantly improved compared to code pseudorange results. When applying post-processing algorithm to L1/L2 dual frequency band data, as shown in Figure 3.34(b), there are only 32 float solutions out of a total of 0.1% solutions available. Furthermore, the improvement in terms of 2D RMS will be to reach up to 0.0035 m.

The carrier phase method (Figure 3.34) clearly demonstrates a significant improvement in 2D RMS positioning accuracy compared to the code pseudorange (Figure 3.35). Residuals are an important parameter for evaluating the predicted

values and real measurement results. From the top and middle panels of Figure 3.35, it can be observed that the code pseudorange residuals RMS is 0.580 m, whereas the carrier phase residuals RMS is only 0.007 m, indicating an increase of three orders of magnitude.

The carrier phase measurement (Figure 3.35) shows good accuracy for both results. However, in Figure 3.35(a), the single L1 band result indicates that at the beginning of the measurement, the output solutions are difficult to "fix". It is easy to notice that there are some orange points separated at the bottom of the figure. As illustrated in Figure 3.35(a), due to the convergence of residuals not being ready in the initial stage of L1 single band measurement, the cost function Eq.(2.49) does not meet the threshold for integer ambiguity validation, resulting in only float solutions being present. Meanwhile, during post-processing, satellites' signals from L2 frequency band are added into Extended Kalman Filter (EKF) for positioning estimation. The inclusion of more frequency independent data provides better initialization accuracy and improves convergence time during measurement update process. This is why dual frequency band results can find fixed solutions right from the start and have more fixed solutions compared to single band's data.

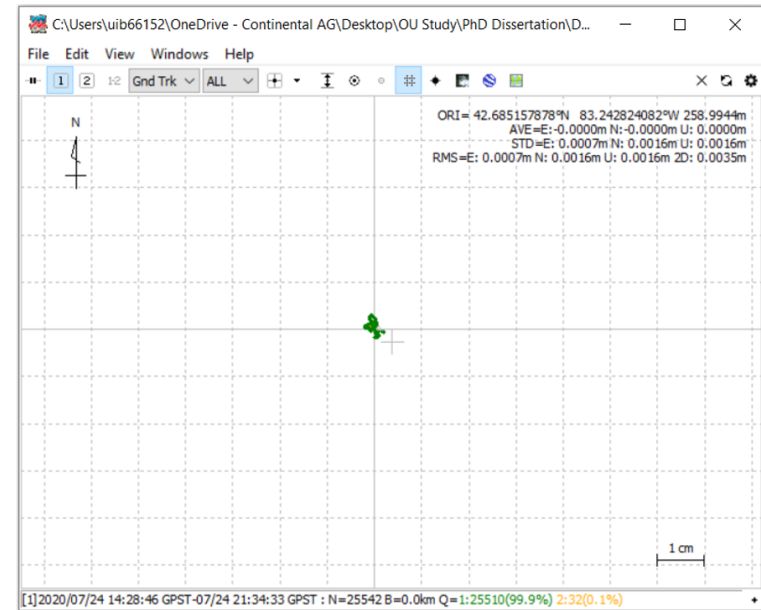


(a)

1D RMS = E: 0.0020m, N:0.0022m, U:0.0034m

2D RMS = 0.0059 m

Q= 1 Fixed:25318(99.1%) 2 Float:224(0.9%)



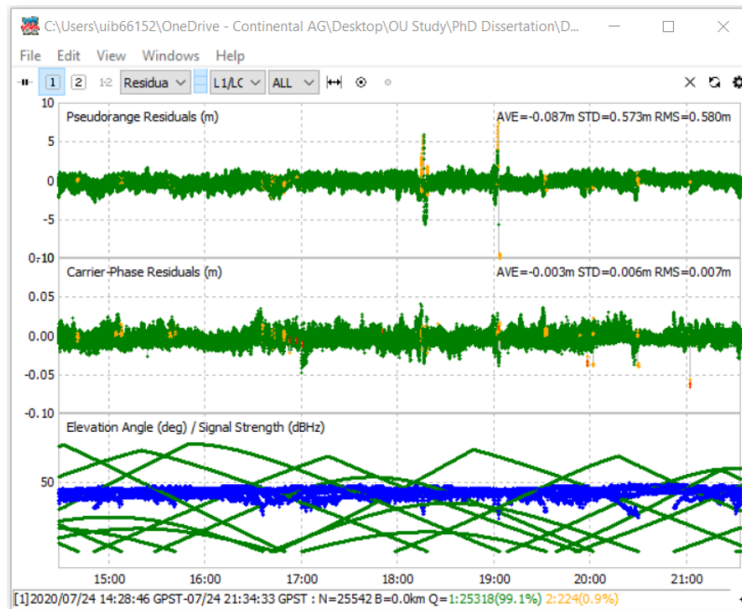
(b)

1D RMS = E:0.0007m, N:0.0016m, U: 0.0016m

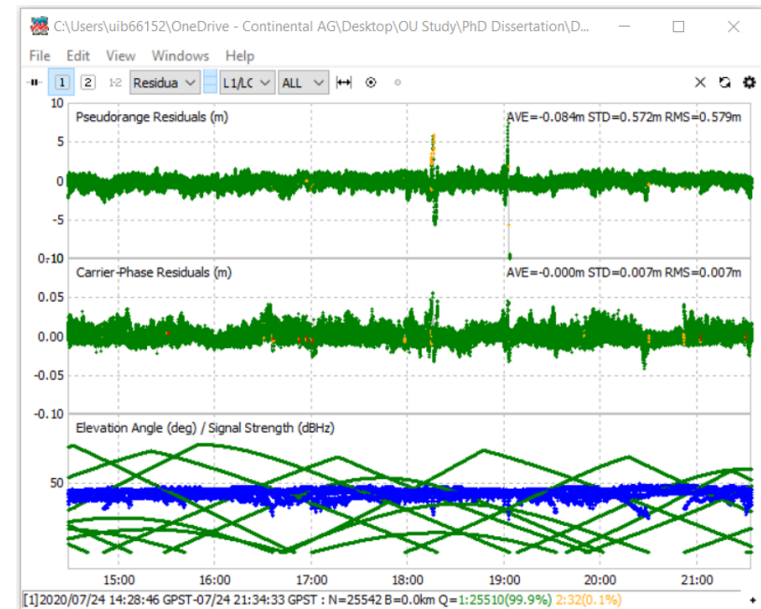
2D RMS = 0.0035 m

Q= 1 Fixed:25510(99.9%) 2 Float:32(0.1%)

Figure 3.34: (a) Carrier phase L1 static result; (b) Carrier phase L1/L2 static result



(a) L1 Residuals



(b) L1/L2 Residuals

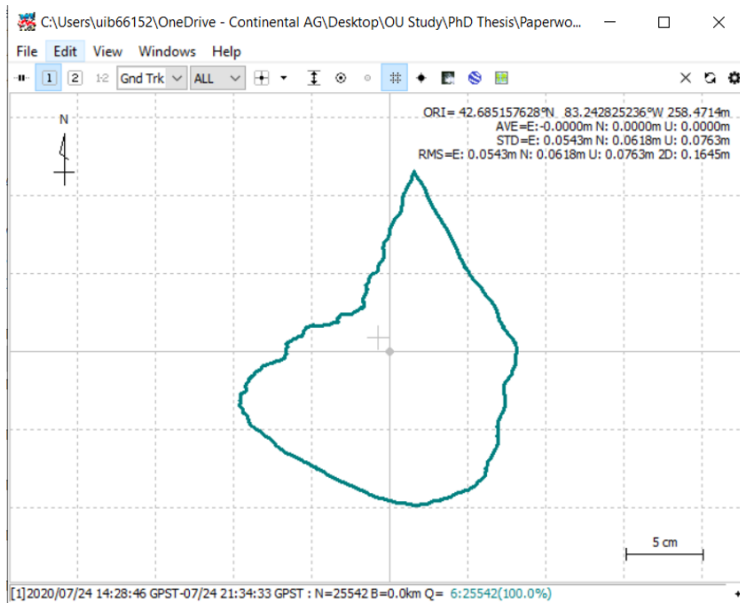
Figure 3.35: (a) Carrier phase L1 band residuals; (b) Carrier phase L1/L2 dual band residuals

3.2.5.3 PPP L1/L2 Result in Ultra-short Baseline The post-processing correction files are obtained using the Precise Point Positioning (PPP) method introduced in Section 2.4.4. The receiver phase center variation files are imported from the Matlab plot phase GUI tool in next section. The data extracted includes on-vehicle phase center variation results (Figure 3.41) for the L1 frequency band (Figure 3.18(c)) and the L2 frequency band (Figure 3.19(c)).

Post processing setup

- Positioning mode: PPP-static
- Filter type: Smoother combined solution with forward and backward filter solution
- Elevation mask angle: 10°
- Dynamics mode of the rover receiver: OFF
- Earth tides correction: ON
- Ionosphere correction: Broadcast
- Troposphere correction: Saastamoinen mode
- Clock data: igr21155.clk
- Earth rotation parameter file: igr21155.erp
- Precise ephemeris: igr21155.sp3

The results shown in Figure 3.36 represent the PPP method without and with receiver phase center variation correction (Figure 3.41). The 2D RMS results are 0.1645 m and 0.1629 m respectively.

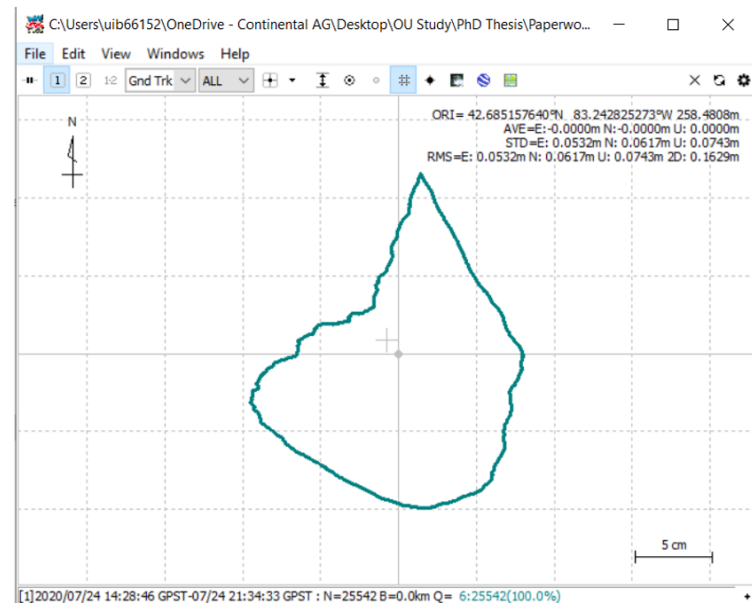


(a)

1D RMS = E: 0.0543m, N:0.0618m, U:0.0763m

2D RMS = 0.1645 m

Q= 6 PPP 25542(100%)



(b)

1D RMS = E:0.0532m, N:0.0617m, U: 0.0743m

2D RMS = 0.1629 m

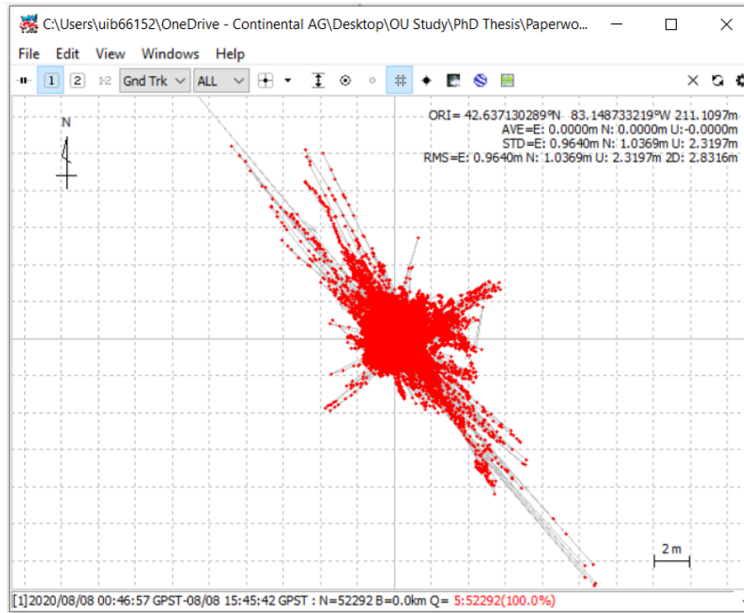
Q= 6 PPP 25542(100%)

Figure 3.36: (a) PPP L1/L2 static result; (b) PPP L1/L2 static result with receiver phase center variation correction

3.2.5.4 Code Pseudorange Single L1 and DGNSS L1/L2 Result in Short Baseline In a short baseline test with over 14 hours of observation (Figure 3.29), the single L1 code pseudorange result and L1/L2 dual-band code differential GNSS pseudorange results were derived from the same raw data, using a short baseline of 9.35 km. The ground tracking (Gnd Trk) results are shown in Figure 3.37(a) and (b), respectively, with a scale of 2 meters. The observation time is twice that of the ultra-short baseline test. As shown in the L1 single band results, ionospheric scintillation has a greater impact on the northwest and southeast directions. Comparing it to the L1/L2 dual-band calculation, ionospheric scintillation affects much less, resulting in an improved 2D RMS accuracy from 2.83 m to 1.13 m. However, both methods still provide meter-level accuracy.

Post processing setup

- Elevation mask angel option: 10°
- Receiver dynamics: off
- Earth Tides Correction: on
- Ionosphere Correction: broadcast
- Troposphere Correction: Saastamoinen model
- Satellite Ephemeris Clock: broadcast ephemeris

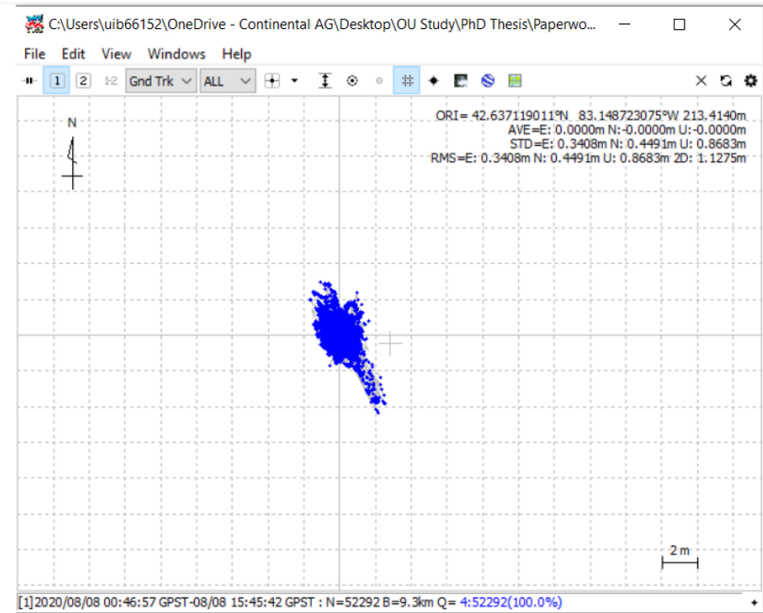


(a)

1D RMS = E:0.9640m, N:1.0369m, U: 2.3197m

2D RMS = 2.8316 m

Q=5 Single: 52297 (100%)



(b)

1D RMS = E:0.3408m, N:0.4491m, U: 0.8683m

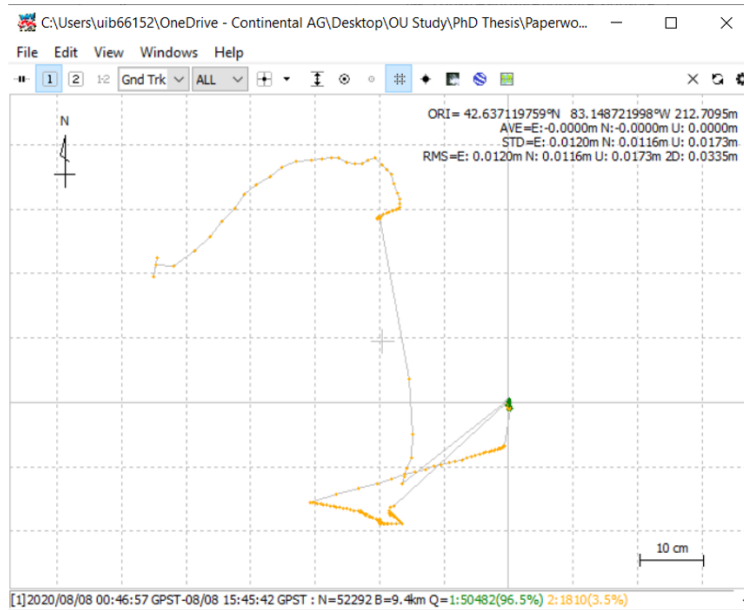
2D RMS = 1.1275 m

Q=4 DGNSS: 52297 (100%)

Figure 3.37: (a) L1 code single GNSS; (b) L1/L2 code DGNSS measurement

3.2.5.5 Carrier Phase Static L1 and L1/L2 Result in Short Baseline The result shown in Figure 3.38 displays the green fixed solutions and orange float solutions, representing the integer and unconstrained solutions of the least-squares problem Eq.(2.49). In this case, a validation threshold of ratio-test is selected $R_{thres} = 3$, post-processing setup

- Elevation mask angel option: 10°
- Receiver dynamics: off
- Earth Tides Correction: on
- Ionosphere Correction: broadcast
- Integer ambiguity resolution options: fix and hold
- Integer ambiguity validation option: 3

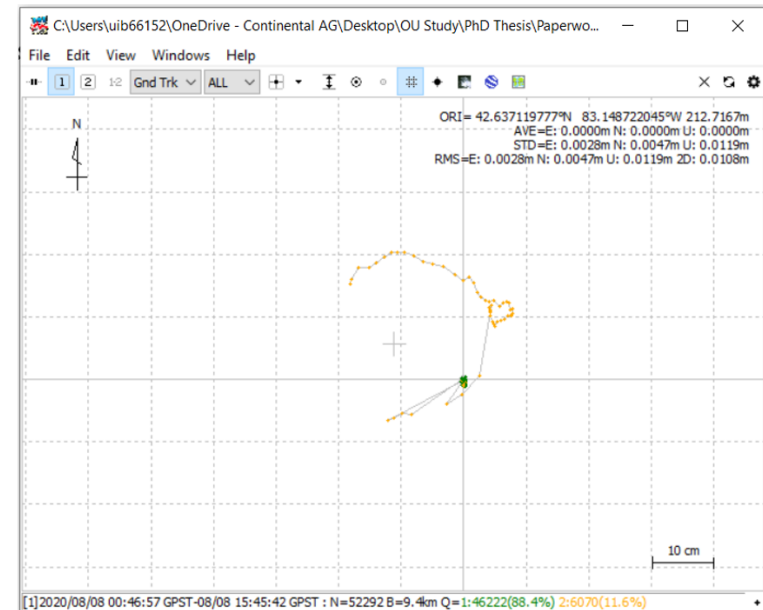


(a)

1D RMS = E:0.0120m, N:0.0116m, U: 0.0173m

2D RMS = 0.0335 m

Q=1 Fixed:50482(96.5%) 2 Float:1810(3.5%)



(b)

1D RMS = E:0.0028m, N:0.0047m, U: 0.0119m

2D RMS = 0.0108 m

Q=1 Fixed:46222(88.4%) 2 Float:6070(11.6%)

Figure 3.38: L1 and L1/L2 carrier phase in short baseline: (a) L1 GNSS; (b) L1/L2 dual-band GNSS

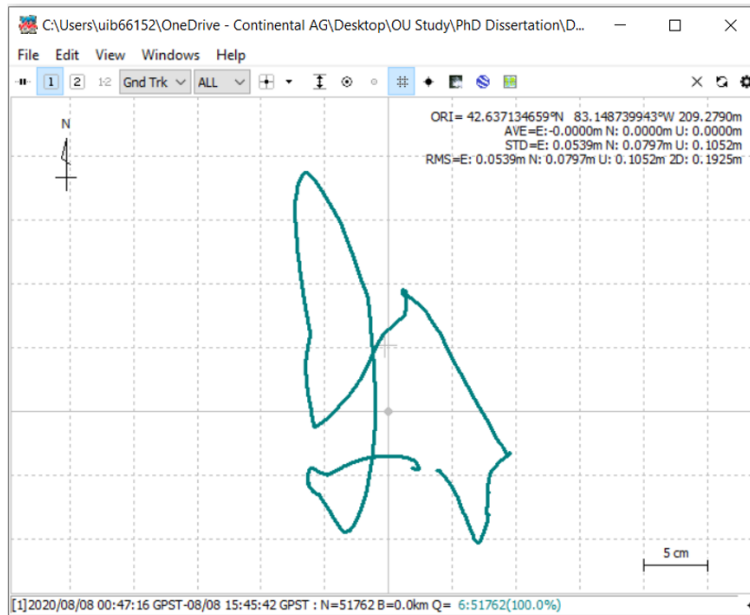
3.2.5.6 PPP Static L1/L2 and Phase Center Correction Result in Short Baseline

The post-processing configuration is similar to the ultra-short baseline setting. The receiver phase center variation file (Figure 3.41) can be imported into the receiver antenna calibration file, while the antenna placement remains the same as in the previous setup. The correction data is provided by IGR 21175 data.

Post processing setup

- Filter type: Smoother combined solution with forward and backward filter solution
- Elevation mask angel option: 10°
- Dynamics mode of the rover receiver: OFF
- Earth tides correction: ON
- Ionosphere correction: Broadcast
- Troposphere correction: Saastamoinen mode
- Clock data: igr 21175.clk
- Earth rotation parameter file: igr21175.erp
- Precise ephemeris: igr21175.sp3

The results shown in Figure (3.39) represent the PPP method without and with correction for receiver phase center variation. The 2D RMS values are 0.1925 m and 0.1920 m respectively. Incorporating the phase center variation improves positioning accuracy to the centimeter level.

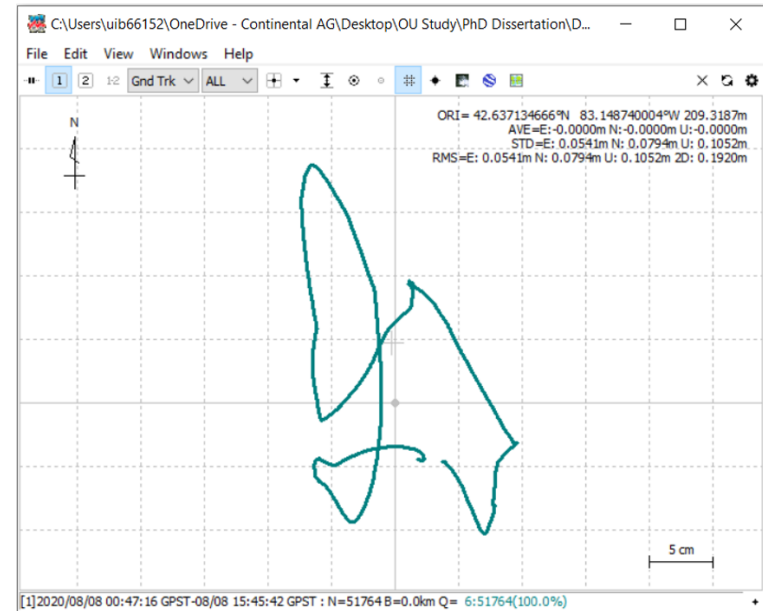


(a)

1D RMS = E: 0.0539m, N:0.0797m, U:0.1052m

2D RMS = 0.1925 m

Q= 6 PPP 51762(100%)



(b)

1D RMS = E: 0.0541m, N:0.0794m, U:0.1052m

2D RMS = 0.1920 m

Q= 6 PPP 51762(100%)

Figure 3.39: L1 and L1/L2 PPP short baseline

3.3 MATLAB GUI for Phase Center Plot and Analysis

The result is analyzed by a self-programming Matlab Graphical User Interface (GUI). which is including the test result import, phase center variation/ phase center offset plot by frequencies and the standard Antenna Exchange File (ANTEX) export [49].

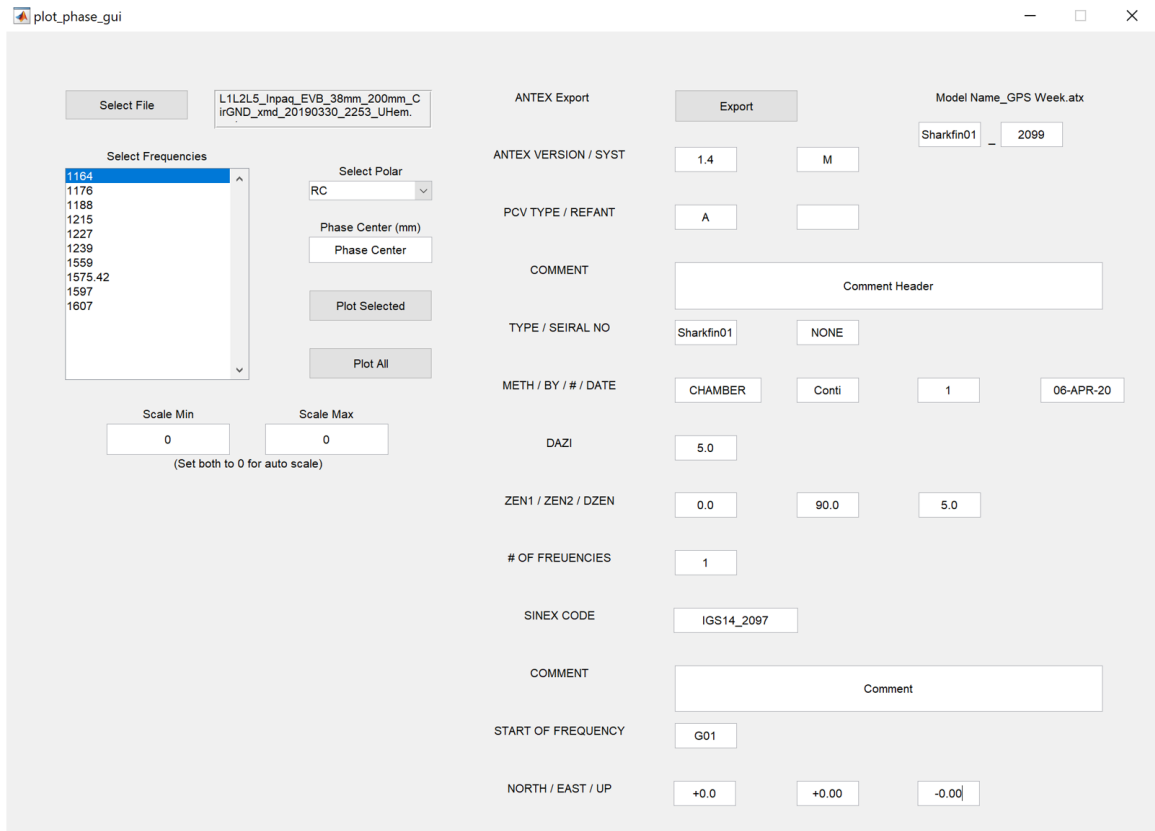


Figure 3.40: Matlab GUI for phase center plot

On the left side of the GUI, it is phase center plot area. The results are including chamber measure phase center variation and calculated phase center offset. On the right side of the GUI, it is ANTEX export setting.

ANTEX VERSION: Format version (1.4)

SYST: Satellite system, 'M': Mixed

PCV TYPE: Phase center variation type, 'A': absolute values

METH: Calibration method.

DAZI: Increment of the azimuth, 0 to 360 with increment 'DAZI' (in degree), common value for 'DAZI': 5.0

ZEN1 / ZEN2 / DZEN: Definition of the grid in zenith angle. In this case, the degree of zenith angle scans from 0.0 to 90.0 degree, increment of 5.0 degrees.

OF FREQUENCIES: Number of Frequencies, for which phase patterns are stored for the current antenna.

SINEX CODE: Name of the antenna calibration model to be used in the SINEX format. (for referencing purposes)

START OF FREQUENCY: Record indicating the start of a new frequency section. The satellite system flag ('G', 'R', 'E', 'C', 'J', 'S') has to be specified together with the frequency number code that has to be consistent with the RINEX definition.

GPS: 'G01' – L1.

NORTH / EAST / UP: Receiver antenna: Eccentricities of the mean antenna phase center relative to the antenna reference point (ARP). North, east and up component (in millimeters).

Sharkfin01L1L2L5_2099.atx - Notepad

File Edit Format View Help

1.4 M

Comment Header

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1 06-APR-20 METH / BY / # / DATE

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START OF ANTENNA
SERIAL NOTYPE / SERIAL NO
DAZI
ZEN1 / ZEN2 / DZEN
OF FREQUENCIES
SINEX Code
COMMENT

START OF FREQUENCY
NORTH / EAST / UP

NOAZI	0.00	1.65	0.92	-1.82	-2.06	-3.69	-6.20	-7.76	-10.13	-12.53	-14.07	-16.45	-18.76	-21.52	-22.11	-23.37	-25.61	-29.50	-33.80
0.0	0.00	1.52	1.49	-2.10	-4.81	-2.79	-7.91	-7.31	-13.39	-15.60	-9.54	-16.74	-22.72	-28.35	-25.89	-27.17	-34.68	-43.18	-49.52
5.0	0.00	1.94	-0.62	-3.18	-6.86	-4.38	-8.45	-8.31	-15.83	-19.77	-18.16	-25.91	-26.94	-26.36	-26.66	-33.82	-44.47	-46.51	-53.53
10.0	0.00	3.12	-0.23	-3.20	-3.40	-5.62	-7.34	-9.24	-16.43	-21.69	-23.61	-27.63	-23.95	-26.87	-29.64	-42.28	-43.63	-48.31	-57.10
15.0	0.00	2.21	0.36	-4.98	-5.54	-2.42	-5.30	-12.15	-18.71	-19.71	-24.10	-26.86	-24.96	-37.06	-33.95	-43.42	-47.09	-50.98	-52.10
20.0	0.00	3.06	-0.03	-5.11	-7.44	-3.92	-6.64	-13.43	-18.79	-19.65	-20.77	-26.65	-32.10	-37.65	-33.60	-40.24	-42.07	-51.22	-56.24
25.0	0.00	3.46	0.05	-2.93	-7.56	-8.23	-9.96	-11.29	-17.02	-19.87	-21.54	-25.43	-35.14	-36.81	-30.38	-38.67	-46.54	-50.25	-54.65
30.0	0.00	4.08	-0.15	-4.86	-5.72	-8.05	-11.18	-10.56	-17.67	-24.20	-25.19	-29.70	-32.60	-38.19	-32.60	-46.74	-53.23	-55.13	-62.46
35.0	0.00	0.86	-2.27	-6.24	-7.04	-6.54	-11.88	-13.17	-18.93	-27.46	-28.90	-30.75	-38.00	-37.42	-37.26	-45.25	-50.98	-50.52	-64.03
40.0	0.00	3.83	-0.53	-1.41	-5.37	-9.68	-12.25	-14.26	-19.44	-26.24	-26.68	-31.73	-32.85	-36.49	-36.07	-44.31	-47.41	-48.74	-60.21
45.0	0.00	4.33	0.81	-2.46	-2.26	-6.16	-11.69	-14.91	-19.62	-22.04	-24.39	-28.12	-30.05	-31.95	-38.04	-41.61	-46.29	-48.63	-62.45
50.0	0.00	-0.85	-2.82	-7.81	-4.73	-3.16	-12.01	-17.46	-21.95	-23.61	-25.74	-31.95	-31.28	-32.59	-44.10	-47.45	-44.81	-44.92	-76.25
55.0	0.00	0.72	-1.77	-4.70	-5.79	-6.26	-12.45	-18.52	-21.87	-26.12	-27.35	-32.84	-32.51	-37.38	-49.46	-49.90	-47.52	-47.72	-70.15
60.0	0.00	1.71	-1.88	-6.27	-3.04	-8.88	-13.73	-19.98	-21.29	-27.07	-29.07	-33.88	-36.52	-42.36	-47.53	-47.11	-45.84	-53.26	-68.63
65.0	0.00	-0.45	-1.41	-9.32	-4.57	-3.89	-12.56	-16.74	-21.88	-23.52	-27.85	-33.48	-33.41	-40.73	-46.71	-44.03	-43.75	-51.63	-67.17
70.0	0.00	-1.95	-1.67	-6.05	-8.14	-6.36	-14.15	-18.31	-20.62	-27.09	-29.14	-33.97	-37.18	-41.09	-44.42	-43.84	-44.90	-51.64	-58.96
75.0	0.00	3.07	-1.29	-3.93	-6.93	-12.27	-14.71	-19.90	-22.93	-30.59	-31.55	-35.79	-38.61	-42.20	-44.33	-43.33	-44.32	-49.60	-57.85
80.0	0.00	-0.14	-3.26	-9.00	-3.78	-7.51	-14.85	-18.59	-23.30	-28.90	-32.25	-34.56	-38.06	-41.96	-42.78	-38.57	-41.51	-43.79	-50.85
85.0	0.00	-2.99	-4.48	-9.44	-6.70	-5.71	-14.93	-20.76	-23.14	-29.21	-32.12	-34.86	-39.07	-39.64	-40.49	-36.59	-38.67	-44.87	-52.33
90.0	0.00	-0.58	-5.62	-5.79	-8.18	-10.69	-16.97	-22.18	-22.71	-28.40	-34.15	-37.17	-37.52	-39.35	-37.92	-36.18	-38.64	-46.75	-53.15
95.0	0.00	-1.54	-7.29	-10.90	-5.61	-10.79	-16.48	-21.21	-21.84	-27.44	-32.41	-37.05	-37.14	-37.39	-37.65	-36.90	-41.32	-47.07	-53.03
100.0	0.00	-5.02	-6.28	-11.87	-6.12	-8.88	-15.12	-18.04	-19.46	-25.76	-29.06	-31.67	-33.02	-34.14	-32.74	-31.55	-34.77	-44.69	-52.37
105.0	0.00	-2.02	-4.35	-3.72	-11.04	-12.82	-14.81	-16.13	-17.93	-24.66	-27.65	-29.25	-30.62	-33.30	-31.98	-32.16	-35.67	-44.06	-49.94
110.0	0.00	0.33	-5.70	-5.15	-9.49	-16.90	-16.91	-18.13	-19.74	-25.70	-27.75	-29.31	-30.55	-34.73	-34.05	-33.12	-34.08	-41.95	-49.31
115.0	0.00	-2.99	-4.06	-9.33	-5.54	-11.48	-15.55	-17.08	-20.18	-24.69	-27.36	-30.00	-29.98	-30.70	-30.70	-29.59	-35.06	-43.67	-48.02
120.0	0.00	-1.10	-4.94	-4.05	-8.21	-12.19	-15.78	-16.00	-19.96	-23.90	-25.51	-27.28	-28.11	-29.27	-30.58	-31.74	-36.95	-42.75	-46.95
125.0	0.00	1.66	-2.93	-2.16	-7.45	-14.98	-16.73	-17.87	-21.28	-23.65	-25.67	-26.71	-27.59	-29.06	-30.65	-32.37	-37.95	-44.03	-45.60

Figure 3.41: An example test: Sharkfin01L1L2L5 2099

3.4 Summary

In this chapter, the main research work includes antenna performance evaluation and antenna field testing. The research flowchart is shown in Figure 3.42. A square patch multi-band GNSS antenna is tested using a standalone EVB, a compact shark-fin antenna, and the same shark-fin antenna mounted on the car roof. The evaluation of antenna performance includes passive gain, axial ratio, phase center variation, and calculated phase center offset [40]. A standard Antenna Exchange File (ANTEX) is generated based on the PCV results. During the field testing, two scenarios are considered: a 40 m ultra-short baseline and a 9.35 km short baseline. The comparison of results between these two baselines is listed in

Table 3.5, including code pseudorange, carrier phase measurements, and PPP method with single L1 band as well as L1/L2 dual-band respectively. By applying PPP method without and with phase center variation correction, it can be observed that the positioning accuracy improves when using phase center variation correction even without RTK subscription service. The PPP position achieves centimeter-level accuracy which surpasses DGNS (meter-level) but falls behind carrier phase (millimeter-level) in ultra-short baseline calculation. These results align with our expectations for different positioning methods.

- The carrier phase method can significantly improve position accuracy compared to the code pseudorange method.
- Dual frequency band correction is much better than single frequency band correction.
- Both carrier phase and code pseudorange methods experience degradation in correction performance when baseline distance is increased, while the PPP method maintains the same level of position accuracy for ultra-short and short baseline distances.
- By using antenna phase center correction, position accuracy can be improved to millimeter-level precision.

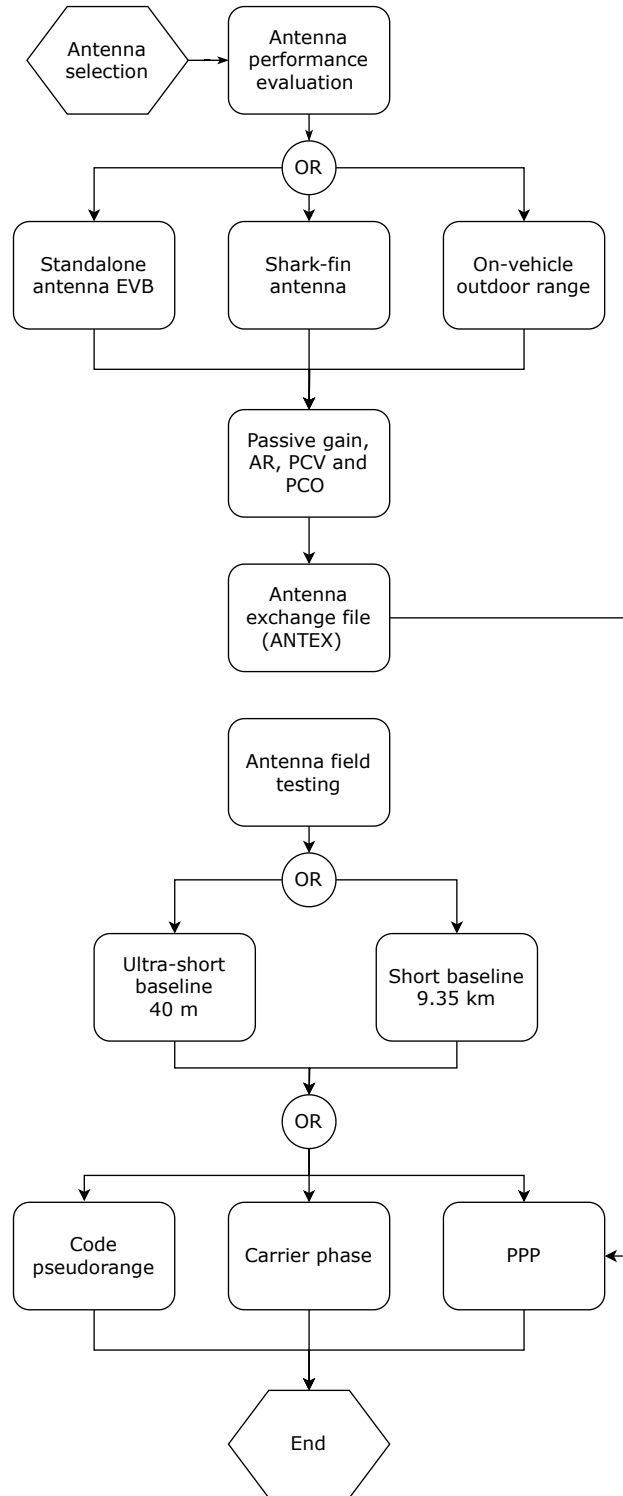


Figure 3.42: Research flowchart

Table 3.5: Position accuracy between ultra-short and short baseline

	Ultra-short Baseline (40 m)				Short Baseline (9.35 km)			
	East RMS	North RMS	Up RMS	2D RMS	East RMS	North RMS	Up RMS	2D RMS
Code L1 [m]	0.3589	0.7442	1.2308	1.6524	0.9640	1.0369	2.3197	2.8316
Code L1/L2 [m]	0.1784	0.3098	0.5946	0.7149	0.3408	0.4491	0.8683	1.1275
Phase L1 [m]	0.0020	0.0022	0.0034	0.0059	0.0120	0.0116	0.0173	0.0335
Phase L1/L2 [m]	0.0007	0.0016	0.0016	0.0035	0.0028	0.0047	0.0119	0.0108
PPP L1/L2 [m]	0.0543	0.0618	0.0763	0.1645	0.0539	0.0797	0.1052	0.1925
PPP PCV L1/L2 [m]	0.0532	0.0617	0.0743	0.1629	0.0540	0.0794	0.1052	0.1920

CHAPTER FOUR

V2V ANTENNA DESIGN AND RURAL PROPAGATION SCENARIOS CHANNEL ANALYSIS

Cellular Vehicle-to-Everything (C-V2X) is an innovative component of LTE technology, and the latest New Radio V2X (NR-V2X) will play a crucial role in the 5G era, particularly for facilitating interaction between Internet-of-Things (IoT) device and the automotive network [50]. With the advent of 5G cellular networks, advanced intelligent Multiple Input and Multiple Output (MIMO) systems, as well as Mobile Edge Computing (MEC) technologies, enhance the overall network's resilience and reduce latency compared to traditional communication networks. The fundamental communication air interface of this system encompasses both cellular communication User-to-UTRAN (Uu) mode and Direct Communication PHY Sidelink (PC5) mode 4. Uu interface has been extensively employed in Vehicle to Network (V2N), exhibiting improved transmitter-receiver scenario changes [51][52][53][54][55][56]. Previous studies have primarily focused on V2V channel propagation in urban/suburban scenarios [57], underground indoor garages with antennas positioned at a height of 1.7 m [58], as well as determining inter-vehicle communications channel models with antennas [59]. According to SAE J2945/1 Section 6.4 RF performance requirements for detailed specifications [3], an omni-directional radiation pattern on the roof is essential for antennas. IEEE 802.11p standard [60], along with 3GPP TR36.885 [61] and TR37.885 [62] documents only consider one omni-directional antenna. In contrast, authors in [63] proposed four antenna positions on the rooftop accompanied by simulated radiation patterns. And the authors in [64] "with the height of transmitter and receivers antennas relatively close to the ground level make the V2V channel significantly

different from the well studied channels of other technologies such as cellular network". In the real world, due to the vehicle structure and size limitations, it is difficult to realize this type of omni antenna. Some automotive Original Equipment Manufacturers (OEMs) design a sharkfin antenna on the rear area of the roof, which is the highest point of the entire vehicle. This could be an optimal position for the V2X antenna. However, in recent years, panoramic glass roofs have become a modern trend in car design. Since there is no electrical wall boundary on the front side of the antenna due to the glass material, significant attenuation occurs in its radiation pattern. One possible solution is to metalize the glass roof; however, this would involve high costs for materials and additional processing work. On the technical side, the Network Access Device (NAD) module design has only one transmit path based on 3GPP TS 36 Group release 14 [65], Therefore, the LTE-V2X system can only support one transmit/receive path. In the novel NAD design, Cyclic Delay Diversity (CDD) is introduced in dual transmit chains following 3GPP TS 36 Group release 15 [66], which includes two Tx/Rx paths and could be an ideal NR-V2X configuration. The recommendation position for these two antennas is front-rear directional placement since most communication scenarios occur in front and rear areas [11][67][68], such as Forward Collision Warning, Lane Change Warning, Not Pass Warning discussed in Section 2.1. Alternative antenna positions may consider left-right rear minor placement. Overall, careful consideration of vehicle structure and styling is required for vehicle antenna design and placement [69].

Meanwhile, the propagation of cellular network channel has been studied over the past decades, and transmission between tower base stations and handheld devices can be carried out Line of Sight (LOS) and Non-Line of Sight (NLOS) conditions [19] [70][71]. However, V2X channel propagation needs to be re-analyzed due to changes in transmitter height and fast movement of vehicle, especially in

small-scale fading where there can be rapid fluctuations in the amplitude of a radio signal over a short period of time or distance traveled [19].

4.1 Research Methodology

This chapter's research work focuses on the analysis of Band 47 (5855-5925 MHz) V2X antenna performance in a rural area propagation channel at different heights, including small-scale fading and real-world path loss testing. The study includes the following scenarios:

- omni-directional antenna as reference at 1.6 m height
- directional antenna at 0.6 m height - bumper level
- directional antenna at 1.0 m height - logo level
- directional antenna at 1.6 m height - roof level

The next section will introduce the design of a low-profile directional V2X antenna for automotive applications. Following that, in Section 4.4, channel characteristics such as receiving power and delay spread in an open field rural area can be extracted from simulation results based on ray-tracing simulation and visualization of wave propagation through dielectric surfaces.

- two-ray ground reflection pathloss model as reference
- rural-area propagation channel simulation by ray-tracing
- real-world on-vehicle testing

Finally, the real-world on-vehicle measurement needs to be verified and compared with the reference antenna height factor (Table 4.1) from SAE J3161/1: On-Board System Requirement for LTE-V2X Safety Communications [1][72].

Table 4.1: Antenna height correction factor [1]

Antenna Height (meter)	Antenna Height Factor (dB)
2.5	1.9
2.4	1.8
2.3	1.8
2.2	1.7
2.1	1.6
2.0	1.4
1.9	1.2
1.8	1
1.7	0.7
1.6	0.4
1.5	0
1.4	-0.4
1.3	-0.9
1.2	-1.4
1.1	-2.1
1.0	-2.8
0.9	-3.6
0.8	-4.5
0.7	-5.6
0.6	-6.8
0.5	-8.3

4.2 V2V Antenna Design

The antenna, as the primary component in the Radio Frequency (RF) front-end link, plays a crucial role in determining the overall system performance. In order to meet the stringent requirements of automotive industry, various aspects of the V2V antenna need to be carefully considered, including industrial feasibility, material cost-effectiveness and ease of installation. Following SAE J2945/1 [3] and SAE J3161/1 [72] standards and real-world implementation experiences, it is recommended that the 3GPP release 15 [66] compliant antenna should adopt a dual-channel paths configuration. As discussed in the previous sections, two possible

placements for antenna are either at the front-rear or left-right positions. In typical scenarios [73], front-rear communication can effectively cover most usage scenarios. Consequently, many OEMs prefer placing antennas on the front-rear windshield based on these considerations. Therefore, a directional antenna with low profile and simple structure that facilitates easy assembly is required for this system.

The whole antenna includes the following elements: a plastic protective housing, a microstrip feed patch radiator and Fakra[®] connector. The antenna is designed using CST Microwave Studio[®], which is a Commercial Off-The-Shelf (COTS) three-dimensional, full wave electromagnetics (EM) field Time Domain solver, as highlighted in Figure 4.1(a). The V2V system requires vertical polarization of EM propagation, and the radiation patch needs a two-orthogonal difference phase feeding network. Due to the crowded front windshield sensor-farm area, the antenna is limited to horizontal placement as shown in Figure 4.1(b), which represents a Computer Aided Design (CAD) view. A real mass production sample is built up shown in Figure 4.1(c). Its dimensions are limited to $80 \times 30 \times 30 mm^3$. The antenna could be used in pairs to cover both front and rear coverage areas and it is tested in a near-field ETS[®] AMS-8923 anechoic chamber at SGS[®] Suzhou lab in China, as depicted in Figure 4.1(d). The test setup follows the same coordinate system as shown Figure 4.1(a).

The antenna housing material is Polycarbonate- Acrylonitrile Butadiene Styrene (PC-ABS) with a Dielectric Constant (DK) set at 2.98 in simulation. After fabricating real samples and comparing them with the simulation, as shown in Figure 4.2, the resonant frequency of the return loss for both matches well within the working frequency Band 47 (5855-5925 MHz), and the S-parameter (S11) of the return loss is below -20 dB. In common antenna requirements, the return loss lower than -10 dB is necessary. This design takes into consideration production

deviations, and due to uncertainties in raw materials and assembly consistency, there may be variations in the center frequency within a certain range.

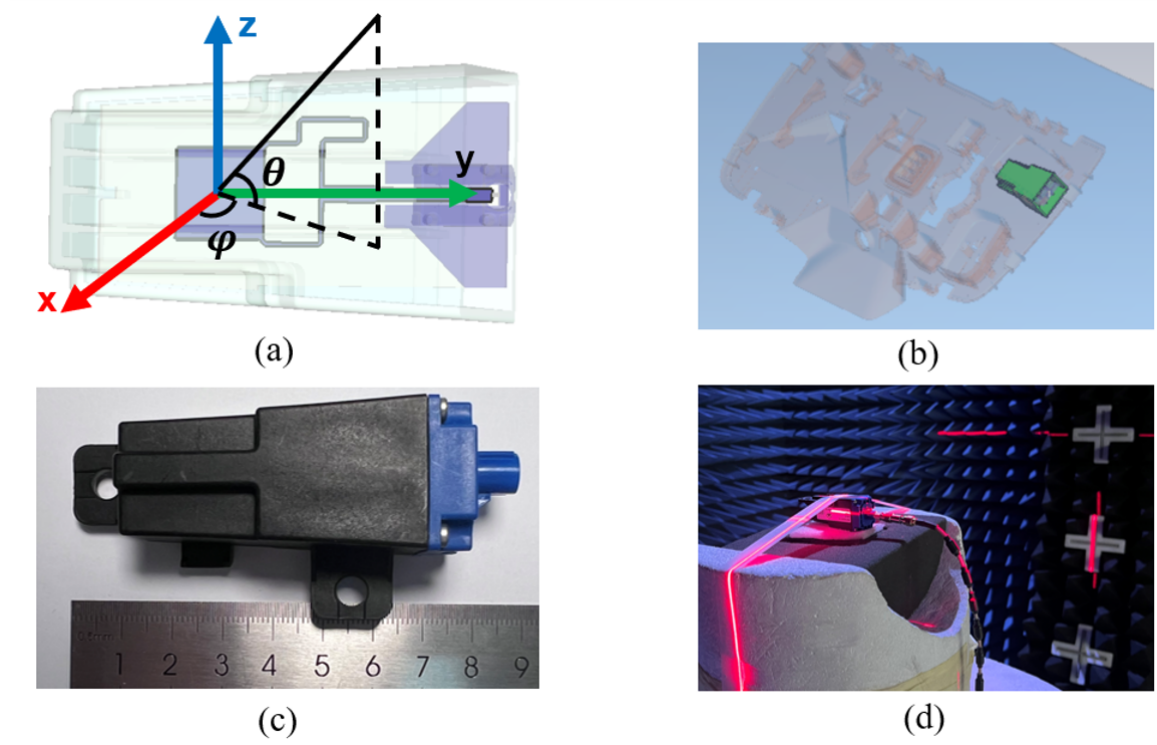


Figure 4.1:

- (a) Isometric overview of the patch antenna with Cartesian coordinate system;
- (b) Antenna placement on sensor farm area in CAD file;
- (c) A mass production sample with dimension;
- (d) Antenna under test in ETS[®] AMS-8923 anechoic chamber.

According to the J3161/1 test specification [72], the minimum 70% coverage of Transmit Signal Strength (TxSS) is 15 dBm,

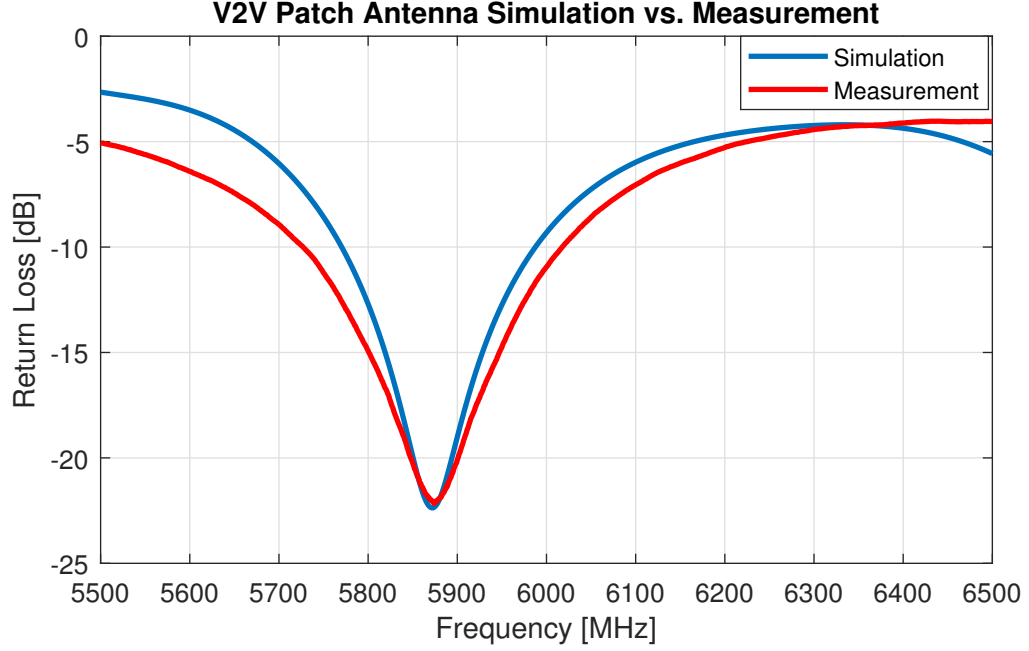


Figure 4.2: Return loss result of simulation and real antenna sample result

$$TxSS(\theta, \varphi) = EIRP(\theta, \varphi) + f_{HF}(h) \quad (4.1)$$

$$EIRP(\theta, \varphi) = Tx_{power} - Loss + Gain_{Ant}(\theta, \varphi) \quad (4.2)$$

where Tx_{power} represents the transmission power of the V2V Telematic Control Unit (TCU), $Loss$ denote the loss caused by cable and connector, $f_{HF}(h)$ is the height factor of antenna place, and $Gain_{Ant}(\theta, \varphi)$ refers to the free space far-field antenna gain. Assuming that $TxSS = TxSS_{min} = 15$ dBm, $Tx_{power} = 23$ dB, the cable & connector loss is 5 dB, and the reference antenna height factor is 0, with a given value of $h = 1.5$ m [69].

Table 4.2: Minimum linear average free space antenna gain (70% coverage)

	Linear Avg. Gain $89^\circ \leq \theta \leq 91^\circ$	70% Gain $89^\circ \leq \theta \leq 91^\circ$	Linear Avg. Gain $91^\circ \leq \theta \leq 96^\circ$ $84^\circ \leq \theta \leq 89^\circ$
Front Zone $-60^\circ \leq \varphi \leq 60^\circ$	-3 dBi	-6 dBi	-6 dBi
Side Zone $60^\circ \leq \varphi \leq 120^\circ$ $240^\circ \leq \varphi \leq 300^\circ$	-9 dBi	-12 dBi	-12 dBi
Rear Zone $120^\circ \leq \varphi \leq 240^\circ$	-3 dBi	-6 dBi	-6 dBi

$$Gain_{Ant}(\theta, \varphi)_{min} = TxSS_{min} - Tx_{power} + Loss - f_{HF}(h) = -3dB \quad (4.3)$$

In case of two antenna Tx diversity mentioned in [72],

$$TxSS_{Diversity}(\theta, \varphi) = Max[TxSS_1(\theta, \varphi), TxSS_2(\theta, \varphi)] \\ + DiversityGain[TxSS_1(\theta, \varphi) - TxSS_2(\theta, \varphi)] \quad (4.4)$$

with CDD technology in 3GPP TS 36 Group release 15 [66], the two antennas can work simultaneously to provide diversity gain in the antenna radiation coverage area. According to specification [72], the Tx Signal Strength is limited by front, rear and side zones. Based on Eq.(4.3) the calculated minimum $Gain_{Ant}(\theta, \varphi)_{min} = -3$ dB, and the overall antenna gain requirements in different angle zones are shown in Table 4.2.

Figure 4.4 shows the comparison between simulation and anechoic chamber measurement for the far-field radiation gain pattern of a single antenna in terms of elevation pattern (Figure 4.4a) and azimuth pattern (Figure 4.4b). In Figure 4.4(a),

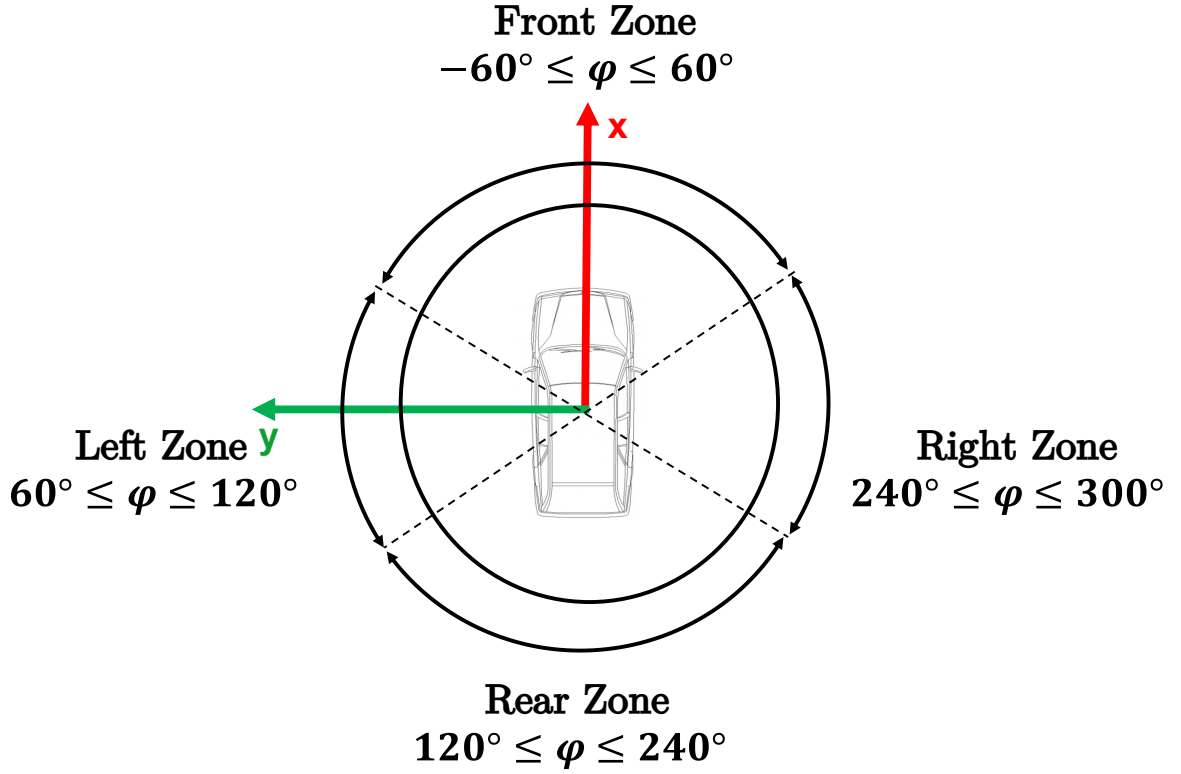


Figure 4.3: Definition of Tx signal strength zones

the elevation ($\varphi = 0^\circ$) gain pattern is displayed, with the antenna's main beam covering Theta (θ) from 84° to 96° , as defined by SAE J3161/1 test specification [72]. The main beam angle definition takes into account different signal incident elevation angles in vehicle communication. The green curve represents the simulation result, which matches with the blue curve obtained from real sample chamber measurements. Figure 4.5 illustrates that the azimuth gain radiation pattern covers Phi (φ) angles ranging from -60° (300°) to 60° , encompassing most of the communication area in the direction of incident signal drive. By using this same antenna for both front and rear vehicle placements, a diversity radiation pattern is

calculated using Eq.(4.4), demonstrating an improved coverage on the side zone in Figure 4.5. A full spherical far-field 3D radiation pattern can be exported by considering all ergodic elevation and azimuth angles, which will be utilized for channel simulation in the next section.

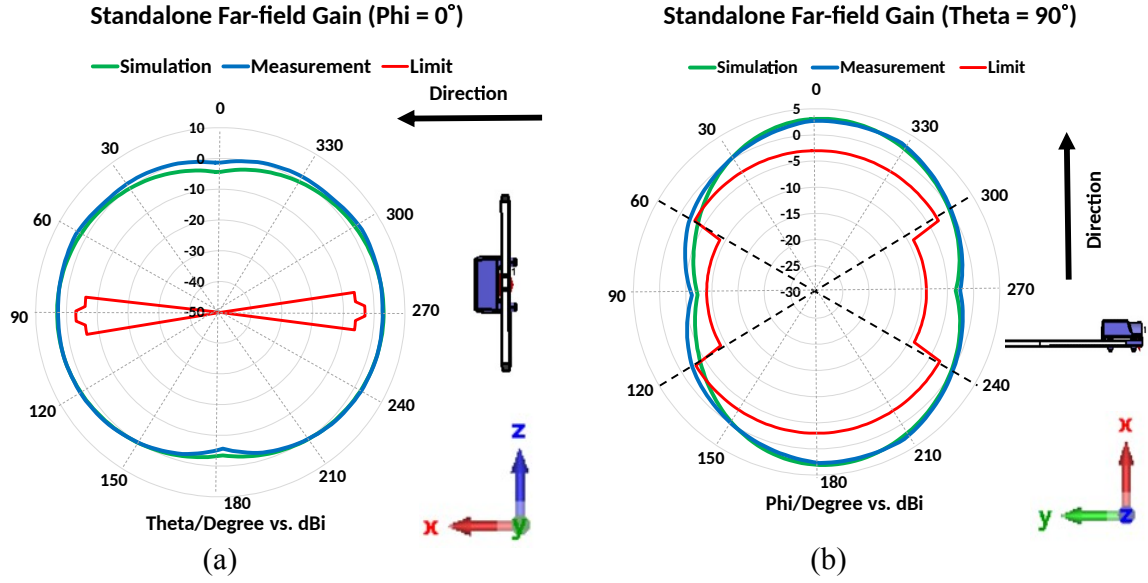


Figure 4.4: Far-field radiation gain pattern simulation vs. anechoic chamber measurement result;

(a) Elevation cut radiation gain pattern ($\Phi = 0^\circ$);

(b) Azimuth cut radiation gain pattern ($\Theta = 90^\circ$).

4.3 On-vehicle Chamber Measurement

The performance of the antenna is highly dependent on the surrounding environment. To assess its real working performance, an on-vehicle antenna is tested

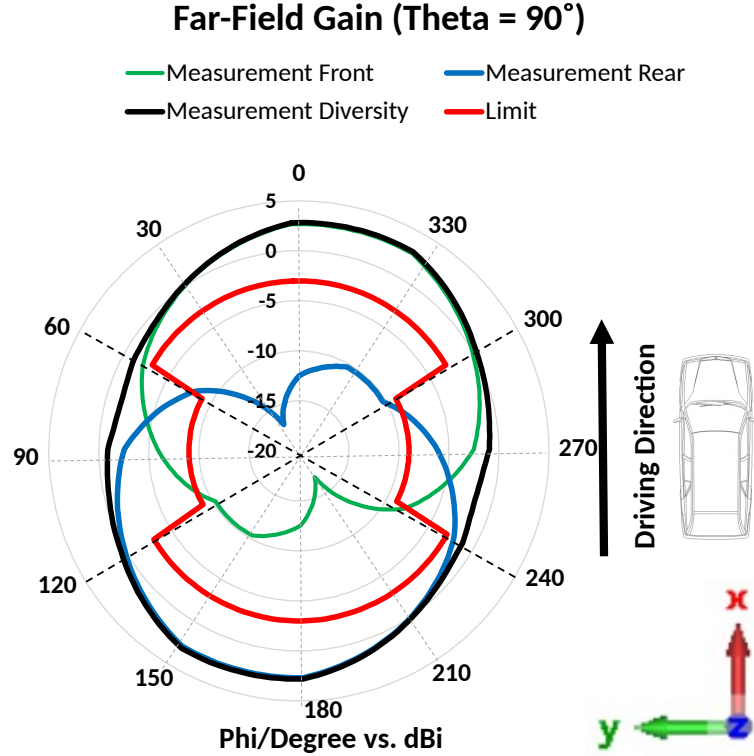


Figure 4.5: Far-field radiation diversity gain pattern (Theta = 90°).

in SGS[®] Suzhou lab using the MVG-SG[®] 3000 anechoic chamber: a multi-probe full vehicle OTA test system. The chamber configuration is shown in Table 4.3. In Figure 4.6, an on-vehicle antenna is being tested in this chamber. The test coordinator follows a right-hand coordinate system as depicted in the picture, with the x-axis representing the vehicle's driving direction. Three different rear antenna heights were considered, as illustrated in Figure 4.7: 0.6 m, 1.0 m, and 1.6 m.

The MVG-SG 3000 is a full-scale antenna anechoic chamber that provides a three-dimensional (3D) plane for evaluating the antenna radiation performance. The most common method to analyze the performance is by using the elevation

Table 4.3: SGS Suzhou lab configuration

MVG-SG 3000	
Chamber size	15 m (L) $\times$ 13 m (W) $\times$ 10 m (H)
Probe array	111 dual polarized probes
Elevation angle range	$\theta = 0^\circ$ to 110° , $\Delta\theta = 1^\circ$
Frequency range	400 MHz to 10 GHz
Measurement distance	6 m
Vehicular turntable	8 m diameter, 3500 kg max load-bearing
Post-processing	NF2FF, TWSE, Insight



Figure 4.6: On-vehicle testing under MVG-SG 3000 full-scale chamber

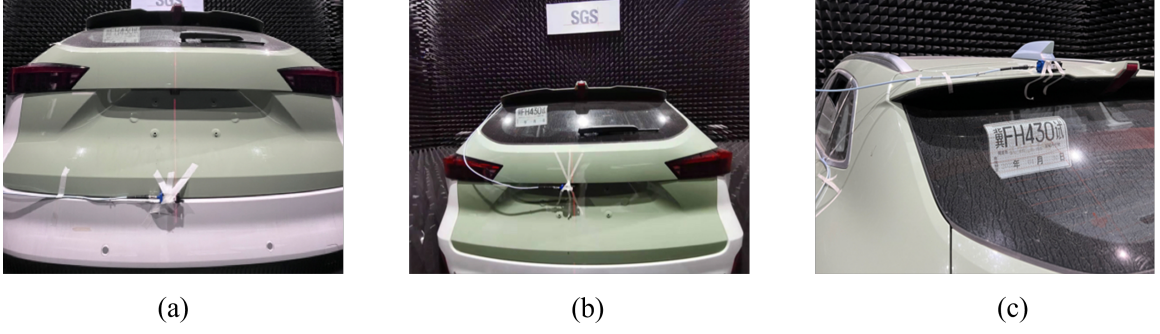


Figure 4.7: Rear antenna heights placement;
(a) antenna height at 0.6 m - bumper level;
(b) antenna height at 1.0 m - logo level;
(c) antenna height at 1.6 m - roof level.

plane ($x - z$ cut) and azimuth plane ($x - y$ cut) radiation patterns. The front antenna is mounted on the front windshield in the sensor farm area. The results of the elevation plane can be found in Figure 4.8(a). The rear antennas at three different heights are shown in Figure 4.8(b). Three different elevation zones with limits are defined as shown in Table 4.4. Additionally, the results of the azimuth plane can be found in Figure 4.9 and Table 4.5.

The antenna radiation patterns exhibit many ripples due to the surrounding environment of the integrated antenna on the vehicle, such as diffracting and scattering parts of the car body. From the on-vehicle results, it can be observed that the diversity gain of elevation and azimuth plane in Figure 4.10 remains constant in the front region but shows different performance in the rear region. The front antenna is fixed in the front windshield, while the rear antennas are positioned at three different heights: 0.6 m, 1.0 m, and 1.6 m.

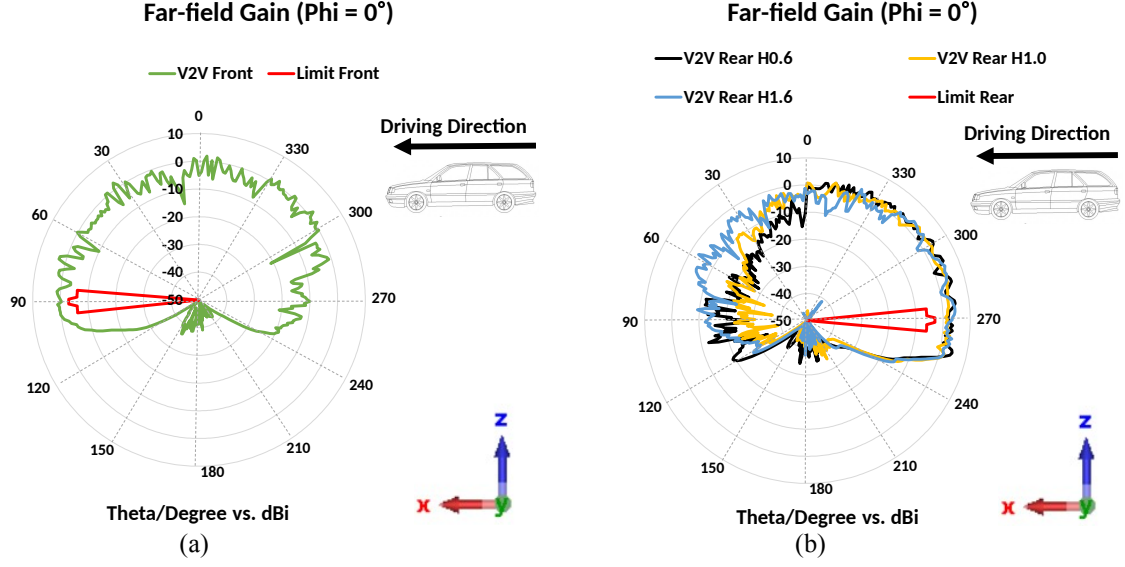


Figure 4.8: Elevation plane in $x - z$ cut of front and rear antenna results

Table 4.4: The elevation plane result of front and rear antenna results

	Limit (dBi)	H 0.6 m Avg. gain(dBi)	H 1.0 m Avg. gain(dBi)	H 1.6 m Avg. gain(dBi)
Front Zone $84^\circ \leq \theta \leq 89^\circ$	-6	0.37	0.32	0.46
Front Zone $89^\circ \leq \theta \leq 91^\circ$	-3	0.24	0.24	0.43
Front Zone $91^\circ \leq \theta \leq 96^\circ$	-6	0.92	0.85	1.00
Rear Zone $271^\circ \leq \theta \leq 276^\circ$	-6	3.05	2.04	4.50
Rear Zone $269^\circ \leq \varphi \leq 271^\circ$	-3	2.34	1.53	3.15
Rear Zone $264^\circ \leq \varphi \leq 269^\circ$	-6	3.10	1.65	1.68

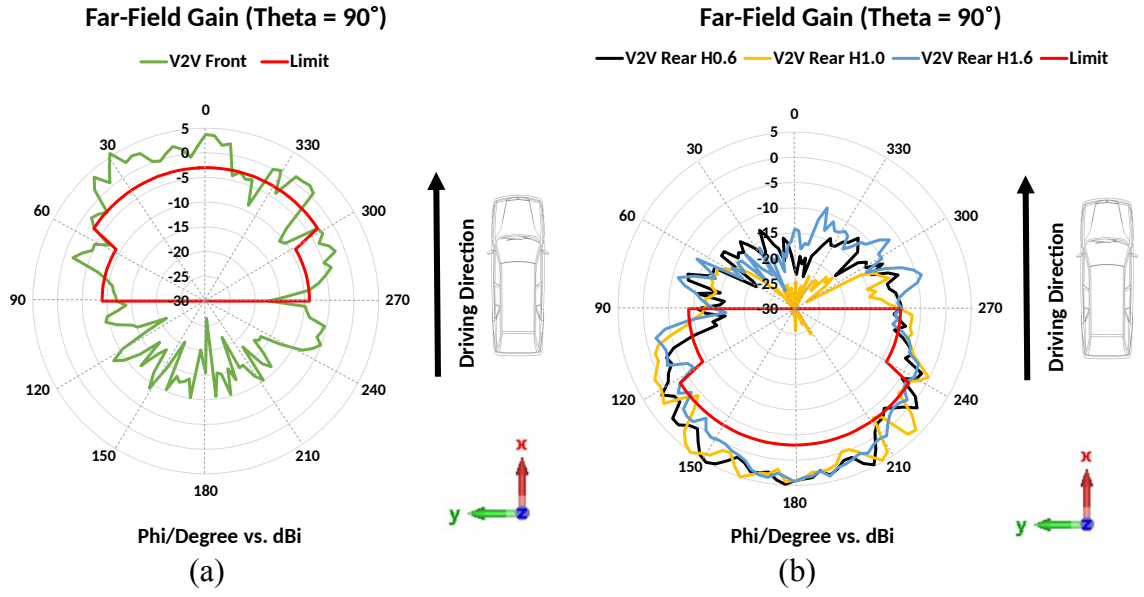


Figure 4.9: Azimuth plane in $x - y$ cut of front and rear antenna results

Table 4.5: The azimuth cut result

	Limit (dBi)	H 0.6 m Avg. gain(dBi)	H 1.0 m Avg. gain(dBi)	H 1.6 m Avg. gain(dBi)
Front Zone $-60^\circ \leq \varphi \leq 60^\circ$	-3	-0.64	-0.60	-0.14
Left Zone $60^\circ \leq \varphi \leq 120^\circ$	-9	-5.95	-4.63	-5.08
Rear Zone $120^\circ \leq \varphi \leq 240^\circ$	-3	1.73	1.75	1.19
Right Zone $240^\circ \leq \varphi \leq 300^\circ$	-9	-4.00	-3.86	-3.69

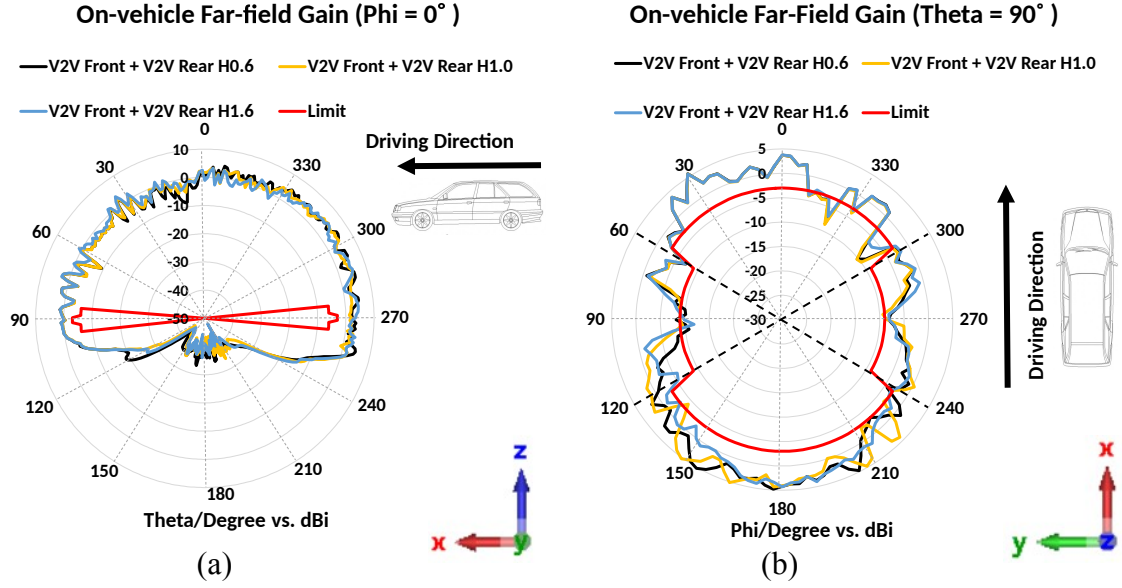


Figure 4.10: Elevation and Azimuth plane of front and rear antenna

4.4 Rural Scenario Channel Simulation

As introduced in Section (2.3.1.3), the two-ray path loss model can be verified using a Commercial-off-the-Shelf (COTS) software called Altair Feko WinProp[®], as shown in the simulation setup depicted in Figure 4.11.

The output power of the transmit antenna is set to +23 dBm, using an omni-directional antenna at 5900 MHz. The transmission (Tx) antenna height is 1.5 m. The ground size is 400 m $\times$ 60 m, with dry ground material being used. The trajectory Route 0 extends to a distance of 300 m away. On this trajectory, the received power result can be found at the bottom of Figure 4.11, where the power level is measured in dBm according to the colormap on the right-side legend. Considering the installation of vehicle antennas as a factor, subsequent simulation work will consider three different heights for the receiving (Rx) antenna: bumper

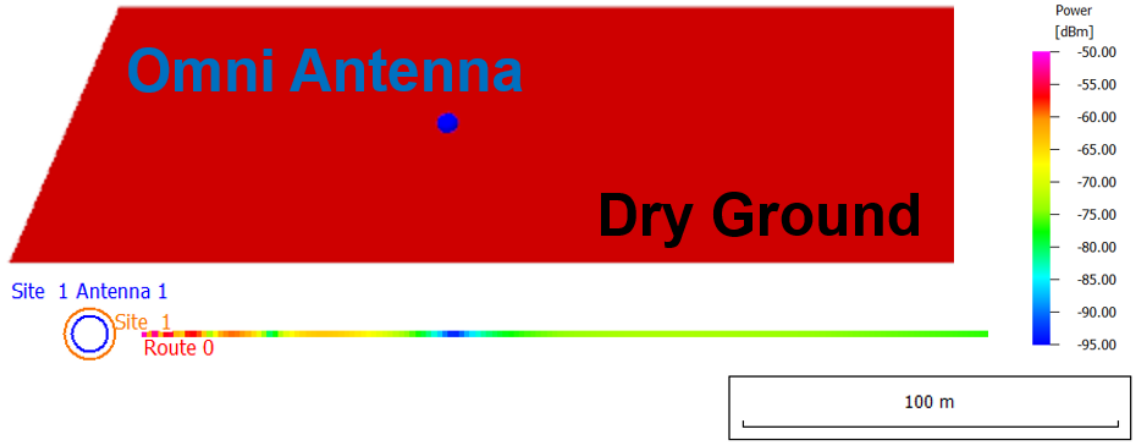


Figure 4.11: Two-ray ground reflection model of simulation setup by Feko WinProp

level at 0.6 m, sedan truck or SUV logo level at 1.0 m, and SUV spoiler level at 1.6 m.

The path loss results of the receiving antenna at different heights are presented in Figure 4.12, following the simulation setup shown in Figure 4.11. The curves exhibit ripples with varying phase construction or deconstruction. When the Tx-Rx distance is large, the phase difference is not significant. The E-field attenuates in a fluctuating manner, and its amplitude continually decreases depending on $distance^2$.

The Figure 4.12 illustrates the analysis of different antenna heights with receiving power. It can be observed that the severity of ground reflection increases as the Rx antenna height increases. When the Rx antenna height is 0.6 m (blue curve), there is one oscillation. When the height reaches 1.0 m, there is still one oscillation (orange curve), which is similar to the height of 1.6 m. This observation can be explained by Fresnel zone theory. By using Eq.(2.15) for two-ray ground

reflection path loss, it can be calculated that when $h_t = 1.5$ m, and the first Fresnel zone at $h_r = 0.6$ m is 68 m, indicated by a blue point and dashed line on the graph. The remaining heights of $h_r = 1.0$ m and 1.6 m can also be calculated using Eq.(2.15), with their first Fresnel zone distances being 114 m and 182 m respectively.

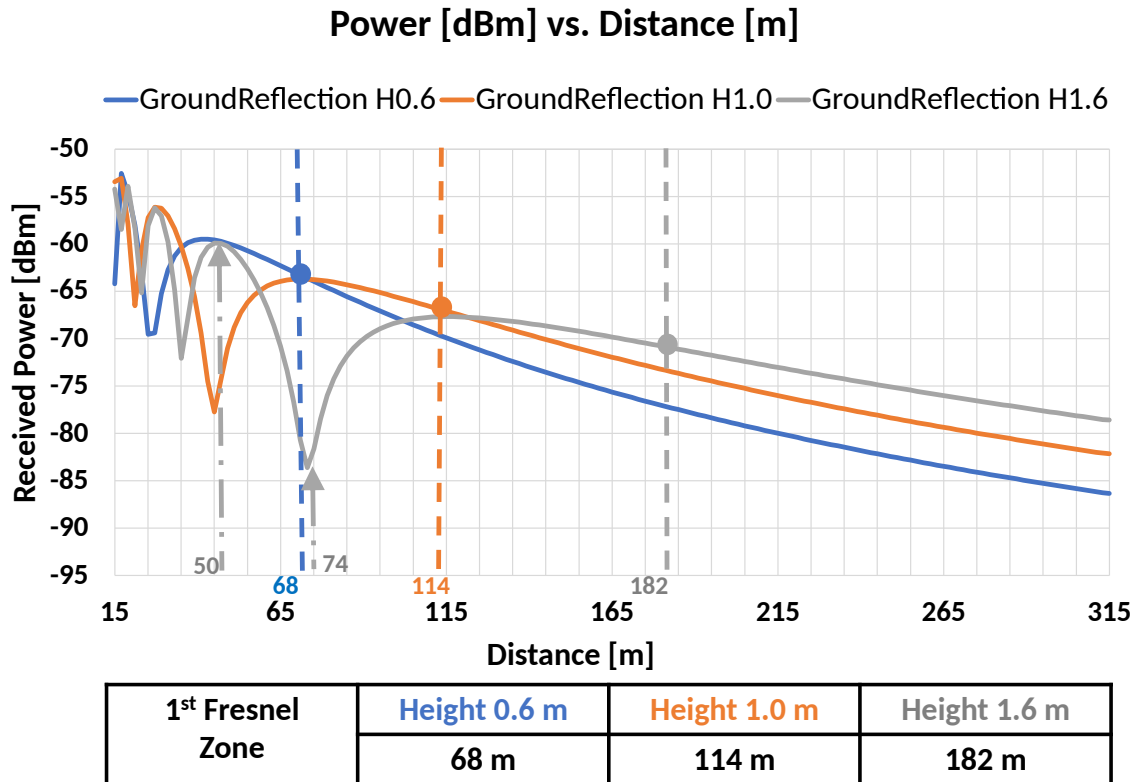


Figure 4.12: Two-ray ground reflection path loss simulation

The detail of phase contribution of the receiving level is shown in Table 4.6. There are a total of 2 rays with an Rx antenna height of 1.6 m, and Ray 0 represents the first LOS ray from the transmitter's omni-directional antenna to the

Table 4.6: Ground reflection H1.6 @50 m

Ray	Interactions	Delay (ns)	Phase(deg)	Signal Level (dBm)
0	0	171.86	341.12	-53.16
1	1(1R,0D,0T,0S)	172.11	324.93	-55.46

Table 4.7: Ground reflection H1.6 @74 m

Ray	Interactions	Delay (ns)	Phase(deg)	Signal Level (dBm)
0	0	255.25	340.06	-56.59
1	1(1R,0D,0T,0S)	255.41	152.91	-58.15

receiver position directly. Ray 1 interacts with 1R (reflection). At a distance of 50 m, both Ray 0 and Ray 1 have similar phases, resulting in constructive addition of their signal powers. However, under the same conditions described in Table 4.7 at a distance of 74 m, these two rays have an almost half-period phase difference, leading to deconstruction of the signal power due to their phase discrepancy.

Meanwhile, a rural road scenario is created in the Feko WinProp[®] simulation (Figure 4.13). The scenario includes a vehicle body made of metal and glass windshield, a ground composed of grass, and bi-directional boulevard with two car lanes. The transmit antennas (blue circle: omni-directional antenna and blue arrow: V2V patch directional antenna) are placed on the front windshield at a height of 1.5 m. The radiation pattern of the patch directional antenna is directed towards the vehicle's direction, as described in Section 4.2. The antenna configuration and material properties used in the simulation are shown in Table 4.8 and Table 4.9.

The reference position of the Tx antenna is set at 0 m, and the starting point of the Rx antenna is set 15 m away from the Tx antenna. The Route 0 follows a

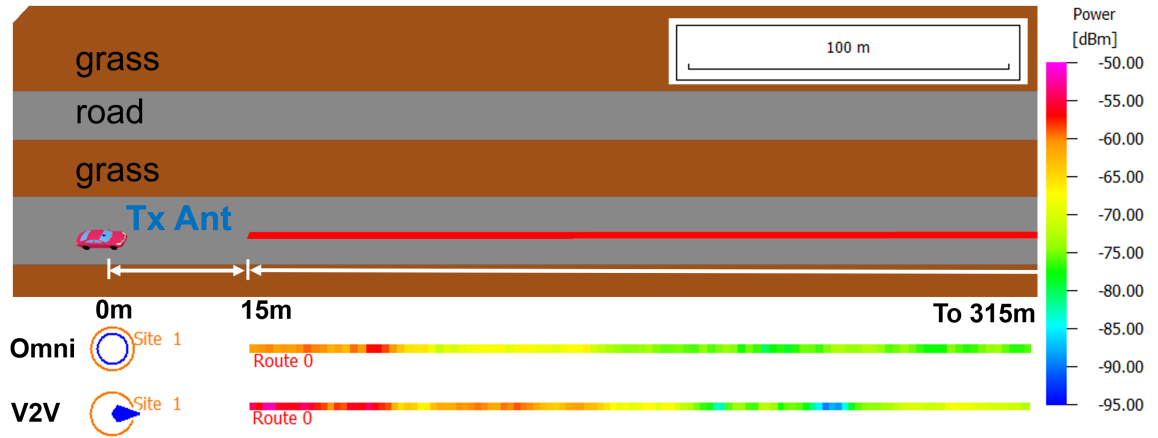


Figure 4.13: Simulation rural scenario and received power result over 315 m range

Table 4.8: Tx-Rx antenna configuration

Tx Omni/ Patch antenna		Receiving antenna	
Height	1.5 m	Low height	0.6 m
Output power	+23 dBm	Mid height	1.0 m
Frequency	5900 MHz	Top height	1.6 m

Table 4.9: Material properties @ 5900 MHZ

	Grass: Loam	Road: Concrete
Transmission loss	2.18 dB	27.66 dB
Reflection loss	12.95 dB	7.51 dB
Incident (Min.)	12 dB	12 dB
Incident (Max.)	19 dB	19 dB
Diffracted	9 dB	9 dB
Rel. permittivity	2.5	9
Rel. permeability	1	1
Conductivity	0.008 S/m	0.194 S/m

sampling trajectory setting from 15 m to 315 m with a step resolution of 1 m.

According to the colormap on the right side of Figure 4.13, the receiving power level in dBm indicates that both types of antennas follow Route 0.

Figure 4.14 illustrates the entire prediction area for receiving power distribution. In Figure 4.14(a), it is evident that the transmit antenna is omni-directional. At the transmission point, the receiving power distribution is widely spread, providing strong energy to antennas around the site. Comparing this with a directional patch antenna, Figure 4.14(b) shows that the main beam of the V2V antenna points towards the driving direction, concentrating most of its power in that direction. Due to its high gain directivity, this vertically polarized wave allows distinguishing between two directional roads and a middle loam based on propagation properties; however, an omni-directional antenna cannot detect such details.

The curvature result in Figure 4.15 includes three types of antenna heights and an omni-directional antenna at the reference height of 1.6 m. The variation trend of the received signal power is similar to the theoretical two-ray path loss.

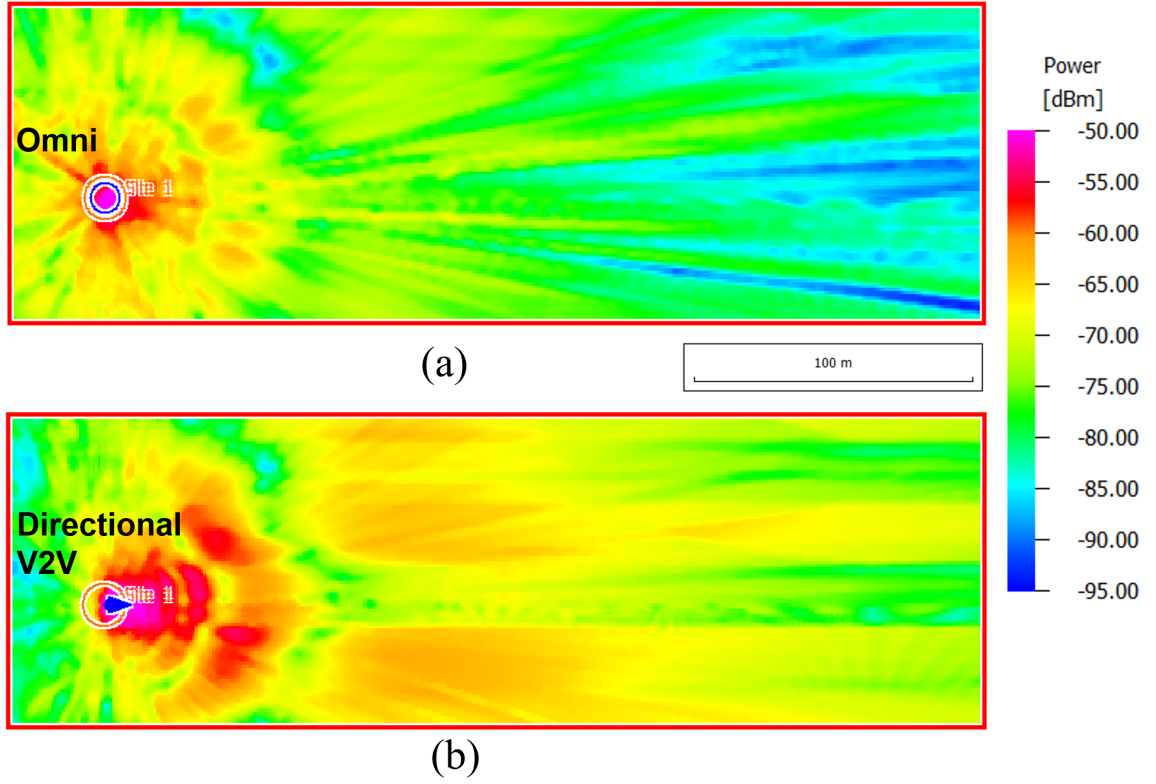


Figure 4.14: Rural scenario predicted receiver power of Omni vs. V2V patch directional antenna with height 1.6 m;
(a) Tx Omni-directional antenna;
(b) Tx V2V patch directional antenna

The receiver power decreases with increasing distance. By using the same V2X directional patch antenna at different height elevations, it can be observed that higher antenna placement leads to a higher gain on the Tx/Rx far-end distance (yellow, green, and red curves). In closed areas, due to ray propagation, the power experiences oscillation caused by phase construction and deconstruction. This explains the sudden change in curve level over a short distance and accounts for small-scale fading in this area. In Section 4.2, it is shown that the patch antenna

has a maximum gain of 3.1 dBi at horizontal plane as depicted in Figure 4.4. The benefits of using a directional antenna are evident from comparing the red curve (patch antenna) receiving more power than the black curve (omni-isotropic antenna). By employing front-rear diversity directional antennas, greater system gain can be achieved compared to using only one omni-directional antenna.

Figure 4.16 refers to the delay spread result obtained from Figure 4.15. It describes the delay spread with Tx and Rx distance. The delay spread remains constant at shorter distances, but shows a significant increase beyond 30 m. Different antennas exhibit varying performances in propagation routing, as depicted in Figure 4.14(a), where the omni-directional antenna broadcasts signals around the transmitter site. Due to the vehicle's metal structure, rays from the omni-directional antenna experience more reflections inside the vehicle, resulting in higher ray amplitude and delay spread compared to directional antennas of similar height. In far-field areas, the omni antenna transmits more diverse paths of rays, which explains why its delay spread is higher than that of a directional patch antenna.

The peak delay spread occurs at shorter distances with the same directional antenna placed at different heights. The delay spread is higher for antennas placed at higher levels compared to those placed at middle and low heights. This trend can be attributed to the increased propagation path diversity of higher-level antenna placements compared to lower positions. As the distance between the transmitter and receiver increases, the delay spread gradually decreases.

The Table 4.10 presents the numerical results obtained from Figure 4.16. The higher placement of the antenna leads to a greater delay spread due to path ray diversity. As introduced in Section 2.3.2, the coherence bandwidth and coherence time can be calculated by Eq. (2.18) and Eq. (2.20). Thus the calculated 0.9 coherence bandwidth at an antenna height of 0.6 m, 1.0 m, and 1.6 m are 2.06 MHz,

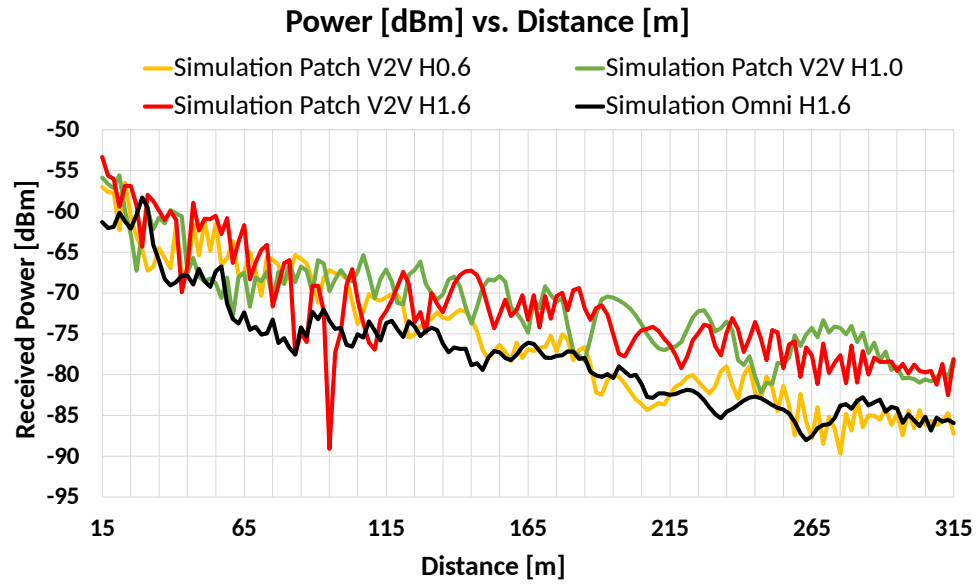


Figure 4.15: Receiver power through Route 0 trajectory with different antenna configuration

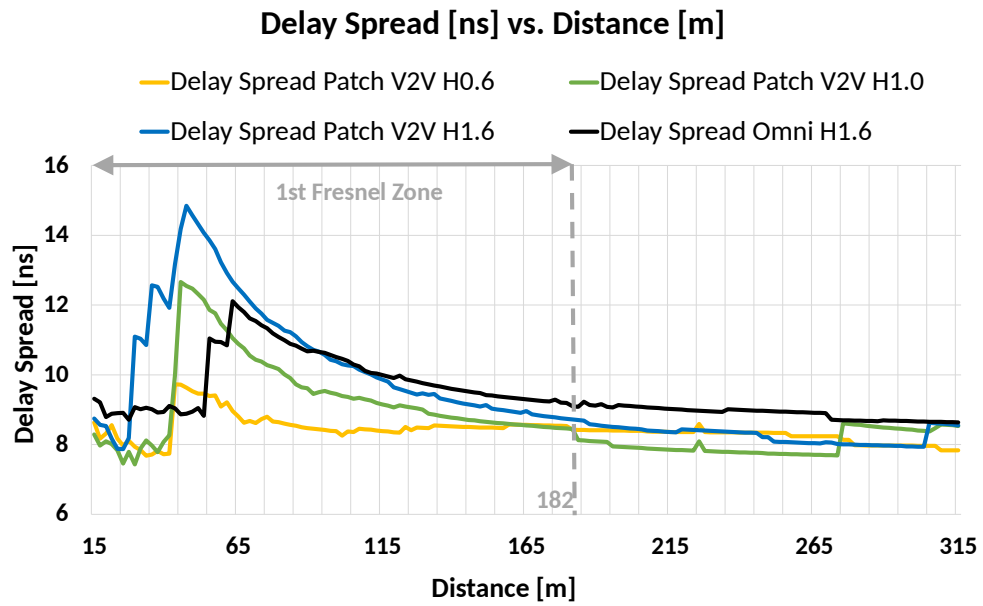


Figure 4.16: Delay spread vs. distance on simulation at rural scenario

Table 4.10: RMS delay spread with different antenna height configuration

	V2V H0.6	V2V H1.0	V2V H1.6	Omni H1.6
Delay Spread Average	8.41 ns	8.76 ns	9.49 ns	9.45 ns
Delay Spread STDEV	0.37 ns	1.14 ns	1.70 ns	0.82 ns
Delay Spread 50%	8.39 ns	8.50 ns	8.75 ns	9.09 ns
Delay Spread 95%	9.31 ns	11.38 ns	13.21 ns	11.14 ns
Maximum Delay Spread	9.72 ns	12.57 ns	14.85 ns	12.11 ns
0.9 Coherence BW	2.06 MHz	1.58 MHz	1.35 MHz	1.65 MHz
OFDM N=64 Spacing	0.156 MHz			

1.58 MHz, and 1.35 MHz respectively, while for the omni antenna at a height of 1.6 m it is simulated as 1.65 MHz. The V2V application operates within the frequency band of 5905-5915 MHz with a bandwidth of 10 MHz. The modulation method used for V2V transmission is OFDM which employs N=64 subcarriers with each carrier channel having a bandwidth of $10\text{MHz} \div 64 = 0.1562\text{ MHz}$, from Table 2.1. Due to the benefits of OFDM technology, the coherence bandwidth (the narrowest coherence bandwidth (BW) of a directional patch antenna at a height of 1.6 m is 1.35 MHz) is much wider than each subcarrier channel's spacing (0.156 MHz). This is one advantage of novel code modulation in maintaining signal propagation in a state of flat fading rather than frequency-selective fading.

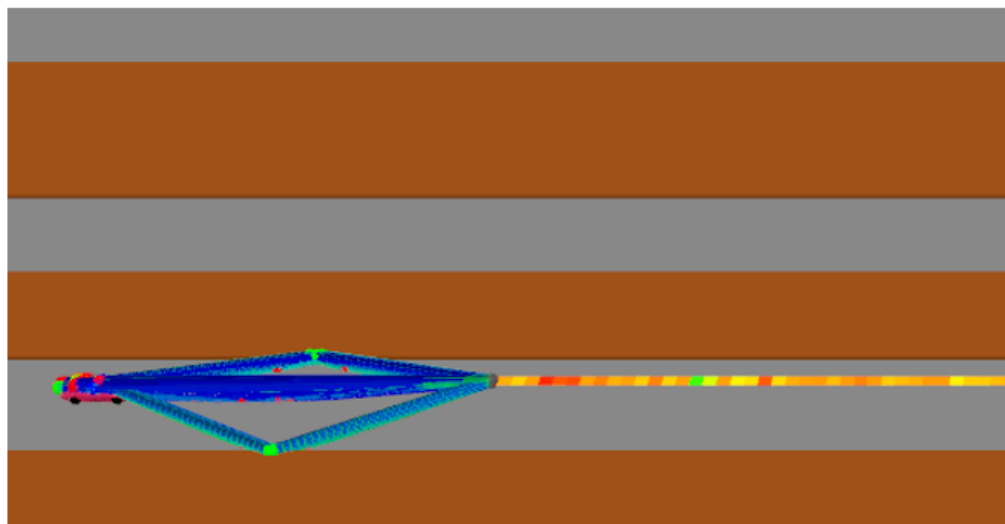
Intersymbol interference (ISI) occurs when OFDM symbols overlap in time at the receiver. To avoid ISI, the cyclic prefix (guard interval) should exceed the channel maximum delay spread T_m [22],

$$T_{\text{GI}} > T_c \approx T_m \quad (4.5)$$

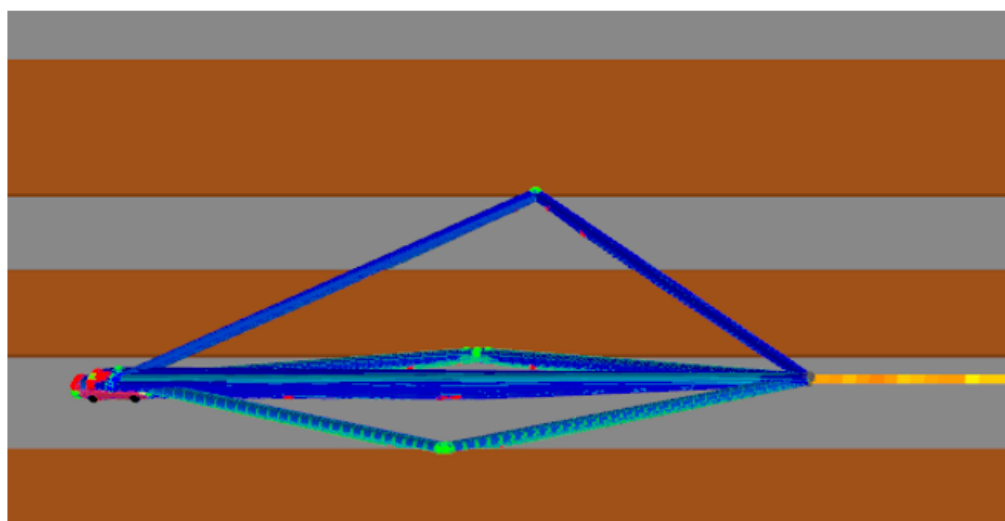
where $T_{GI} = 1.6\mu s$ from Table 2.1 , $T_m = 14.85ns$ of the maximum delay spread of V2V antenna at 1.6 m.

As an illustration of antenna height at 1.6 m, Figure 4.17 (a) and (b) depict sudden changes in delay spread at distances of 27 m and 49 m. Table 4.11 presents the top three rays at a distance of 27 m with an antenna height of 1.6 m. The primary ray, Ray 1 (0R,0D,2T,0S), indicates a path with two windshield transmissions followed by direct transmission to the receiving point. The second ray, Ray 2 (1R,0D,2T,0S), includes two transmissions and an additional reflection on the ground; while the third ray, Ray3 (0R,1D,1T,0S), involves one diffraction from the side grass loam and one windshield transmission. In this simulation case, a maximum of 100 rays per pixel are calculated; other rays exhibit similar levels of delay.

Referring to Figure 4.18, when comparing the receiving signal configuration at 27 m, there is a significant increase in delay spread from 8.56 ns to 14.57 ns at 49 m. The simulation result (Figure 4.17b) indicates that Ray 38, 53, 67, and 85 experience diffractions from the opposite side of the boulevard. Table 4.12 provides the double distance of the delay values as per Eq.(2.16), which explains the reason for the increase in delay spread. In rural scenarios simulations, it is observed that delay spread is correlated with ray propagation path routing, resulting in different delays in each snapshot due to diverse paths caused by complex surrounding environments contributing to excess delay bins analysis.



(a)



(b)

Figure 4.17: (a). The ray illustration at 27 m with antenna height at 1.6 m;
 (b). The ray illustration at 49 m with antenna height at 1.6 m.

Table 4.11: The ray detail at 27 m with antenna height at 1.6 m

Ray	Interactions	Delay	Phase	Signal Level
1	2 (0R,0D,2T,0S)	95.13 ns	104.18 deg	-61.28 dBm
2	3(1R,0D,2T,0S)	95.67 ns	353.56 deg	-66.67 dBm
3	2(0R,1D,1T,0S)	95.14 ns	290.07 deg	-67.30 dBm

Table 4.12: The ray detail at 49 m with antenna height at 1.6 m

Ray	Interactions	Delay	Phase	Signal Level
1	2(0R,0D,2T,0S)	168.52 ns	90.80 deg	-66.25 dBm
2	3(1R,0D,2T,0S)	168.52 ns	200.91 deg	-69.24 dBm
3	2(0R,1D,1T,0S)	168.52 ns	273.04 deg	-72.81 dBm
...				
38	4(1R,1D,2T,0S)	232.90 ns	45.72 deg	-88.72 dBm
53	4(1R,1D,2T,0S)	232.91 ns	51.79 deg	-91.08 dBm
67	4(1R,1D,2T,0S)	232.92 ns	91.20 deg	-93.94 dBm
85	4(1R,1D,2T,0S)	232.88 ns	166.86 deg	-98.30 dBm

The received power and delay spread versus distance for the V2V antenna at a height of 1.6 m are shown in Figure 4.18. It can be clearly observed that there is a sudden change in the delay spread and fading of the received power as distance increases, indicating that large-scale fading becomes the dominant factor in wave propagation. The delay spread is below 8 ns after passing through the first Fresnel zone.

4.5 Transmission through the Windshield

In Table 4.11 and Table 4.12, there is a type of path called T (transmission) that passes through a dielectric material, which should not be overlooked. It can be observed that all the rays contain at least one or two transmissions when they pass

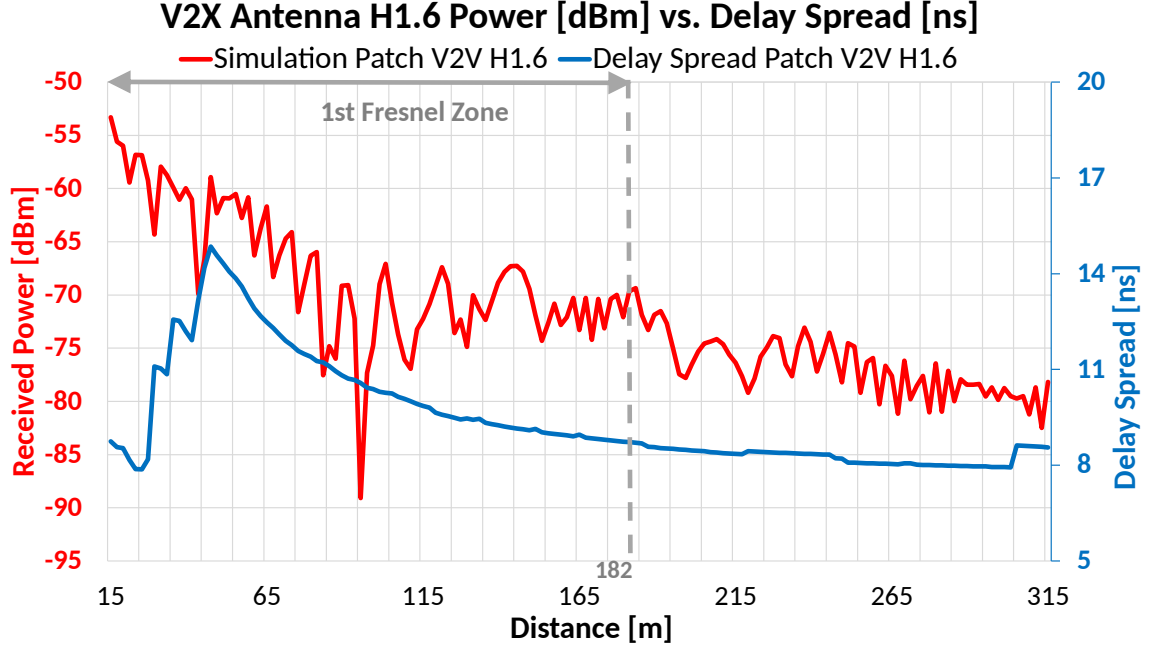


Figure 4.18: Simulation received power and delay spread on antenna height at 1.6 m

through the glass. According to Figure 2.14, an incident wave hits the starting point of the surface of different dielectric materials (initial hits), and some waves are transmitted while others are reflected and then re-transmitted. This phenomenon requires revising the simulation with an integrated windshield and inner bracket CAD file. By utilizing CST Asymptotic Solver[®] (Figure 4.19), the radiation pattern obtained from Section (4.2) can be used as a field source in the simulation boundary condition (highlighted in blue box).

The ray-tracing technique can provide accurate channel statistics by considering the actual physical propagation [74][75]. Figure 4.20 illustrates the propagation path of the rays up to two orders. Figure 4.21 presents the far-field radiation directivity results without and with glass effects. Without the glass, the

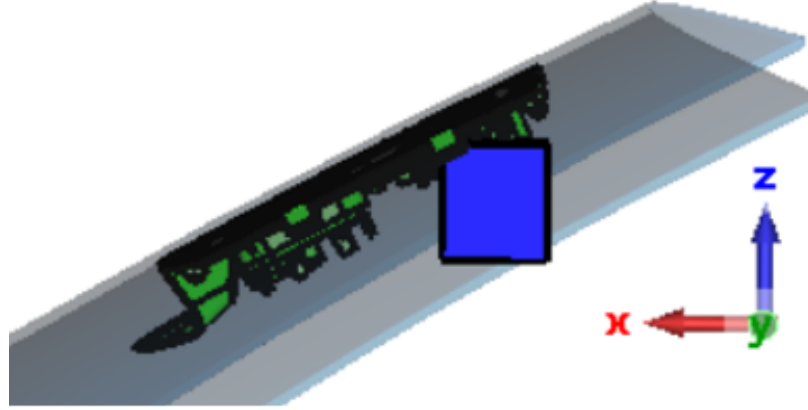


Figure 4.19: Field source antenna simulation with integrated windshield in CST

antenna pattern is round and smooth, with a 3 dB beam width of up to 110° and a maximum directivity of 5.96 dBi. However, when the windshield and surrounding attachments are present, ripples appear at the beam edge due to phase construction and deconstruction of transmission waves. There is a null in the beam at an angle of $\theta = 125^\circ$, which corresponds to tangency with the windshield surface. The presence of this dielectric material causes small ripples in far-field gain within a certain range of Theta angles.

The ray path through the glass side view can be seen in Figure 4.22(a) and (b), which includes observations up to 2 hit orders and E-Field amplitude. In Figure 4.22(a), the blue/green/red rays represent direct paths of 0/1/2 hit orders, respectively. For example, when a ray hits the dielectric surface, first-order rays reflect from the material as shown by the green curve. As indicated by E-Field amplitude in Figure 4.22(b), reflection path energy is not dominant (illustrated in cold colors), and most of it would continuously go through second-hit order as shown by warm color rays on top of the image and amplitude levels on the color

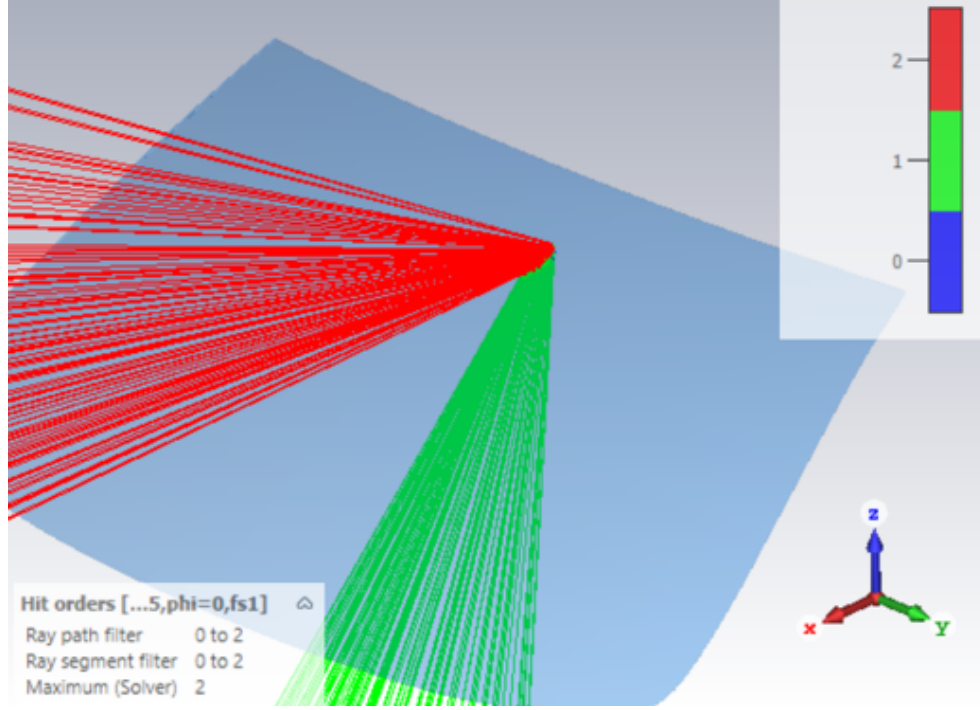


Figure 4.20: Wave propagation up to 2 order rays in windshield simulation

map. The CST Asymptotic Solver[®] ray-tracing analysis provides additional information from Feko WinProp[®] rural vehicle scenario simulation, with both software tools providing coherent results that mutually validate each other regarding glass effect on antenna radiation performance.

4.6 On-vehicle Field Testing

The simulation work provides a method to analyze the performance and channel information of the V2V antenna. However, verification of multipath propagation needs to be conducted through real-world vehicle testing. The test setup takes place in an open-field rural area with a two-direction road and two car lines in each direction. This scenario represents a typical rural environment, with

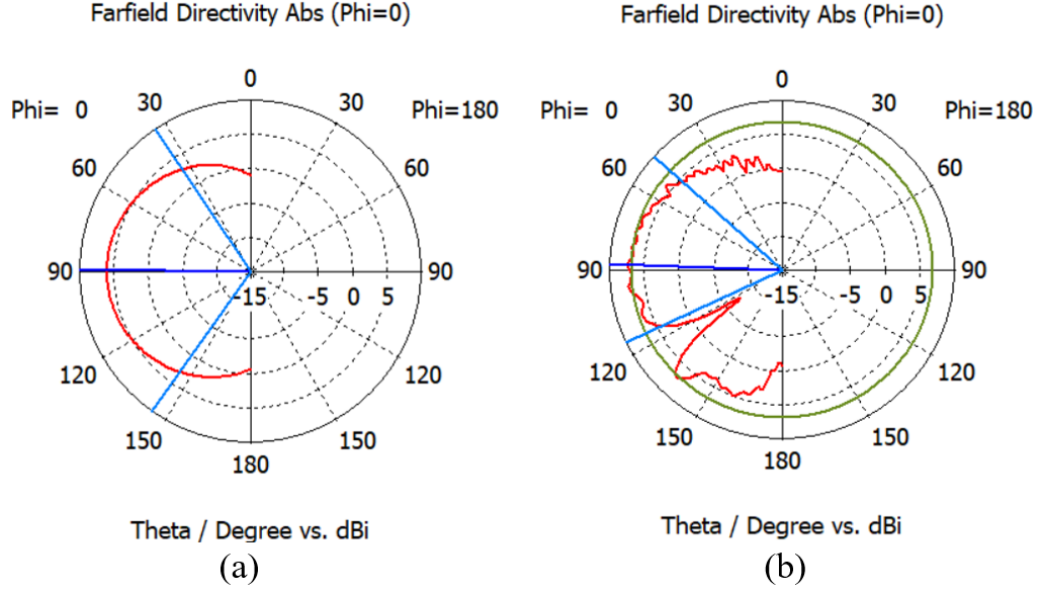
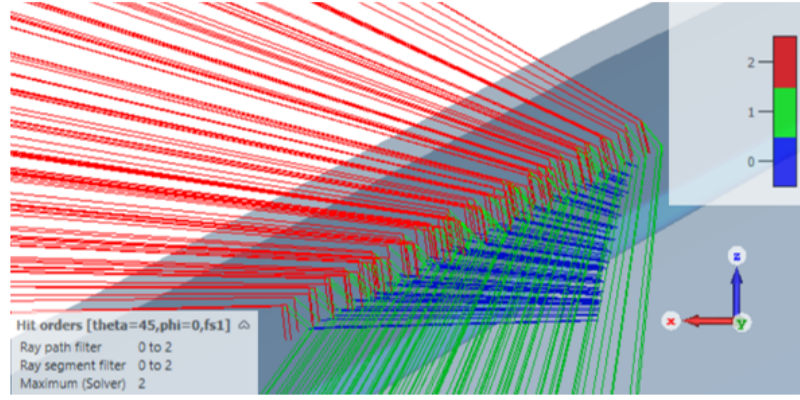


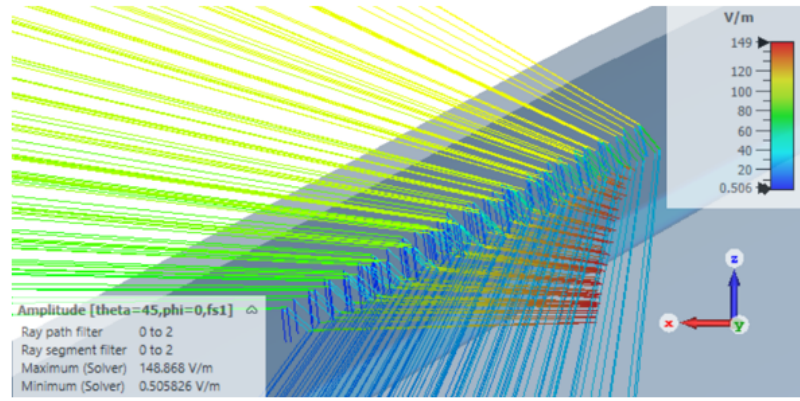
Figure 4.21: Simulated far-field directivity radiation pattern;
(a) without glass;
(b) with glass

surroundings similar to those in the previous simulation scenario (Figure 4.13). This trial work is part of China C-V2X Large-scale Pilot Test Activity held in the City of Suzhou, China.

A stationary car is positioned at the left end of the road, with a Tx antenna mounted on the front windshield, as indicated in Figure 4.23(a), which is designed and analyzed in Section 4.2. A rover is located at a starting point with an Rx V2V antenna pointed towards the Tx antenna. The distance between the Tx antenna and Rx antenna is 15 m, which matches the simulation scenario setup. In Figure 4.23(b), the red line represents the moving trajectory until it reaches a distance of 315 m away. During vehicle testing, consideration is given to the height factor of



(a)



(b)

Figure 4.22: (a) Side observation view of rays up to 2 hit orders;
(b) Rays E-Field amplitude at Theta 45 degree

receiving antenna. The Rx antenna height configuration options are set at 0.6 m, 1.0 m, and 1.6 m respectively, as shown in Figure 4.24.

The configuration of both Tx and Rx antennas is the same as the Feko WinProp[®] simulation in Table 4.8, with receiving power ranging from -50 to -95 dBm over a distance of 315 m. As shown in Figure 4.25, when the distance between Tx and Rx antenna increases, the receiving power attenuates further. By comparing



(a)



(b)

Figure 4.23: On-vehicle field testing setup;
 (a) Transmitting antenna (Tx Ant) and Receiving antenna (Rx Ant) setup on the vehicle;
 (b) A trajectory demonstration of road test on a rural area.



0.6 m



1.0 m



1.6 m

Figure 4.24: Receiving antenna height placements.

three levels of antenna heights, it can be observed that at far-field distances, the receiving power of the low-level antenna ($H = 0.6$ m) is significantly lower than that of the middle-level ($H = 1.0$ m) and high-level ($H = 1.6$ m) antennas. The average power is calculated in Table 4.13. In the far-end region, the measured average receiving powers are -78.47 dBm, -81.43 dBm, and -84.55 dBm for low to high placements respectively.

As a reference antenna with a height of 1.5 m, corresponding antenna factors are calculated as -5.92 dB, -2.80 dB, and 0.16 dB respectively which are in good agreement with reference SAE J3161/1A [72]. This degradation factor due to low antenna placement height could impact far-field transmitting power and received sensitivity. Thus, this height factor should be considered during vehicle design stage along with V2V communication link budget calculation.

The results are applicable to the far region with proper antenna height calculation. However, they may not consider the small-scale fading in the near region, where the signal power undergoes dynamic changes over short distances. Higher antenna placement leads to deeper and longer fadings. The fading areas are highlighted in black, yellow, and blue in Figure 4.25, representing antenna placements of 0.6 m, 1.0 m, and 1.6 m respectively. Consequently, the minimum receiving power is -70.53 dBm, -79.80 dBm and -83.07 dBm in the near-field region.

The following Figure 4.26 (a) (b) and (c) illustrate the receiving power of ground reflection model, rural area simulation and real-world field measurement, with Rx antenna heights at 0.6 m, 1.0 m and 1.6 m respectively. The three results, as observed from Figure 4.26 (a), show a strong agreement with the receiving power over distance. In the near region (15 - 68 m), fluctuations in receiving power are evident, indicating small-scale fading in the V2V communication system. The black curve result reveals that the fading occurs between distances of 31 to 39 m, with a

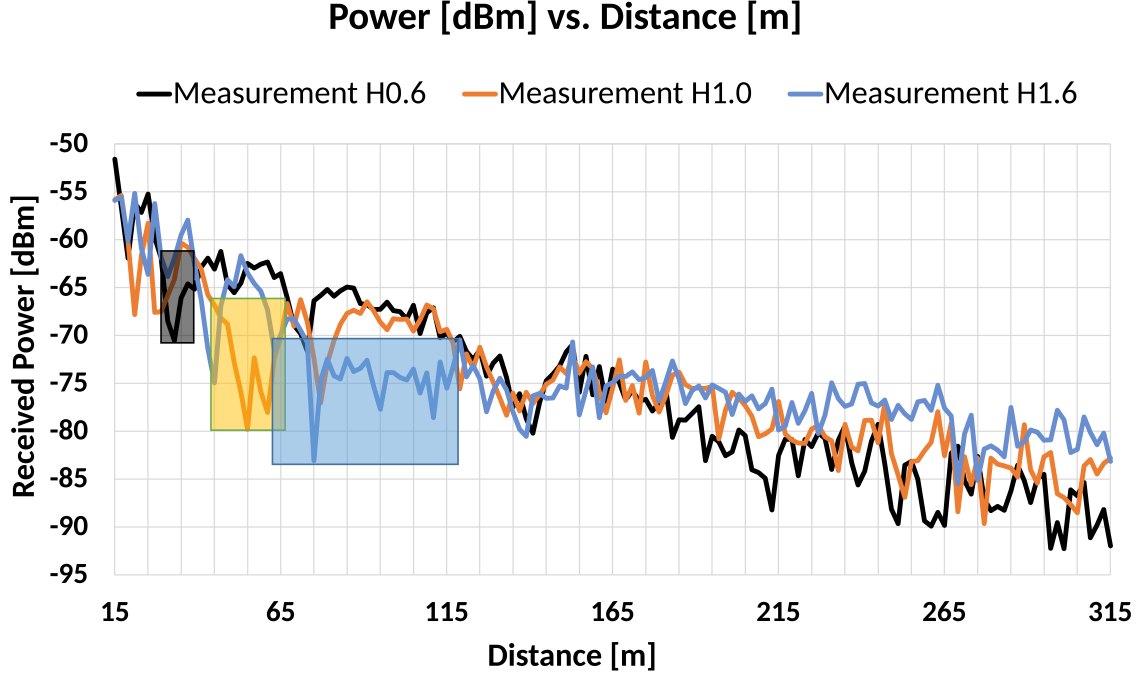


Figure 4.25: Receiving power vs. distance at three different antenna height configurations.

minimum receiving power of -70.53 dBm highlighted in the black area. Theoretical calculations and simulations also demonstrate similar behaviors.

Meanwhile, the antenna height result at 1.0 m is presented in Figure 4.26 (b), which exhibits similar behaviors in both the near and far regions. The fading area will extend from 41 to 65 m, with a minimum value of -79.80 dBm.

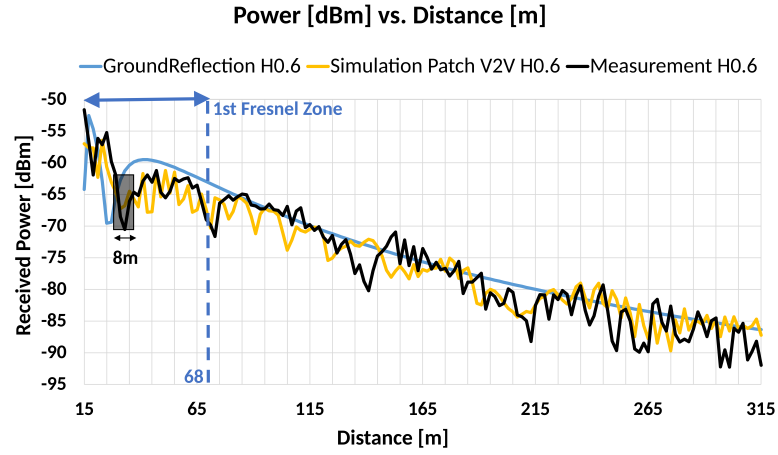
As the antenna height increases to 1.6 m, as shown in Figure 4.26 (c), the propagation channel experiences longer and deeper fades in both the near region and fading area. For distances between 63 and 117 m, the signal persists over a long distance within a deep fading zone and the minimum receiving power of -83.07 dBm. It presenting a significant challenge for maintaining communication

Table 4.13: Average receiving power vs. antenna height factor

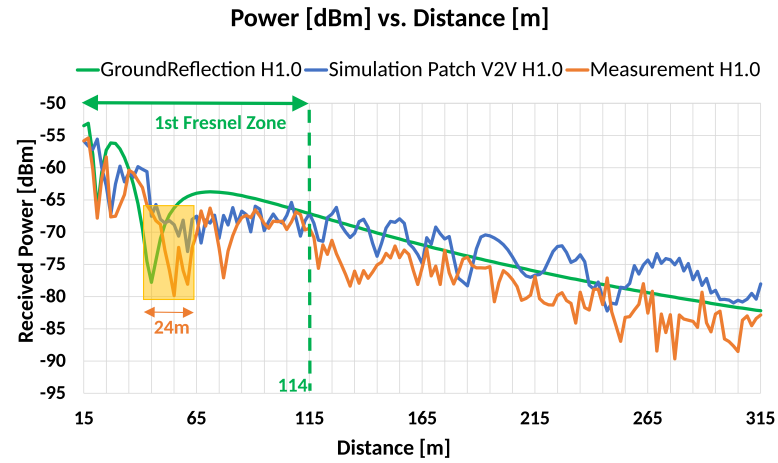
	Average Rx Power (182-315 m)		
	H 0.6 m	H 1.0 m	H 1.6 m
2-ray Simulation	-82.11 dBm	-78.06 dBm	-74.89 dBm
Rural Simulation	-83.16 dBm	-76.24 dBm	-76.68 dBm
Measurment	-85.55 dBm	-81.43 dBm	-78.47 dBm
Antenna Factor: reference height 1.5 m			
Meas. Antenna Factor	-5.92 dB	-2.80 dB	0.16 dB
Ref. Antenna Factor [72]	-6.80 dB	-2.80 dB	0.40 dB

connection status in such a rapidly fluctuating area. The weak signal connection makes the communication signal susceptible to distortion and interruption.

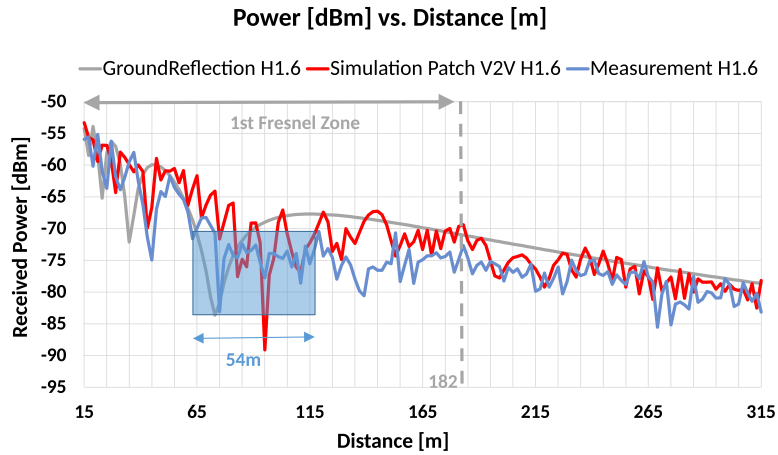
The ground-reflection model has a good prediction ability for large-scale fading, and the path loss suffers continual attenuation in the far-end area. The rural scenario ray-tracing simulation is in good agreement with real-world measurements on both short and long travel distances. In particular, the simulation result could predict a kind of small-scale fading characterized by rapid fluctuations of the received signal power. The most important factor for antenna position selection is that higher antennas have better gain benefits in far regions, while they have more small-scale fading impact in near regions, vice versa. Thus, the antenna should be placed at a middle level on vehicles to balance both short and long distance communication performance.



(a) antenna height at 0.6 m;



(b) antenna height at 1.0 m;



(c) antenna height at 1.6 m;

Figure 4.26: Receiving power of two-ray ground-reflection, rural scenario simulation and real-world measurement with 1st Fresnel zone

4.7 Summary

V2X technology has become an essential element of modern automotive driving, and ensuring a stable communication link for end-to-end transmission is crucial. The channel environment and transmission conditions have undergone changes due to the utilization of new frequency bands and application scenarios. This chapter presents a comprehensive study on Direct Communication PC5 mode 4, considering the latest advancements in antenna design within the industry. It concludes that antenna placement plays a significant role in overall RF link budget calculation, emphasizing the importance of considering antenna location during the carline design stage. Conducted in a rural area, V2V channel analysis encompasses theoretical analysis, virtual ray-tracing simulation, and real-world on-vehicle testing to provide insights into actual installation of RF antenna in vehicles. Additionally, research on channel fading includes small-scale fading transmission within the first Fresnel zone and far-end large-scale fading transmission. By analyzing closed area channel delay spread and OFDM frame characteristics, this study explains how V2V code modulation system can overcome frequency-selective fading. Integrating antenna design with vehicle-specific channel propagation analysis will serve as a guideline for subsequent V2V antenna design.

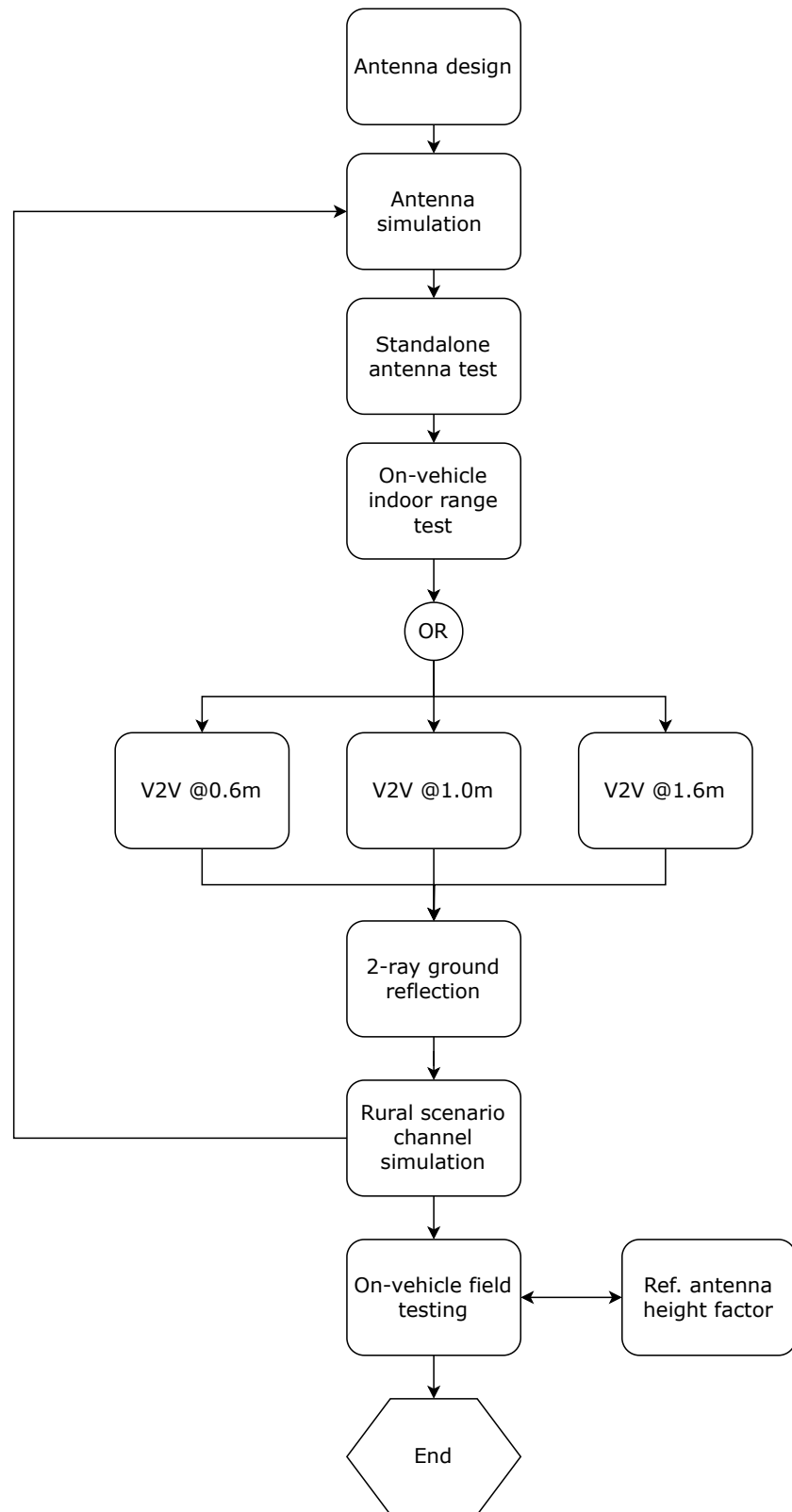


Figure 4.27: Research flowchart

CHAPTER FIVE

GENERAL CONCLUSION

This dissertation presently focused on two main areas: the characterization of phase center for multi-band GNSS antenna and the design of V2V antenna, along with channel analysis in rural propagation scenario. The research highlights the importance of high-precision positioning systems and reliable C-V2X wireless networks, particularly in the automotive Telematics Control Unit (TCU).

A low-cost multi-band L1/L2/L5 GNSS patch antenna to characterize its PCV and PCO in a standalone patch antenna and in a shark-fin multi-function antenna vehicle level at indoor and outdoor measurement ranges. Due to the shark-fin antenna's complex interior configuration, such as nearby radiator, which detunes the GNSS patch antenna electrical wall and magnetic wall boundary condition. By the impact of that, the GNSS antenna's PCV and PCO results exhibited worse performance than the standalone antenna on a ground plane. A proper solution is to fix phase-center motion error $\varepsilon(r, PCV)$ and $\varepsilon(r, PCO)$ in high precise positioning application. Each unique vehicle platform and antenna location requires the calibration profile of PCV and PCO. By using this antenna calibration file (.atx) could estimate the antenna phase-center motion error-free result. In high accuracy positioning, single, double and triple difference could eliminate the major error sources, except antenna phase error. On Chapter 3 presents a method to solve phase error $\varepsilon(r, PCV)$ and $\varepsilon(r, PCO)$ in Eq.(2.61) in automotive applications. From a classic method to evaluate the antenna by using gain and axial ratio, moreover by using the far-field phase data to show the PCV and calculated PCO based on the PCV result. Although gain and axial ratio results do not contain the phase

information, they have a strong correlation with PCV and PCO result. A better gain and axial ratio imply a better phase result as well. They provide an alternative way to evaluate PCV and PCO quickly at early of the development stage. As the requirement of precise GNSS navigation evolves, it requires PCV measurement and PCO calculation at the validation stage.

As V2X technology becomes an essential part of current automotive driving, a stable communication link ensures an end-to-end link. The channel environment and transmission condition have changed under the new frequency band and new application scenarios. This study conducted a profound study of V2V communication with received power level analysis in a rural channel environment as function of antenna type and height via simulation and field trails. The antenna design considered the latest industry developments and concluded the antenna placement is a significant parameter in the whole Radio Frequency (RF) link budget calculation and more importantly the antenna location should be considered in advance early in the design stage. Based on a rural area scenario, V2V channel characteristic is performed via theoretical analysis, virtual ray-tracing simulation and real world on-vehicle testing with RF antenna in actual vehicle installations. In this research, channel fading is analyzed in both the small-scale fading transmission in the 1st Fresnel zone and the far-end large-scale fading transmission. The combination of mid-level antenna placement and integrated vehicle channel propagation analysis provided a guideline for the subsequent automotive manufacture V2V antenna design.

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