

PERFORMANCE EVALUATION AND EXPERIMENTAL VERIFICATION OF VEHICULAR VISIBLE LIGHT COMMUNICATION

A Dissertation

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PERFORMANCE EVALUATION AND EXPERIMENTAL VERIFICATION OF VEHICULAR VISIBLE LIGHT COMMUNICATION

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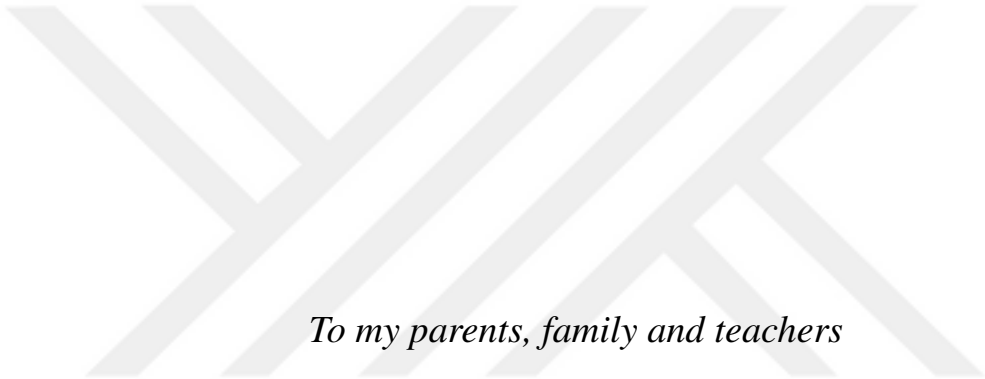
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To my parents, family and teachers

ABSTRACT

Vehicular development relies heavily on data communications for supporting some future applications. Vehicle platoon and Autonomous Vehicles (AV) are some of the applications that will be commercially available in this decade. From that point, Intelligent Transportation System (ITS) become one of the most attractive fields for the applications of wireless communications where Decentralized Environmental Notification Messages (DENM), Co-Operative Awareness Messages (CAM) or any application-specific messages could be exchanged among vehicles. Those messages could be further utilized by traffic management which will participate in reducing both trip time and environmental pollution. This motivates automotive manufacturers to rely on vehicular connectivity to catch up with the market development through technology implementation. In ITS applications, a reliable, high data rate and low latency communication link is required. Dedicated Short-Range Communication (DSRC) and Cellular vehicle-to-everything (CV2X) are the conventional Radio Frequency (RF) technology for ITS applications. Due to the RF bandwidth limitation and Electromagnetic Interference (EMI) issues, Visible Light Communication (VLC) has emerged as powerful wireless access technology by offering a large license-free bandwidth and immunity to EMI. VLC is a possible candidate for Vehicle-to-Vehicle (V2V), Vehicle-to-Infrastructure (V2I) as well as Vehicle-to-Pedestrian (V2P) applications. Particularly, VLC has a robust performance in user-dense environments and is considered a complementary technology.

The automotive manufacturers have recently increased the use of Light-Emitting Diodes (LEDs) in the vehicle's bodies. Since VLC allows the dual use of LED for illumination and communication purposes, this allows VLC to be considered as a complementary approach for vehicular connectivity. VLC system modulates the intensity of the light source by superimposing the desired information signal upon the driving Direct Current (DC) which is selected according to the desired operating point taking

into account the amplitude constraints. Typical modulation choice for VLC systems is On-Off Keying (OOK) and pulse modulation techniques while other modulation techniques such as Orthogonal Frequency Division Multiplexing (OFDM) have been further proposed to support ultra-high-speeds on the order of Gbps.

There are some concerns for practically implementing the Vehicle-to-Everything (V2X) VLC applications. For instance, the practical implementation aspects and the hardware availability. Despite the shortage in the commercially available products for vehicular VLC, the frontend has to be designed and its effect should be considered in the system implementation. Moreover, the effect of the outdoor environment, such as sunlight, has to be characterized for accurate system design. In addition, as in any communication system, channel modelling plays a critical role in vehicular VLC systems, especially with the asymmetrical pattern of the vehicular light source, it needs to be experimentally modelled. Another critical aspect is the effect of mobility and the location of the photodetector on the vehicle must be properly identified for having a robust vehicular VLC link in different road shapes. As a benchmark, the performance of vehicular VLC should be further compared with RF.

Motivated by these, we firstly investigate the baseband processing and the experimental implementation of different modulation techniques for VLC systems. After that, we design a frontend that supports high power transmission of vehicular LEDs and investigate its experimental effects on the system performance. We then characterize the impairments of the outdoor environment and identify the proper choice of the lens combination that suits the vehicular environment. Additionally, we experimentally model the pathloss of the vehicular asymmetrical light source. This model is further extended to be used for a special case (truck-to-truck VLC system) of the truck's vertical oscillation. Counting on these, we investigate the critical aspect of the vehicular VLC link performance in the presence of mobility for passenger vehicles as well as for heavy vehicles (i.e., Truck). For benchmarking purposes, we further implement a hybrid system (VLC/ RF) for comparison purposes as well as deploy a switching algorithm to satisfy a specific Quality of Service (QoS).

ÖZETÇE

Araç geliştirme, gelecekteki bazı uygulamalarını desteklemek için büyük ölçüde veri iletişimine aktarır. Araç müfrezesi ve otonom araçlar (AV), bu on yılda ticari olarak temin edilebilecek uygulamalardan bazılarıdır. Bu noktadan sonra, akıllı ulaşım sistemleri (ITS), merkezi olmayan çevresel bildirim mesajlarının (DENM), kooperatif farkındalık mesajlarının (CAM) veya herhangi bir uygulamaya özel mesajın birbirleriyle değiş tokuş edilebileceği kablosuz iletişim uygulamaları için en çekici alanlardan biri haline geldi. araçlar. Bu mesajlar, hem yolculuk süresinin hem de çevre kirliliğinin azaltılmasına katkıda bulunacak trafik yönetimi tarafından daha fazla kullanılabilir. Bu, otomotiv üreticilerini, teknoloji uygulaması yoluyla pazar gelişimini yakalamak için araç bağlantısına güvenmeye motive ediyor. ITS uygulamalarında son derece güvenilir, yüksek veri hızı ve düşük gecikmeli bir iletişim bağlantısı gereklidir. Özel kısa menzilli iletişim (DSRC) ve Hücreli araçtan her şeye (CV2X), ITS uygulamaları için geleneksel radyo frekansı (RF) teknolojisidir. Mevcut RF spektrumunun bant genişliği sınırlaması ve elektromanyetik girişim (EMI) sorunları nedeniyle, görünür ışık iletişimi (VLC), lisanssız geniş bir bant genişliği ve EMI'ye karşı bağışıklık sunarak güçlü kablosuz erişim teknolojisi olarak ortaya çıkmıştır. Özellikle VLC, kullanıcı yoğun ortamlarda güçlü bir performansa sahiptir ve veri boşaltma yoluyla WiFi'ye tamamlayıcı bir teknoloji olarak kabul edilir. Ayrıca VLC, hastane yoğun bakım üniteleri, nükleer santraller gibi RF kısıtlı alanlarda güvenli bir alternatif olarak kullanılabilir.

Otomotiv üreticisi son zamanlarda araç gövdelerinde ışık yayan diyotların (LED'ler) kullanımını artırıyor. VLC, aydınlatma ve iletişim amaçları için LED'in ikili kullanımına izin verdiğinden, bu, VLC'nin araç bağlantısı için tamamlayıcı bir yaklaşım olarak değerlendirilmesini sağlar. VLC sistemi, istenen bilgi sinyalinin sürüş doğru akımı (DC) üzerine bindirerek ışık kaynağının yoğunluğunu modüle eder. DC

değeri, genlik kısıtlamaları dikkate alınarak istenilen çalışma noktasına göre seçilir. Modüle edici sinyalin frekansı çok yüksek olduğundan, ı ışık yoğunluğunun titremesi insan gözü tarafından gözlemlenemez. VLC sistemleri için tipik modülasyon seçimi, açma-kapama anahtarlama (OOK) ve darbe modülasyon teknikleri iken, dikey frekans bölmeli çoğullama (OFDM) gibi diğer modülasyon teknikleri, Gbps düzeyinde ultra yüksek hızları desteklemek için ayrıca önerilmiştir.

V2X uygulaması için VLC bağlantısının pratik uygulaması için bazı endişeler var. Örneğin, pratik uygulama yönleri. Araç VLC'si için ticari olarak temin edilebilen ürünlerdeki kıtlığa rağmen, ön uç tasarlanmalı ve sistem tasarımında etkisi göz önünde bulundurulmalıdır. Ayrıca, güneş ışığı gibi dış ortamın etkisi karakterize edilmelidir. Ayrıca, herhangi bir iletişim sisteminde olduğu gibi, kanal modelleme araç VLC sistemlerinde kritik bir rol oynar, özellikle araç ışık kaynağının asimetrik modeli ile deneysel olarak modellenmesi gerekir. Diğer bir kritik husus, hareketliliğin etkisidir ve fotod-epektörün araç üzerindeki konumu, farklı yol şekillerinde (yani düz ve kavisli yollar) sağlam bir araç VLC bağlantısına sahip olmak için uygun şekilde tanımlanmalıdır. Bir kıyaslama olarak, araç VLC'sinin performansı RF ile daha fazla karşılaştırılmalıdır.

Bunlardan motive olarak, öncelikle VLC sistemleri için temel bant işleme ve farklı modülasyon tekniklerinin deneysel uygulamasını araştırıyoruz. Daha sonra, araç LED'lerinin yüksek güç iletimini destekleyen önyüz tasarlıyor ve sistem performansı üzerindeki deneysel etkilerini araştırıyoruz. Daha sonra dış ortamdaki bozulmaları ve araç ortamına uygun lens kombinasyonunun doğru seçimini karakterize ediyoruz. Araç VLC sistemi için bir diğer kritik husus, araç asimetrik ışık kaynağının yol kaybını deneysel olarak modelleyerek kanal modelinin ayrıca ele alınmasıdır. Bu model, özel durum için kullanılmak üzere daha da genişletildi (kamyonlardan kamyonlara VLC sistemi). Bunlara güvenerek, hem binek araçlar hem de ağır araçlar (yani Kamyon) için hareketlilik varlığında araç VLC bağlantı performansının kritik yönünü araştırıyoruz. Kıyaslama amacıyla, ayrıca karşılaştırma amacıyla bir hibrit sistem (VLC/RF) uyguluyor ve belirli bir hizmet kalitesini (QoS) karşılamak için bir anahtarlama algoritması uyguluyoruz.

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ACRONYMS

ACO-OFDM	Asymmetrically-Clipped Optical Orthogonal Frequency Division Multiplexing
APD	Avalanche Photodiode
API	Application Programming Interface
AV	Autonomous Vehicles
AWG	Arbitrary Waveform Generator
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
CAM	Co-Operative Awareness Messages
CIR	Channel Impulse Response
CINR	Carrier to Interference Noise Ratio
CP	Cyclic Prefix
CV2X	Cellular vehicle-to-everything
DAC	Digital to Analog Conversion
DC	Direct Current
DCO-OFDM	Direct Current Optical Orthogonal Frequency Division Multiplexing
DENM	Decentralized Environmental Notification Messages
DSRC	Dedicated Short-Range Communication
EMI	Electromagnetic Interference
FEC	Forward Error Correction
FFT	Fast Fourier Transform
FOV	Field of View
FPGA	Field Programmable Gate Arrays
IFFT	Inverse Fast Fourier Transform
IM/DD	Intensity Modulation direct-detection

ITS	Intelligent Transportation System
LED	Light-Emitting Diode
LFRX	Low Frequency Receiver
LFTX	Low Frequency Transmitter
LoS	Line of Sight
ML	Maximum Likelihood
mmW	millimeter Wave
NLoS	Non-Line of Sight
OFDM	Orthogonal Frequency Division Multiplexing
OOK	On-Off Keying
PDF	Probability Density Function
PCB	Printed Circuit Board
PDF	Probability Density Function
PDR	Packet Delivery Ratio
PHY	Physical Layer
PRBS	Pseudorandom Binary Sequence
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
RF	Radio Frequency
RRC	Root-Raised-Cosine
RSSI	Received Signal Strength Indication
SC	Selection Combining
SHR	Synchronization Header
SNR	Signal-to-Noise Ratio
U-OFDM	Unipolar Orthogonal Frequency Division Multiplexing
USRP	Universal Software Radio Peripheral
V2I	Vehicle-to-Infrastructure
V2P	Vehicle-to-Pedestrian
V2V	Vehicle-to-Vehicle

V2X	Vehicle-to-Everything
VLC	Visible Light Communication
WiFi	Wireless Fidelity



TABLE OF CONTENTS

DEDICATION	iii
ABSTRACT	iv
ÖZETÇE	vi
ACKNOWLEDGEMENTS	viii
ACRONYMS	ix
LIST OF TABLES	xiv
LIST OF FIGURES	xv
I INTRODUCTION	1
1.1 Overview on Vehicular Connectivity	1
1.2 Literature Review and Motivation	2
1.3 Dissertation Outline	4
1.4 Organization	5
II SIGNALLING MODELS FOR VLC SYSTEMS	7
2.1 Introduction	7
2.2 OFDM-Based VLC System	8
2.3 OOK-Based VLC System	17
2.4 Selection of Proper Lens Combination	27
2.5 Conclusion	31
III MULTIHOP TRANSMISSION FOR END-TO-END CONNECTIVITY	32
3.1 Introduction	32
3.2 System Model	33
3.3 Effect of Outdoor Noises	35

3.4	Error-Rate Analysis	36
3.5	Conclusion	41
IV	VEHICULAR VLC LINKS IN THE PRESENCE OF MOBILITY . . .	42
4.1	Introduction	42
4.2	System Model	43
4.3	Measurement Campaign and Analysis	47
4.4	BER Analysis and Measurements	51
4.5	Conclusion	55
V	VEHICULAR VLC LINKS IN THE PRESENCE OF OSCILLATION .	56
5.1	Introduction	56
5.2	Vehicular VLC Channel Modelling	56
5.3	Vehicular Vertical Oscillation Modelling	63
5.4	Error Rate Performance	67
5.5	Results and Discussion	69
5.6	Conclusion	75
VI	HYBRID VLC/RF SYSTEM	77
6.1	Introduction	77
6.2	VLC frontend Design and Validation	77
6.3	mmW Measurement Campaign on Static Conditions	79
6.4	Measurement Results of the Hybrid VLC/RF on Mobility Conditions .	81
6.5	Switching Algorithm Deployment	85
6.6	Conclusion	89
VII	CONCLUSIONS AND FUTURE WORK	90
	REFERENCES	93
	VITA	105

LIST OF TABLES

1	Components and related specifications	15
2	Experimental set-up parameters for U-OFDM VLC System	15
3	Experimental set-up parameters for OOK VLC system.	22
4	Experimental set-up parameters of multihop vehicular VLC system. . .	35
5	Comparison between measured and analytical values	51
6	Comparison of the available vehicular channel models	58
7	Experimental set-up parameters for optical channel modelling.	59

LIST OF FIGURES

1	Block diagram of U-OFDM system under consideration.	9
2	Transmitted frame structure.	12
3	Block diagram of the experimental set-up.	14
4	(a) Effective channel per subcarrier and (b) SNR per subcarrier.	16
5	Measured BER values and comparison with theoretical results.	16
6	Block diagram of VLC system under consideration.	18
7	OOK frame structure.	20
8	Block diagram of the experimental set-up.	23
9	Experimental scenarios: (a) LOS (b) NLOS.	23
10	Constellation and eye diagram of received signal at 1 meter in LOS scenario with strict alignment (a) without lens (b) with lens.	24
11	Experimental measurements: (a) estimated SNR versus distance (b) measured BER versus SNR.	25
12	Receiver SNR versus reception angle for LOS scenario: (a) $\theta_{TX} = 0^\circ$ (b) $\theta_{TX} = 10^\circ$	25
13	Receiver SNR versus reception angle for NLOS scenario at $d_w = 0.5$ m: (a) $\theta_{TX} = 60^\circ$ (b) $\theta_{TX} = 90^\circ$	26
14	Selection of proper lens combination: Pictures taken from experiment.	27
15	Lens combinations.	28
16	Scenarios under consideration.	29
17	Comparison of optical and effective channel coefficients.	29
18	Effective channel coefficient for different lens combination at $\theta = 0^\circ$	30
19	Effective channel coefficient for different lens combination at $\theta = 10^\circ$	30
20	Vehicular multihop system.	33
21	Block diagram of of the experimental set-up of multihop vehicular VLC systems.	34

22	Measurements of the receive electrical SNR on 4th June 2020.	36
23	Measurements of the receive electrical SNR for individual hops: (a) daytime and (b) nighttime.	37
24	BER for individual hops (System with three hops is assumed): (a) day-time and (b) nighttime.	39
25	Overall multihop system's BER performances: (a) daytime and (b) nighttime.	40
26	Measurements of the receive electrical SNR with and without filter employment on 4th June 2020.	41
27	The experimental set-up.	44
28	Two tracks on the university campus where the measurements are taken.	47
29	Connection drop percentage for different deviation angle.	48
30	Received SNR at (a) Track 1 and (b) Track 2. Day time measurements with lens employment are assumed.	49
31	Histograms of measured Signal-to-Noise Ratio (SNR) at the output of the selection combiner. Day time measurements are assumed.	50
32	BER Comparison between numerical and closed-form expression with different orders.	53
33	BER Comparison between numerical and closed-form expression with different orders.	54
34	The measured illumination pattern of the utilized low-beam headlamp.	59
35	Comparison of the proposed path loss expression and other benchmarks.	61
36	Comparison of the proposed path loss expression for vehicle's low-beam headlamps with measured data ($D_R = 10$ mm and $d_h = 0$ m are assumed): (a) Photodetector is lower than headlights and (b) photodetector is higher than headlights.	62
37	Comparison of the proposed path loss expression for vehicle's low-beam headlamps with measured data for different receiver aperture diameter ($z = 0$ m and $d_h = 0$ m are assumed).	63
38	Comparison between the proposed path loss expression for vehicle's low-beam headlamps and the measured data for different lateral distances ($D_R = 10$ mm and $z = 0$ m are assumed).	64
39	(a) V2V system under consideration (b) experimental set-up.	65

40	Measured vertical displacement of Ford FMAX truck at: (a) 30 km/h, (b) 50 km/h, (c) 70 km/h and (d) 90 km/h.	66
41	Relation between the vertical oscillation variance and the truck speed. .	67
42	Numerical BER and closed-form for $\mu = 0$ m ($D_R = 10$ mm, $d_h = 0$ m and $v = 50$ km/h are assumed).	70
43	Effect of relative mean on error rate performance ($D_R = 10$ mm and $d_h = 0$ m are assumed) at: (a) $v = 30$ km/h and (b) $v = 50$ km/h.	72
44	Effect of lateral shift on error rate performance ($D_R = 10$ mm, $\mu = 0$ m and $v = 50$ km/h are assumed).	72
45	Effect of truck's speed on error rate performance ($D_R = 10$ mm, $d_h = 0$ m and $\mu = -0.1$ m are assumed).	73
46	BER versus photodetector's relative vertical location for: (a) Different inter-trucks distances ($D_R = 10$ mm, $d_h = 0$ m and $v = 50$ km/h are assumed) and (b) different truck speed values ($d = 10$ m and $D_R = 10$ mm and $d_h = 0$ m are assumed).	74
47	Error rate performance at the optimal vertical location on the receive vehicle of $\mu_{Opt} = -0.43$ m ($D_R = 10$ mm, $d_h = 0$ m and $v = 50$ km/h are assumed).	75
48	The new designed PCB for the vehicular headlight.	78
49	VLC system bandwidth before and after equalization.	78
50	Experimental environment: (a) Dens area and (b) open area.	79
51	Received signal at dens area for different intervehicle distances of (a) 0 m (b) 1 m and (c) 2 m.	80
52	Received signal at open area for different intervehicle distances of (a) 0 m (b) 1 m and (c) 2 m.	81
53	mmW link data rates at (a) Dens area and (b) open area.	82
54	Experimental set-up of hybrid system.	82
55	Under-consideration Roads.	83
56	Environment of the experimental campaign: (a) Side road and (b) main road.	83
57	Comparison between VLC and mmW systems on mobility conditions for: (a) Main roads and (b) side roads.	84

58	Flow chart of the hard switching Algorithm.	85
59	Flow chart of the Hysteresis Algorithm.	86
60	Flow chart of the proposed Algorithm.	87
61	Data rates of different Algorithms at: (a) Main road and (b) side road. .	88
62	Power efficiency of different switching Algorithms.	89



CHAPTER I

INTRODUCTION

1.1 Overview on Vehicular Connectivity

Intelligent transportation system (ITS) has emerged to confront many challenges such as smoothing traffic flow, ensuring road safety and reducing environmental impacts [1, 2]. Moreover, ITS is considered the main foundation of automotive future applications that mainly count on high throughput and low latency data transfer among vehicles. The deployment of ITS requires the existence of reliable vehicular connectivity which includes vehicle-to-vehicle (V2V) and vehicle-to-infrastructure (V2I) communications. These are commonly referred as vehicle-to-everything (V2X) communications [3]. Wireless Fidelity (WiFi) and cellular access solutions (e.g., DSRC, CV2X) have been typical choices for ITS deployments [4]. The existing research and standardization efforts on V2X have mainly focused on the use of radio frequency (RF) technologies. Such radio solutions are however sensitive to electromagnetic interference EMI. The impact of current V2X communications on the RF spectrum usages is negligible due to low levels of current market penetration. However, with the widespread adoption of ITS in the near future, limited RF bands licensed for ITS can quickly suffer from high interference levels, particularly in high-density traffic scenarios. Visible light communication (VLC), on the other hand, has emerged with the advantages of operating in an unlicensed band and the immunity to EMI [5]. The availability of light-emitting diode (LED) in vehicles exteriors paves the way for the LED to be used as wireless transmitters [6] as an alternative or complementary technology to RF-based V2X communications.

Vehicular VLC is in its infancy in many aspects and requires further research efforts in several research areas in order to achieve a robust vehicular VLC system. For example, most works on the propagation modelling and characterization of vehicular VLC

channels are mainly obtained through the simulation method which requires experimental validation. Even after the validation, the effect of the practical implementation, as well as the frontends, has to be considered. Moreover, for vehicular networking, special attention should be devoted to mobility as well as to the propagation channel. A key concept to enable connectivity in vehicular VLC is that of multihop transmission, where the signal transmitted from the source vehicle can reach the destination vehicle through a number of intermediate vehicles termed "relays". In addition, hybrid VLC/RF links need to be investigated to ensure link availability under all conditions. In light of the above open research topics, a robust VLC link could be achieved to support ITS deployment.

1.2 Literature Review and Motivation

Most of the available research in literature on vehicular VLC focuses only on indoor environment [7–29]. In these works, microcontroller [7–16], Arbitrary Waveform Generator (AWG) [17–23], Field Programmable Gate Arrays (FPGA) [24, 25] and Universal Software Radio Peripheral (USRP) [26–29] are used as platforms. All these works [7–29] are limited to the use of indoor-type LEDs and the effects of background environment are not considered. Some other attempts for investigating the performance of vehicular VLC were carried out in an outdoor environment in [30–43]. For example, in [42] Arduino-based vehicular VLC system was implemented at noon time on static conditions and it support data rate up to 9.5 Kbps using indoor-type LED. With on-off keying OOK modulation scheme and Manchester line coding, a 100 % Packet Delivery Ratio (PDR) up to 30 m distance was achieved . In [43], IEEE 802.15.7 Physical Layer (PHY) vehicular VLC system was implemented using USRP platform. At 200 KHz clock rate, OOK signal with 1/4 Forward Error Correction (FEC), and error free transmission with 11.67 bps at 14 m distance was achieved in shaded area at daytime on static conditions. Despite the availability of few initial attempts to investigate the performance in outdoor environment on static conditions, the critical effect for vehicular mobility needs to be further investigated.

Literature has very few efforts that investigate the critical concern of vehicular mobility [44–49]. For instance, in [44, 45], using the USRP platform and headlight transmitter, the received optical power was measured on a highway in an effort to model the channel characteristics. In [46], using the USRP N200 platform, vehicular taillight as a wireless transmitter and Thorlabs PDA100A as the optical receiver, the received packet ratio was measured on mobility conditions. Their system exhibited unstable behavior and faced an outage due to the ambient light of the road LED signage. Using the Arduino platform, a V2V VLC system was implemented and gave a 9.5 Kbps data rate using indoor-type LED with OOK modulation scheme and Manchester line coding at noontime on a mobile vehicle with a 100 % PDR up to 30 m [47]. In [48], a high reliability MCU-based V2V VLC prototype was designed. It showed a stable performance for 20 m with an indoor-type LED. In [49], a higher data rate of 3.1 Mbps was achieved with indoor-type LED and MATLAB offline processing. Direct Current Optical Orthogonal Frequency Division Multiplexing (DCO-OFDM) modulation was used with 52 data subcarrier and FEC 1/3. The system has an error less transmission for a 25 m distance of V2V implementation.

As the above survey highlights, most of the existing experimental works deploy indoor-type LEDs which exhibit different characteristics from vehicular headlights with asymmetrical intensity distributions. Furthermore, a comprehensive performance analysis of vehicular VLC systems under mobility conditions has not been yet presented. To fill these research gaps, we first investigate the baseband processing and the experimental implementation of different modulation techniques for VLC systems. We then focus on developing the frontend that supports high power transmission of vehicular LEDs and experimentally characterize impairments of the outdoor environment. A critical aspect of the vehicular VLC system is the channel model with its asymmetrical light sources, thus, we experimentally propose a pathloss model through a measurement campaign. Counting on these, we investigate the vehicular VLC link performance in the presence of mobility for passenger vehicles as well as for heavy trucks. Furthermore, we consider different road shapes (i.e., straight and curved roads). For benchmarking

purposes, we further implement a hybrid system (VLC/ RF) for comparison purposes as well as deploy a switching algorithm to satisfy a specific quality of service (QoS).

1.3 Dissertation Outline

The dissertation has five main parts outlined as follows:

In the first part of this thesis, we experimentally evaluate the performance of modulation techniques using software define platforms (i.e., USRP). We first illustrate the experimental set-up and detail the measurement procedures. We considered the implementation of an OOK modulation as well as OFDM. We further examine the effective channel that includes the effects of both the propagation channel and frontends. Furthermore, we measure the Bit Error Rate (BER) and validate the system results through a comparison with the theoretical values. In addition, we investigate the performance by considering different cases and conditions. Furthermore, we identify the proper lens combination to be used for real-time driving scenarios considering both increasing the communication distance and the angular limitations.

The second part illustrates the multihop vehicular VLC system for end-to-end connectivity. Firstly, we experimentally investigate the effect of the background noise variation on the vehicular VLC link. In addition, to extend the communication distance, we characterize the multihop vehicular VLC links at different times. We measure the BER for each individual hop and then obtain the overall performance of the multihop system.

The third part of the thesis focuses on investigating the effect of mobility. We employ multiple photodetectors on the receiving vehicle and identify each position/ orientation through a measurement campaign. We then consider different roads (straight/ curved) and the lens employment/ unemployment cases. We further consider daytime and nighttime measurements.

In the fourth part of the thesis, we consider a special case of truck-to-truck VLC communication. We first experimentally model and propose a 3D vehicular optical

channel. Then we model the truck's vertical oscillation through a measurement campaign. We then investigate different channel and system parameters on the link performance. Furthermore, in an effort to mitigate the effect of the vertical oscillation, we optimize the photodetector location.

In the fifth part, we compare with RF technology. For hybrid system deployment, both technologies must be able to connect with each other's, this is achieved through an Application Programming Interface (API) designed specifically for this purpose. We further propose a switching algorithm to maintain a specific QoS and benchmark with different algorithms in the literature.

1.4 Organization

The dissertation is organized as follows: Chapter 2 focuses on the baseband processing and the experimental implementation of different modulation techniques for VLC systems. It also considers the frontend development and the investigation of the outdoor environment. Chapter 3 focuses on extending the communication distance through multihop vehicular VLC experimental deployment. Chapter 4 focuses on investigating the vehicular VLC link in the presence of mobility. Chapter 5 investigates the mobility effect for a special case (i.e., Truck-to-Truck VLC system) of vertical oscillation. Toward this end, our work further considers a comparison with RF for benchmarking purposes. Therefore, Chapter 6 focuses on implementing a hybrid system (VLC/RF) and deploying a switching algorithm to satisfy a specific QoS. These chapters are organized as follows:

Chapter 2 has three sections. In Section 2.2, we experimentally evaluate the performance of Unipolar Orthogonal Frequency Division Multiplexing (U-OFDM) and further confirm our finding through benchmarking with the theoretical results. In Section 2.3, we experimentally evaluated a pulse modulation type (i.e., OOK). Furthermore, we investigated the performance in Line of Sight (LoS) and Non-Line of Sight (NLoS) conditions. We further tuned our system to have the best performance while

moving to the outdoor measurement campaigns. In Section 2.4, we identify the proper lens combination for vehicular VLC system.

Chapter 3 has four sections. In Section 3.2, we illustrate the experimental environment and the system model for the multihop vehicular VLC system. In Section 3.3, experimentally investigate the effect of the background noises throughout a full day measurement. In addition, to extend the communication distance, we experimentally characterize the vehicular multihop VLC system for individual hop in Section 3.4 as well as for overall hop in Section 3.5.

Chapter 4 has three sections. In Section 4.2, we illustrate the system model and the experimental environment of vehicular VLC on mobility conditions. In Section 4.3, we identify the proper orientation angle of the photodetectors on the receiving vehicle. In Section 4.4, we propose a statistical distribution for the channel coefficient as a result of the vehicular mobility. Moreover, we focus on the error rate analysis and the comparison with the experimental results.

Chapter 5 has four sections. In Section 5.2, experimentally model the vehicular optical channel in terms of the vertical distance. In Section 5.3, we model the truck's vertical oscillation based on measurements. We analyze the error rate and investigate the effect of the vertical oscillation on the system performance in Section 5.4. In Section 5.5, in an attempt to mitigate the degrading effect of the vertical oscillation, we identify the optimal location of the photodetector.

Chapter 6 has four sections. In Section 6.2, we design and validate a new frontend for supporting a higher data rate of the VLC system. In Section 6.3, we conduct a measurement campaign for V2V millimeter Wave (mmW) connectivity. In Section 6.4, we illustrate the experimental set-up for comparing VLC and RF technologies in vehicular domain (i.e., V2V). In Section 6.5, we deploy a proposed switching Algorithm and further compare with different Algorithms for benchmarking purposes.

CHAPTER II

SIGNALLING MODELS FOR VLC SYSTEMS

2.1 Introduction

VLC is based on modulating the light intensity of LEDs. In a VLC system, the information is superimposed upon the DC driving the LED. The DC value is selected according to the desired operating point taking into account the amplitude constraints. Since the frequency of the modulating signal is very high, the flickering of the light intensity cannot be observed by the human eye. Typical modulation choice for VLC systems is OOK and pulse modulation techniques while other modulation techniques such as OFDM have been further proposed to support ultra-high-speeds on the order of Gbps.

In this chapter, we experimentally evaluate the performance of different modulation techniques using software define platforms. The experimental set-up was built upon a pair of modified USRPs where RF daughterboards are replaced by baseband LFTX and LFRX cards and optical frontends. Through a measurement campaign, for OFDM-based VLC system, we present the frequency response of the effective channel including the effects of both propagation channel and frontends. Then, we measure the BER per subcarrier and obtain the overall BER. Finally, we compare the experimental results with theoretical values to confirm our work. While for OOK-based VLC system, we first validate the system results through a comparison with the theoretical values. In addition, we investigate the performance considering different transmitter and receiver orientation angles. Furthermore, we consider both LoS and NLoS scenarios with and without lens employment. We further investigate the advantages and disadvantages of lens employment on the received signal. Then we check the angular limitation of the lenses. Finally, throughout a measurement campaign, we identify the optimal lens

combination to be used for real-time driving scenarios considering both increasing the communication distance and angular limitations.

The rest of the chapter is organized as follows. Section 2.2 presents a detailed OFDM system implementation. First, we present the system model, detail essential experimental aspects and discuss the experimental results. In Section 2.3, we present an OOK-based VLC system. We further discuss the results of the under-consideration LoS and NLoS configurations. Section 2.4 investigate the effect of different lens combination on the vehicular VLC system performance.

2.2 OFDM-Based VLC System

To handle the frequency-selectivity of VLC channels, particularly at high speeds, more recent works are built upon various forms OFDM modified to obtain a real-valued and unipolar signal suitable for driving the light source [50]. Hermitian symmetry is typically used in optical OFDM schemes to make the signal real valued at the transmitter side. To obtain a unipolar signal, various techniques are used. In DCO-OFDM [51], a DC bias is simply added to shift the bipolar signal into the dynamic range of LED. Asymmetrically-Clipped Optical Orthogonal Frequency Division Multiplexing (ACO-OFDM) [52], only odd subcarriers are modulated; this ensures a symmetry in the time domain and makes it possible to clip the non-negative values without any loss of information. Another alternative is U-OFDM [53], also known as Flip-OFDM [54], where each time sample is encoded into a pair of new time samples. For instance, if the sample is positively valued, it is encoded to positive of the magnitude followed by zero. On the other hand, if the sample is negatively valued, it is encoded as a zero followed by the positive magnitude value of the negative sample. By grouping the first samples of each pair, we can obtain the so-called "positive OFDM sub-frame". Similarly, by grouping the second samples of each pair, we can obtain the "negative OFDM sub-frame". The U-OFDM signal is obtained by concatenating positive and negative sub-frames. On the receiver side, the decoding process is simply achieved by subtracting the negative sub-frame from the positive one. It was reported in [53] that U-OFDM

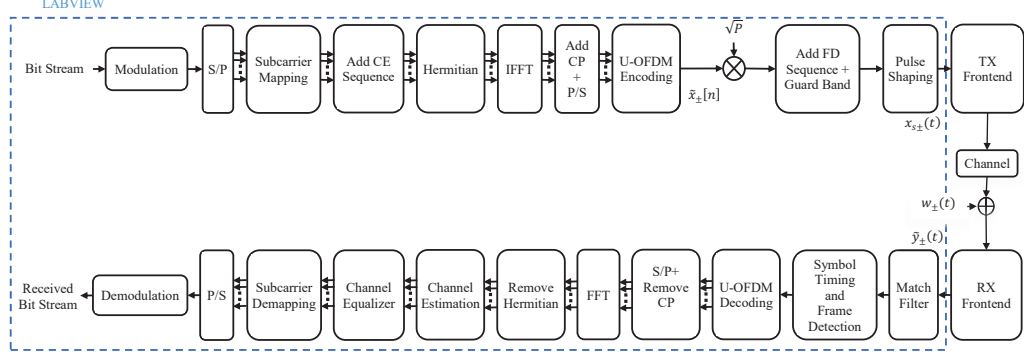


Figure 1: Block diagram of U-OFDM system under consideration.

is more power-efficient than DCO-OFDM and gives better error rate performance than both DCO-OFDM and ACO-OFDM.

Some experimental evaluations of U-OFDM were also reported in [55–57] while utilizing software defined platforms, as a flexible alternative where physical layer functionalities are implemented through modifiable software or firmware, could be found in [58–61]. To the best of our knowledge, implementation of U-OFDM on a software defined platform has not been previously reported.

The block diagram of U-OFDM is illustrated in Figure 1. Let $S_i, i = 1, \dots, N/2 - 1$ denote the M -Quadrature Amplitude Modulation (QAM) symbols where M is the modulation size and N is the number of sub-carriers. In order to ensure that the output of Inverse Fast Fourier Transform (IFFT) is real, Hermitian symmetry is imposed. The resulting signal is given by

$$\mathbf{X} = \begin{bmatrix} 0 & S_1 & \cdots & S_{N/2-1} & 0 & S_{N/2-1}^* & \cdots & S_1^* \end{bmatrix}^T \quad (2.1)$$

Let $\mathbf{X}[k]$ denote the k^{th} element of $\mathbf{X}$. After IFFT, the time domain signal is expressed as $\mathbf{x} = \begin{bmatrix} \mathbf{x}[0] & \mathbf{x}[1] & \mathbf{x}[2] & \cdots & \mathbf{x}[N-1] \end{bmatrix}^T$ whose n^{th} element can be expressed as

$$\mathbf{x}[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} \mathbf{X}[k] \exp(j2\pi nk/N), \quad n=0, 1, \dots, N-1 \quad (2.2)$$

A Cyclic Prefix (CP) with a length of N_{cp} is appended to $\mathbf{x}$ to prevent interference between OFDM symbols and to ensure the circular convolution with the channel. This makes possible the use of a single tap equalizer in frequency domain. After the addition

of CP, we have

$$\tilde{\mathbf{x}} = \begin{bmatrix} \mathbf{x}[N-N_{cp}] & \cdots & \mathbf{x}[N-1] & \mathbf{x}[0] & \cdots & \mathbf{x}[N-1] \end{bmatrix}^T \quad (2.3)$$

Due to Hermitian symmetry, $\tilde{\mathbf{x}}$ is real valued, but still bipolar. Based on U-OFDM structure, each time sample is encoded into a pair of samples. If $\tilde{\mathbf{x}}[n] \geq 0$, it will be encoded as $(\tilde{\mathbf{x}}[n], 0)$. On the other hand, if $\tilde{\mathbf{x}}[n] < 0$, it will be encoded as $(0, -\tilde{\mathbf{x}}[n])$. First and second samples of each pair are assigned, respectively, to the positive and negative sub-frames which are denoted as $\tilde{\mathbf{x}}_+$ and $\tilde{\mathbf{x}}_-$.

Let $p_T(t)$, P , T_s and $\delta(\cdot)$ denote transmit pulse shaping filter, average electrical transmit power, pulse duration and Dirac delta function, respectively. Furthermore, we introduce $\tilde{\mathbf{x}}_{\pm}[n]$ to describe either $\tilde{\mathbf{x}}_+[n]$ or $\tilde{\mathbf{x}}_-[n]$ for a unified notation. The continuous time domain OFDM waveform can be expressed as

$$x_{s\pm}(t) = A \left[\sqrt{P} \sum_{n=0}^{N+N_{cp}-1} \tilde{\mathbf{x}}_{\pm}[n] \delta(t - nT_s) \right] \otimes p_T(t) + V_{DC} \quad (2.4)$$

where A is an amplification factor to mitigate the effects of path loss. Since it is likely that the first term in (2.4) can be still lower than LED turn on voltage, a DC bias (V_{DC}) was further added. The signal goes through the optical channel. At the receiver side, a photodetector captures the light and converts into an electric current. The received electrical signal at the destination can be written as

$$\tilde{y}_{\pm}(t) = [R x_{s\pm}(t) \otimes l(t) \otimes h(t) + w_{\pm}(t)] \otimes p_R(t) \quad (2.5)$$

where R is the photodetector responsivity, $l(t)$ is the LED impulse response, $h(t)$ is the propagation channel impulse response, $p_R(t)$ is the receive pulse shaping filter, and $w_{\pm}(t)$ is the noise term. Replacing (2.4) in (2.5), we have

$$\tilde{y}_{\pm}(t) = \sqrt{P} \left[\sum_{n=0}^{N+N_{cp}-1} \tilde{\mathbf{x}}_{\pm}[n] \delta(t - nT_s) \right] \otimes h_{eff}(t) + V_{RDC}(t) + w_{g\pm}(t) \quad (2.6)$$

In (2.6), we define $h_{eff}(t) = A R p_T(t) \otimes l(t) \otimes h(t) \otimes p_R(t)$ as the “effective channel impulse response” incorporating the effects of propagation channel, LED response, and pulse shaping. In (2.6), we have further defined $w_{g\pm}(t) = w_{\pm}(t) \otimes p_R(t)$ and $V_{RDC}(t) = R V_{DC} \otimes l(t) \otimes h(t) \otimes p_R(t)$.

Let $w_{g\pm}(n)$, $\tilde{y}_{\pm}(n)$, $V_{R_{DC}}(n)$ and $h_{eff}(n)$ denote, respectively, the sampled versions of $w_{g\pm}(t)$, $\tilde{y}_{\pm}(t)$, $V_{R_{DC}}(t)$ and $h_{eff}(t)$. Further, noting that $V_{R_{DC}}(n)$ is the same for all samples, index n can be dropped and $V_{R_{DC}}(n)$ can be written as $V_{R_{DC}}$. Therefore, the sampled version of $\tilde{y}_{\pm}(t)$ can be expressed as

$$\tilde{y}_{\pm}(n) = \sqrt{P}\tilde{x}_{\pm}(n) \otimes h_{eff}(n) + V_{R_{DC}} + w_{g\pm}(n), \quad n = 0, 1, \dots, N + N_{CP} - 1 \quad (2.7)$$

After removing CP from $\tilde{y}_{\pm}(n)$, we have

$$y_{\pm}(n) = \sqrt{P}x_{\pm}(n) \otimes h_{eff}(n) + V_{R_{DC}} + w_{g\pm}(n), \quad n = 0, 1, \dots, N - 1 \quad (2.8)$$

From the descriptions of positive and negative U-OFDM sub-frames, it can be readily verified that $x(n) = x_+(n) - x_-(n)$. Therefore, we have

$$y(n) = y_+(n) - y_-(n) = \sqrt{P}x(n) \otimes h_{eff}(n) + \bar{w}_g(n), \quad n = 0, 1, \dots, N - 1$$

In (2.8), $\bar{w}_g(n) = w_{g+}(n) - w_{g-}(n)$ is the noise term. After Fast Fourier Transform (FFT), the frequency domain signal at the k^{th} subcarrier can be expressed as

$$Y(k) = \sqrt{P}H_{eff}(k)X(k) + \bar{W}_g(k), \quad k = 0, 1, \dots, N - 1 \quad (2.9)$$

where $H_{eff}(k)$ and $\bar{W}_g(k)$ are, respectively, the k^{th} element of N -point FFT of $h_{eff}(n)$ and $\bar{w}_g(n)$.

For detection, we first equalize the received signal by using a single-tap equalizer. Mathematically speaking, we first obtain

$$\widehat{X}(k) = \left(\sqrt{P}\widehat{H_{eff}}(k) \right)^{-1} Y(k), \quad k = 0, 1, \dots, N - 1 \quad (2.10)$$

where $\widehat{H_{eff}}(k)$ is the estimated effective channel coefficient for the k^{th} carrier. Then, we perform Maximum Likelihood (ML) decoding, i.e.,

$$\hat{X}(k) = \arg \min_{X(k) \in \Omega} \left\| \widehat{X}(k) - X(k) \right\| \quad (2.11)$$

where $\|\cdot\|$ is the L_2 norm and Ω is the set of M -QAM symbols.

The signal frame structure illustrated in Figure 2. The start of each frame includes a guard band with a length of L_G and a frame detection training sequence with

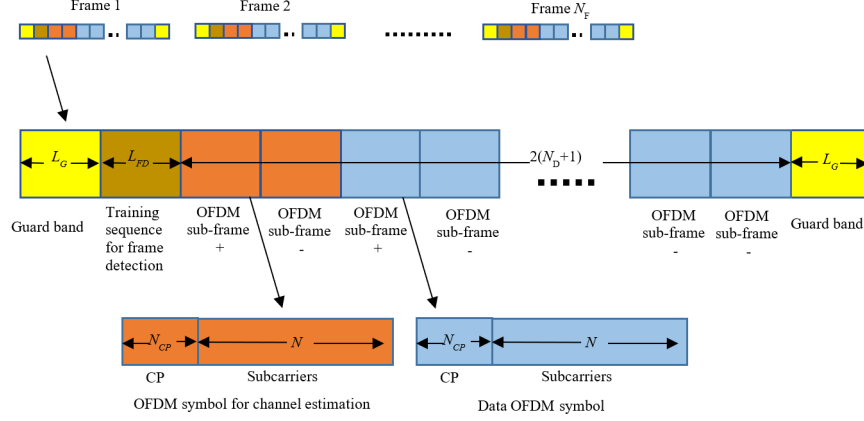


Figure 2: Transmitted frame structure.

a length of L_{FD} . This is then followed by the transmission of $2(N_D + 1)$ OFDM sub-frames, i.e., positive and negative OFDM sub-frames following each other. Each OFDM sub-frame has a cyclic prefix with a length of N_{CP} and data carriers with a length of N . The first two OFDM sub-frames carry the channel estimation training sequence while the other carry data¹. Finally, a guard with a length of L_G is inserted before the transmission of next frame. Therefore, a frame has a total length of $N_{\text{down}} = 2L_G + L_{FD} + 2(N + N_{CP})(N_D + 1)$. We use Root-Raised-Cosine (RRC) filter with a roll-off factor of β for transmitter pulse shaping filter. After pulse shaping, the total number of samples in a transmitted frame is given by $N_{\text{up}} = N_{\text{down}}U$ where U denotes the upsampling factor. In the following, we explain how we perform the symbol timing estimation, frame detection, and channel (and subsequent SNR) estimation.

First, we perform symbol timing which refers to the estimation of the sampling point of the symbol. We consider the maximum output energy method [62, Chapter. 5] which finds the sample point that maximizes the average received energy. Let $Y_{up}(n)$, $n = 1, 2, \dots, N_{\text{up}}$ denote the sampled received frame. Further let $\tau = 1, 2, \dots, U$ denote the possible delay values. Then, we can write

$$\hat{\tau} = \max_{\tau} \left| E \left[(Y_{up}(nU + \tau))^2 \right] \right|, n = 1, 2, \dots, N_{\text{down}} \quad (2.12)$$

¹Due to Hermitian symmetry and unipolar coding, only $N/2 - 1$ of $2N$ carriers actually carry data while the others are loaded with zero.

After symbol timing, frame detection is performed in order to resolve multiple symbol period delays. In other words, we determine the beginning of the frame. For frame detection, we use the training sequence $T_{\text{FD}}(d)$, $d = 0, \dots, L_{\text{FD}} - 1$. By correlating the received signal with the known training sequence for frame detection that is placed at the header of the transmitted frame, the beginning of frame is found at the position of the autocorrelation's peak. Mathematically speaking, the correlator output can be written as

$$\hat{d} = \max_n \left| \sum_{d=0}^{L_{\text{FD}}-1} T_{\text{FD}}(d) Y_{\text{down}}(n+d) \right| \quad (2.13)$$

where $Y_{\text{down}}(n) = Y_{\text{up}}(nU + \hat{\tau})$, $n = 1, 2, \dots, N_{\text{down}}$ is the down-sampled received frame.

For channel estimation, we use the training sequence $T_{\text{CE}}(i)$, $i = 1, \dots, N/2 - 1$ with unit power. Based on (2.9), the corresponding received frequency domain signal for the k^{th} subcarrier can be written as

$$Y(k) = H_{\text{eff}}(k) T(k) + \bar{W}_g(k), \quad k = 0, 2, \dots, N-1 \quad (2.14)$$

Using zero-forcing estimator, the estimated channel gain at k^{th} subcarrier, $k = 0, 1, \dots, N-1$, can be written as

$$\widehat{H_{\text{eff}}}(k) = Y(k) T^{-1}(k) \quad (2.15)$$

It should be noted that the estimated effective channel gain will include the effects of channel and transmitter/receiver frontends, i.e., amplification gain, LED conversion loss, photodetector responsivity, and pulse shaping filters.

To measure the noise, we switch off the LED and measure the received signals. Assuming that L samples are collected, the estimated noise power for the k^{th} subcarrier can be written as

$$\sigma_k^2 = \frac{1}{L} \sum_{l=1}^L |Y_l(k)|^2 \quad (2.16)$$

Since $\sigma_{n_k}^2$ remains the same for all subcarriers, we can drop index k and write it as σ_n^2 .

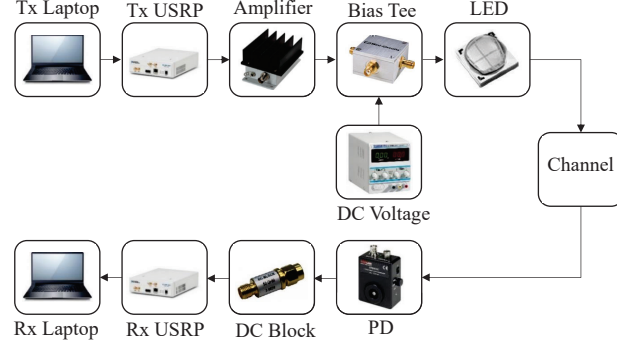


Figure 3: Block diagram of the experimental set-up.

Therefore, SNR at the k^{th} subcarrier can be written as

$$\widehat{SNR}[k] = \frac{\widehat{H_{eff}(k)}^2}{\sigma_n^2} \quad (2.17)$$

Figure 3 illustrates the block diagram of our experimental setup. At the transmitter, LabVIEW software of the transmitter (source laptop) generates U-OFDM symbols. The generated data is transferred from the laptop via Ethernet cable to the modified NI 2920 USRP with Low Frequency Transmitter (LFTX) daughter boards for Digital to Analog Conversion (DAC). The output signal of the USRP is amplified by a power amplifier (Mini-circuits ZHL-32A+) to adjust the signal power for the input of Bias-Tee (Mini-circuits ZFBT-GW+). DC voltage is then added to shift the signal to the linear operation region of the LED. We use an LED (Lumileds LXA7-PW50) with turn-on voltage of 2.5 V and maximum forward voltage of 3.25 V. At the receiver, we use an Avalanche Photodiode (APD) (Thorlabs APD 130 A/M). After the optical signal is converted into an electrical signal, DC component of the received signal is removed. This is then fed to the modified NI 2920 USRP with Low Frequency Receiver (LFRX) daughter board. U-OFDM demodulation is performed on LabVIEW software running on the destination laptop.

The component types and specifications are summarized in Table 1. The transmitter and receiver are placed apart with a link distance of $d = 1.2\text{m}$. Average electrical transmit power (P) is varied in the range of 0.1 mW and 3.6 mW. 4-QAM is employed. The USRP sampling rate is $f_s = 5 \times 10^6$ samples per second. We use an upsampling factor of $U = 80$. We use RRC pulse shape with a roll-off factor of $\beta = 0.5$.

Table 1: Components and related specifications

Component	Brand and Model	Specifications
Amplifier	Mini-circuits ZHL-32A+	0.05 to 130 MHz (Min 25 dB gain)
NI USRP 2920	Ettus LFTX and LFRX boards	0 to 30 MHz
Bias-Tee	Mini-circuits ZFBT-GW+	0.1 to 6000 MHz
Photodetector	Thorlabs APD 130 A/M	400 to 1000 nm
LED	Lumileds LXA7-PW50	High power white LED

Table 2: Experimental set-up parameters for U-OFDM VLC System

Link distance	d	1.2 m
Electrical Noise power	σ_n^2	-61.37 dB
Average electrical transmitted power	P	0.1 mW - 3.6 mW
Modulation order	M	4
Number of data subcarriers	N	32
Length of cyclic prefix	N_{cp}	4
Training sequence length for frame detection	L_{FD}	80
Training sequence length for channel estimation	L_{CE}	15
Guard band length	L_G	6
Number of data OFDM symbols	N_D	368
Number of frames	N_F	60
Upsampling factor	U	80
USRP sampling rate	f_s	5×10^6 samples/sec
RRC pulse shaping filter length	L_{PS}	80
Roll-off factor	β	0.5

The pulse shape is truncated to the length of $L_{PS} = 80$ symbols. Each frame includes a guard band with a length of $L_G = 6$, a frame detection training sequence with a length of $L_{FD} = 80$, channel estimation training sequence with a length of $L_{CE} = 15$, cyclic prefix with a length of $N_{cp} = 4$ and data carriers with a length of $N = 32$. For the convenience of the reader, all parameters are summarized in Table 2.

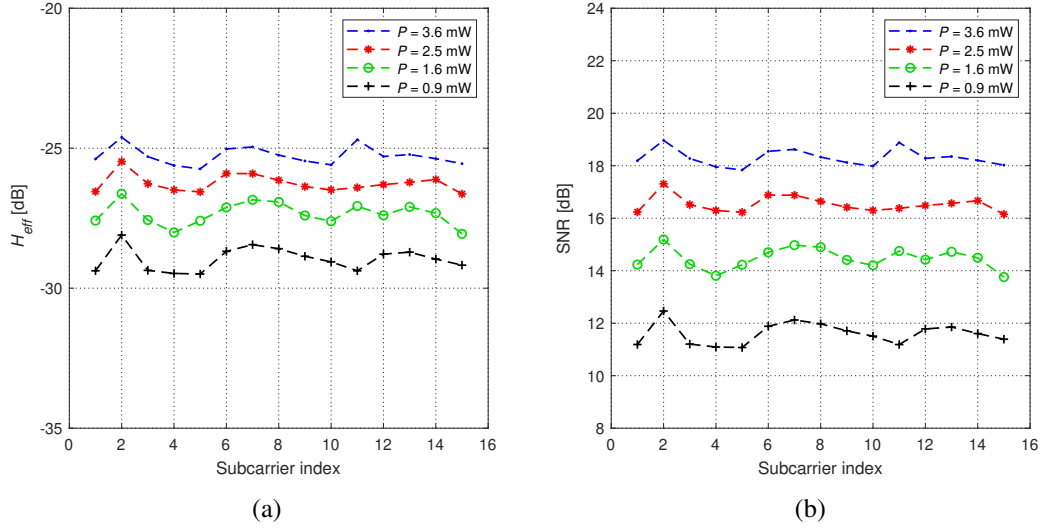


Figure 4: (a) Effective channel per subcarrier and (b) SNR per subcarrier.

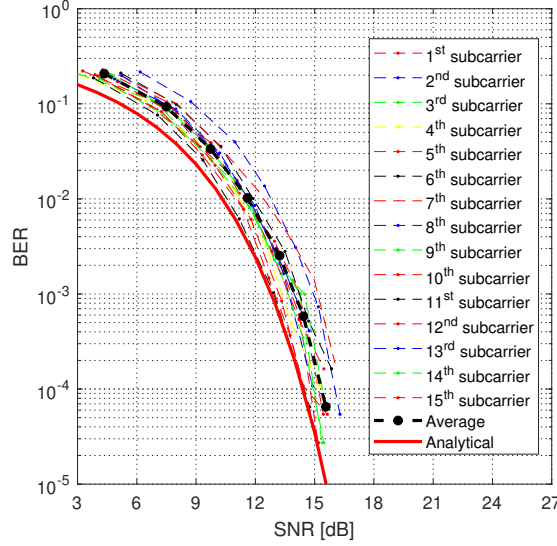


Figure 5: Measured BER values and comparison with theoretical results.

In here, we measured the effective channel gain per each subcarrier $H_{eff}(k)$ and illustrated in in Figure 4.a for different values of average electrical transmit power. It can be readily verified that the channel exhibits frequency-flat characteristics with very little fluctuation in frequency domain. This is rather expected since the transmitter and receiver have a line-of-sight configuration with a good alignment. We further measured noise power based on $L = 100$ noise samples for each subcarrier and, based on (2.16), we determined it as $\sigma_n^2 \cong -61.37$ dB for all subcarriers. Then, based on (2.17), we obtained the received SNR per each subcarrier and illustrated in Figure 4.b

For BER measurements, we transmitted $N_F = 60$ frames. Each frame includes $N_D + 1 = 369$ OFDM sub-frames, one of which is dedicated for channel estimation. Therefore, the total number of transmitted bits is given by $N_F N_D (N/2 - 1) \log_2 M = 60 \cdot 368 \cdot 15 \cdot 2 = 6.624 \times 10^5$ bits. At the receiver side, these were demodulated based on the decision metric in (2.11) ref and compared with actual transmitted bits. The measured BER for each subcarrier is illustrated in Figure 5. Averaging over $N/2 - 1 = 15$ subcarriers, we obtain the measured average BER and plot it as well. As a benchmark, we further include the analytical BER expression. For square QAM constellations, an approximate expression of BER per subcarrier is given by [63]

$$BER[k] \cong \frac{2(\sqrt{M} - 1)}{\sqrt{M} \log_2 \sqrt{M}} Q \left(\sqrt{\frac{3 \text{SNR}[k]}{2(M - 1)}} \right) \quad (2.18)$$

where $Q(\cdot)$ is the Gaussian Q-function. Averaging (2.18) over all data carriers, the average BER can be obtained by

$$BER = \frac{1}{N/2 - 1} \sum_{k=1}^{N/2-1} BER[k] \quad (2.19)$$

It is observed that the characteristics of theoretical and measured BER expressions match each other. There is around 2 dB shift between theoretical and experimental BER curves. This shift decreases as SNR increases. At low SNR values, frame synchronization and symbol timing experience higher estimation errors. At high SNR, the difference reduces to around 1.6 dB because of the better estimation quality.

2.3 OOK-Based VLC System

The typical modulation choice for VLC systems is OOK and pulse modulation techniques. For Vehicular VLC systems, ITS messages are transferred among vehicles with low data rate requirement. Therefore, OOK system could be justified for such vehicular single tap channel for ITS messages.

There is already a rich literature on theoretical aspects of VLC, see e.g., a recent survey [64] and references therein, covering various aspects of channel modelling, physical layer design, and networking. The performance of VLC systems was also evaluated by

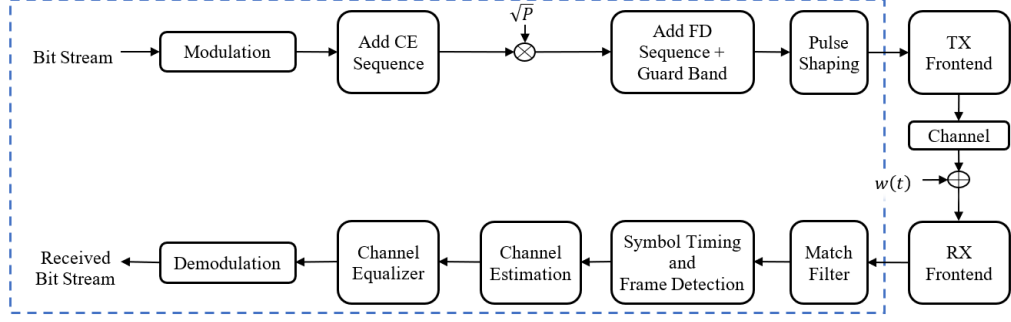


Figure 6: Block diagram of VLC system under consideration.

experimental means in [65]. While most of these studies assumed LoS scenarios, the performance of VLC systems was further demonstrated in NLoS conditions [66–68].

In addition to above experiments conducted mainly with laboratory equipment or dedicated hardware designs on FPGA or microcontrollers, software defined platforms, as a more flexible prototyping tool, were used in VLC experiments in [69]. In here, we present a comprehensive experimental evaluation of a software-defined VLC system with OOK considering both LoS and NLoS scenarios. The modified NI USRPs are coupled with a custom design frontend recently made available by Hyperion Technologies [70]. The receiver has a wide field-of-view of 170 degrees and supports both LoS and NLoS connectivity. In the measurement campaign, we make measurements for SNR and BER. We further investigate the effect of using lens and receiver/transmitter orientation on system performance.

As illustrated in Figure 6, we consider a VLC system with OOK. Let $S(i)$, $i = 1, \dots, L_D$ denote the i^{th} transmitted OOK symbol. The continuous-time waveform can be expressed as

$$x(t) = \left(\sqrt{P_t} \sum_{i=1}^{L_D} S(i) \delta(t - iT_s) \right) \otimes p_T(t), \quad (2.20)$$

where $p_T(t)$, T_s and $\delta(\cdot)$ denote, respectively, transmit pulse shaping filter, pulse duration and Dirac delta function. Here, P_t denotes the transmit power. It is set as $P_t = 0$ if bit ‘0’ is transmitted, otherwise set as $P_t = 2P$. Therefore, the average electrical transmit power becomes P . After adding a DC bias (V_{DC}), the waveform that drives the LED can

be written as

$$x_{\text{LED}}(t) = A \left(\sqrt{P_t} \sum_{i=1}^{L_D} S(i) \delta(t - iT_s) \right) \otimes p_T(t) + V_{DC}, \quad (2.21)$$

where A is an amplification factor. The signal goes through the optical channel. At the receiver side, a photodetector captures the light and converts into an electric current. The received electrical signal at the destination can be written as

$$x_{\text{PD}}(t) = [R x_{\text{LED}}(t) \otimes l(t) \otimes h(t) + w(t)] \otimes p_R(t), \quad (2.22)$$

where R is the photodetector responsivity, $l(t)$ is the LED impulse response, $h(t)$ is the optical channel impulse response, $p_R(t)$ is the receive pulse-shaping filter, and $w(t)$ is the additive white Gaussian noise term with mean μ_n and noise variance of σ_n^2 . Replacing (2.21) in (2.22), we have

$$x_{\text{PD}}(t) = \left(\sqrt{P_t} \sum_{i=1}^{L_D} S(i) \delta(t - iT_s) \right) \otimes h_{\text{eff}}(t) + V_{R_{DC}} + w_g(t), \quad (2.23)$$

where $w_g(t) = w(t) \otimes p_R(t)$ and $V_{R_{DC}} = R V_{DC} \otimes l(t) \otimes h(t) \otimes p_R(t)$. It can be readily checked that the statistics of $w_g(t)$ remain the same as those of $w(t)$. In (2.23), we define $h_{\text{eff}}(t) = A R p_T(t) \otimes l(t) \otimes h(t) \otimes p_R(t)$ as the “effective channel impulse response” incorporating the effects of propagation channel, frontends, and pulse shaping. After sampling the received signal, the i^{th} sample can be expressed as

$$x_{\text{PD}}(i) = \sqrt{P_t} S(i) h_{\text{eff}}(i) + V_{R_{DC}} + w_g(i), \quad i = 1, \dots, L_D \quad (2.24)$$

where $w_g(i)$, $x_{\text{PD}}(i)$ and $h_{\text{eff}}(i)$ denote, respectively, the i^{th} sample of $w_g(t)$, $x_{\text{PD}}(t)$ and $h_{\text{eff}}(t)$. For detection, we first remove the effect of DC bias. Mathematically speaking, we have

$$\widehat{x_{\text{PD}}}(i) \approx \sqrt{P_t} h_{\text{eff}} S(i) + \left(V_{R_{DC}} - \widehat{V_{R_{DC}}} \right) + w_g(i). \quad (2.25)$$

where $\widehat{V_{R_{DC}}}$ is the estimated value of DC-bias. Using a single-tap equalizer, we can

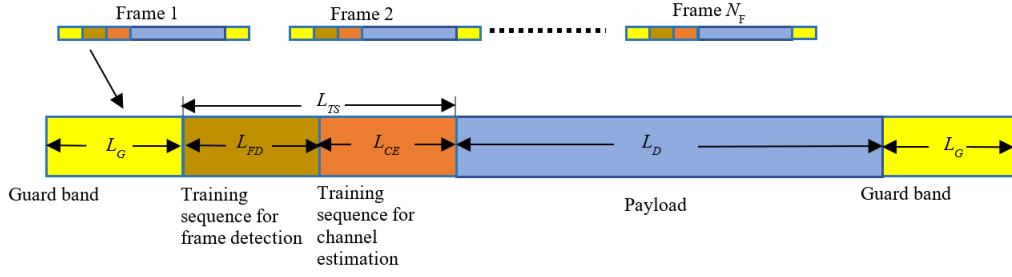


Figure 7: OOK frame structure.

retrieve the message signal as

$$\begin{aligned}\widehat{S}(i) &\approx \left(\sqrt{P_t}\widehat{h_{eff}}\right)^{-1} \widehat{x_{PD}}(i) \\ &= \frac{h_{eff}}{\widehat{h_{eff}}} S(i) + \widehat{w}(i),\end{aligned}\tag{2.26}$$

where $\widehat{w}(i) = \left(\left(V_{R_{DC}} - \widehat{V_{R_{DC}}}\right) + w(i)\right) / \sqrt{P_t}\widehat{h_{eff}}$ and $\widehat{h_{eff}}$ is the estimate of the channel coefficient.

Here, we discuss frame detection, symbol timing, DC bias estimation and channel estimation which are essential for practical implementation. The frame structure is illustrated in Figure 7. Each frame includes a guard band of length L_G followed by a training sequence with a length of $L_{TS} = L_{FD} + L_{CE}$. The training sequence is used for three purposes: i) Frame detection, ii) DC bias estimation and iii) channel estimation. This is then followed by the transmission of L_D data symbols. Finally, another guard band with a length of L_G is inserted before the transmission of the next frame. Therefore, a frame has a total length of $L_{down} = 2L_G + L_{TS} + L_D$. RRC with a roll-off factor of β is used as a transmitter pulse shaping filter. After pulse shaping, the total number of samples in a transmitted frame is given by $L_{up} = L_{down}U$ where U denotes the up-sampling factor.

First, we perform symbol timing to estimate the sampling point of the symbol. For this purpose, we use the maximum output energy method [62]. This finds the sample point where the average received energy is maximized [62, Chapter 5]. Mathematically speaking, this can be obtained as

$$\hat{\tau} = \max_{\tau} \left| E \left[(Y_{up}(iU + \tau))^2 \right] \right|, \quad i = 1, 2, \dots, L_{down}\tag{2.27}$$

where $Y_{\text{up}}(i), i = 1, 2, \dots, L_{\text{up}}$ denote the received samples and $\tau = 1, 2, \dots, U$ denote the possible delay values.

After symbol timing, frame detection is performed in order to resolve multiple symbol period delays [71]. The training sequence for frame detection consists of ones and zeros. Based on (2.25), the corresponding received samples can be expressed as

$$x_{\text{zeros}}(i) = V_{R_{DC}} + w_g(i), \quad i = 1, \dots, L_{\text{zeros}} \quad (2.28)$$

$$\begin{aligned} x_{\text{ones}}(i) &= \sqrt{P_t} h_{\text{eff}} + V_{R_{DC}} \\ &+ w_g(i), \quad i = 1, \dots, L_{\text{ones}} \end{aligned} \quad (2.29)$$

where L_{zeros} and $L_{\text{ones}} = L_{FD} - L_{\text{zeros}}$, respectively, denote the numbers of zeros and ones used for training. For DC-bias estimation, we calculate the average of (2.28) over available samples which yields

$$\widehat{V_{R_{DC}}} = \frac{1}{L_{\text{zeros}}} \sum_{i=1}^{L_{\text{zeros}}} x_{\text{zeros}}(i). \quad (2.30)$$

After removing the estimated DC-bias, we can then express (2.28) and (2.29) as

$$x_{\text{zeros-DC}}(i) = w_{g\text{-DC}}(i), \quad i = 1, \dots, L_{\text{zeros}} \quad (2.31)$$

$$\begin{aligned} x_{\text{ones-DC}}(i) &= \sqrt{P_t} h_{\text{eff}} \\ &+ w_{g\text{-DC}}(i), \quad i = 1, \dots, L_{\text{ones}} \end{aligned} \quad (2.32)$$

where $w_{g\text{-DC}}(i) = (V_{R_{DC}} - \widehat{V_{R_{DC}}}) + w_g(i)$ denotes the effective noise. It can be readily verified that, theoretically, the effective noise is still zero-mean Gaussian with a variance of σ_n^2 . Utilizing (2.31), we can estimate the noise power of the effective noise as

$$\widehat{\sigma_n^2} = \frac{1}{L_{\text{zeros}}} \sum_{i=1}^{L_{\text{zeros}}} |x_{\text{zeros-DC}}(i)|^2. \quad (2.33)$$

On the other hand, using (2.32), we can estimate the effective channel coefficient as

$$\widehat{h_{\text{eff}}} = \frac{1}{\sqrt{P_t} L_{\text{ones}}} \sum_{i=1}^{L_{\text{ones}}} x_{\text{ones-DC}}(i). \quad (2.34)$$

Table 3: Experimental set-up parameters for OOK VLC system.

LOS Link distance	d	[1, 1.5, ..., 3] m
Distance between reflector and Tx-Rx frontends	d_w	0.5 m
Electrical Noise power	σ_n^2	- 48.27 dB
Average electrical transmitted power	P	0 dB
Training sequence length for frame detection	L_{FD}	80
Training sequence length for channel estimation	L_{CE}	80
Guard band length	L_G	6
Number of frames	N_F	200
Number of data symbols per frame	L_D	8000 bit/ frame
Up-sampling factor	U	20
USRP sampling rate	f_s	7×10^6 Sample/sec
RRC pulse shaping filter length	L_{PS}	40
Roll-off factor	β	0.5

Based on (2.33) and (2.34), estimated SNR can be written as

$$\widehat{SNR}[k] = \frac{\widehat{P h_{eff}}^2}{\widehat{\sigma_n^2}}. \quad (2.35)$$

Figure 8 illustrates the block diagram of our experimental set-up. At the transmitter side, the laptop with LabVIEW software generates the digital signal which is transferred to the modified NI 2920 USRP with LFTX daughterboard through an Ethernet cable [72]. After DAC, the signal is delivered to custom-design VLC frontend [73]. Tx frontend is equipped with a Lambertian pattern LED with 60 degrees half angle. Tx frontend has 20 MHz and interchangeable lens mechanism. The optical signal propagates through the channel and captured by the receiver frontend. The Rx frontend has a wide Field of View (FOV) of 170 degree which allows NLoS operation [73]. Both frontends are compatible with USRPs. This is followed by another modified NI 2920 USRP with LFRX daughterboard. The output of USRP is a digital signal which is fed to the receiver laptop for demodulation. All parameters are listed in Table 3

In our measurement campaign, we consider both LoS and NLoS scenarios as illustrated in Figure 9. The length of the straight line between transmitter and receiver is

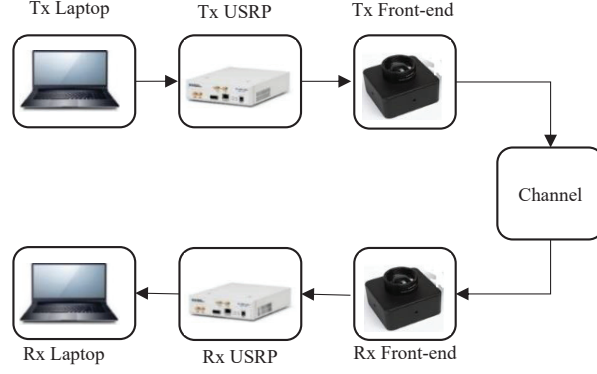


Figure 8: Block diagram of the experimental set-up.

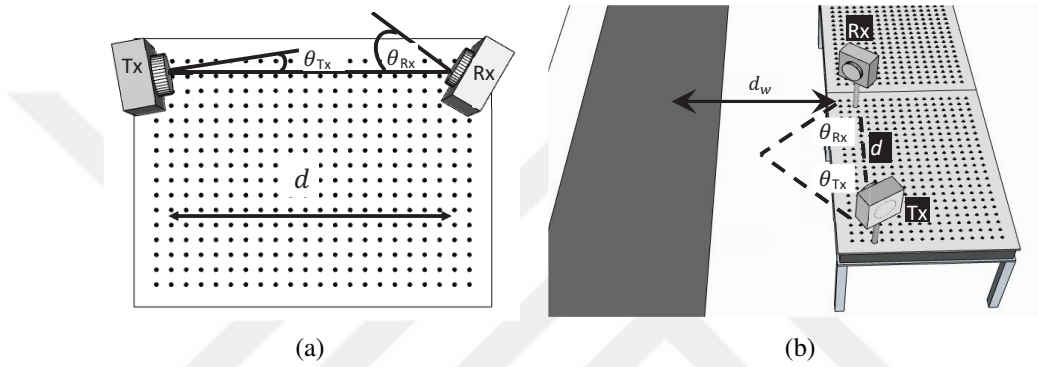


Figure 9: Experimental scenarios: (a) LOS (b) NLOS.

denoted by d . θ_{Tx} and θ_{Rx} respectively define the orientation of transmitter and receiver. In the LOS scenario (see Figure 9.a), we consider two cases: a) strict alignment between Tx and Rx units (i.e., $\theta_{Tx} = 0^\circ$) and b) some shift in the orientation of Tx (i.e., $\theta_{Tx} = 10^\circ$). In the NLOS scenario (see Figure 9.b), transmitter and receiver do not look at each other and the transmission depends only on the reflected light from a whiteboard. d_w denotes the distance between the whiteboard (used as a reflector) and Tx-Rx frontends.

In Figure 10, we present the received constellations and eye diagrams for LOS scenario. We assume strict alignment between transmitter and receivers and set $d = 1$ m. From Figure 10.a, it is observed that received constellation is noisy and the eye diagram is nearly closed when no lens is used. In order to improve the performance, we use a 24.5 mm plano-convex lens at Tx frontend to collimate the transmitted light pattern. This consequently increases the received SNR and gives better results as can be observed in eye diagrams presented in Figure 10.b.

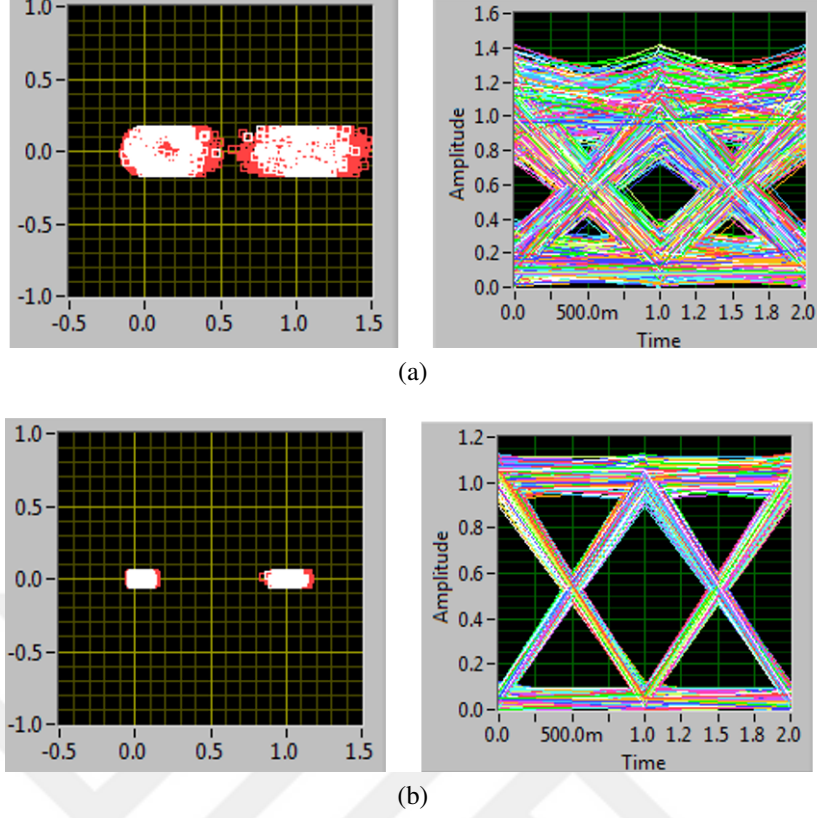


Figure 10: Constellation and eye diagram of received signal at 1 meter in LOS scenario with strict alignment (a) without lens (b) with lens.

In Figure 11, we present SNR and BER performance versus distance. We assume LOS scenario with strict alignment where d varies between 1 m and 3 m with a step size of 0.5 m. The estimate of SNR is obtained based on (2.35) in conjunction with (2.33) and (2.34) where noise power and effective channel are estimated. In Figure 11.a, it can be observed that at a distance of 1 m, 16.87 dB SNR is obtained if no lens is used. This decreases to -3.4 dB at $d = 3$ m. When a plano-convex lens is used at the transmitter, some improvements are observed. Specifically, we obtain SNR values of 26.65 dB and 8.49 dB at $d = 1$ m and $d = 3$ m respectively. These indicate SNR improvements of around 9.8 dB and 11.89 dB in comparison to no lens case.

For BER measurements, we transmit $N_F = 200$ frames for each SNR value. Each frame includes $L_D = 8000$ bit per frame. Therefore, the total number of transmitted bits is $N_F L_D = 200 \times 8000 = 1.6 \times 10^6$ bits. We compare the transmitted and received bits and present the BER for each SNR value in Figure 11.b. As a benchmark, we further include the analytical BER expression for OOK given by $BER = Q\left(\sqrt{\text{SNR}}\right)$ [68]. It can

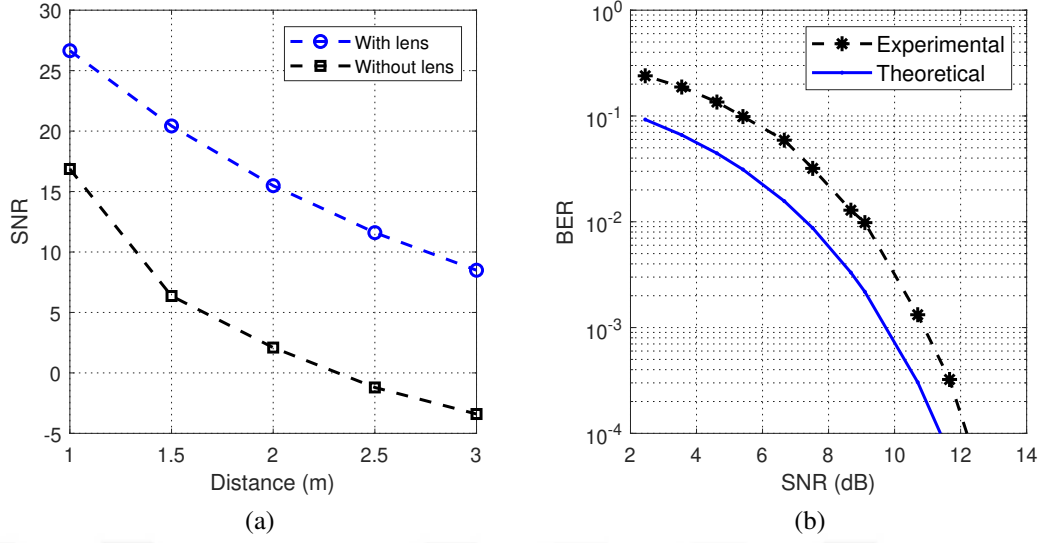


Figure 11: Experimental measurements: (a) estimated SNR versus distance (b) measured BER versus SNR.

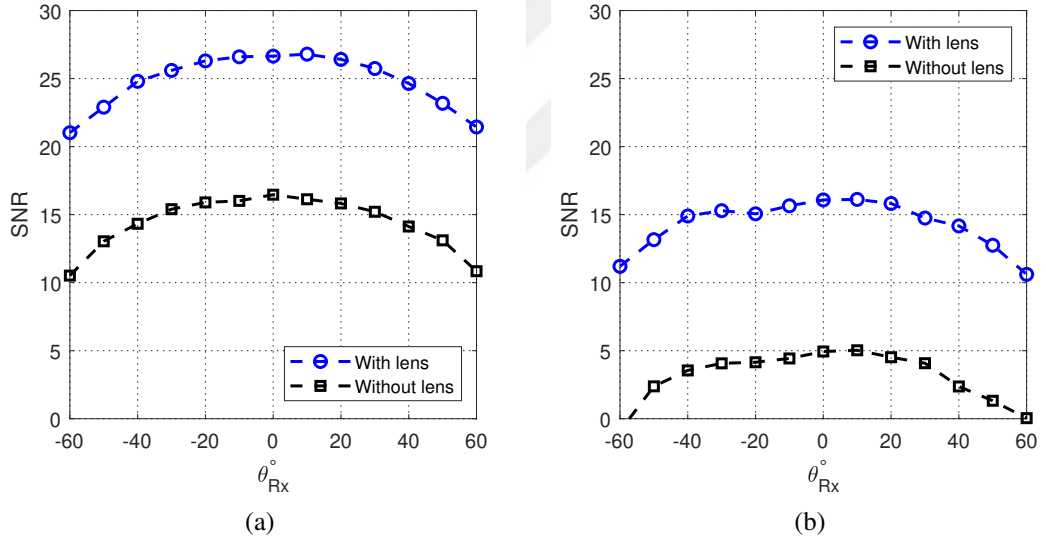


Figure 12: Receiver SNR versus reception angle for LOS scenario: (a) $\theta_{TX} = 0^\circ$ (b) $\theta_{TX} = 10^\circ$.

be readily observed that theoretical and experimental curves have a close match. There is about 1.8 dB shift between the two curves as a result of inaccuracy associated with symbol timing, frame synchronization and channel estimation. As SNR increases, the shift between these two curves decreases since estimation accuracy relatively improves.

In Figure 12, we investigate the effect of transmitter orientation for the LoS scenario. We consider $d = 1$ m and present SNR versus reception angle. For transmitter

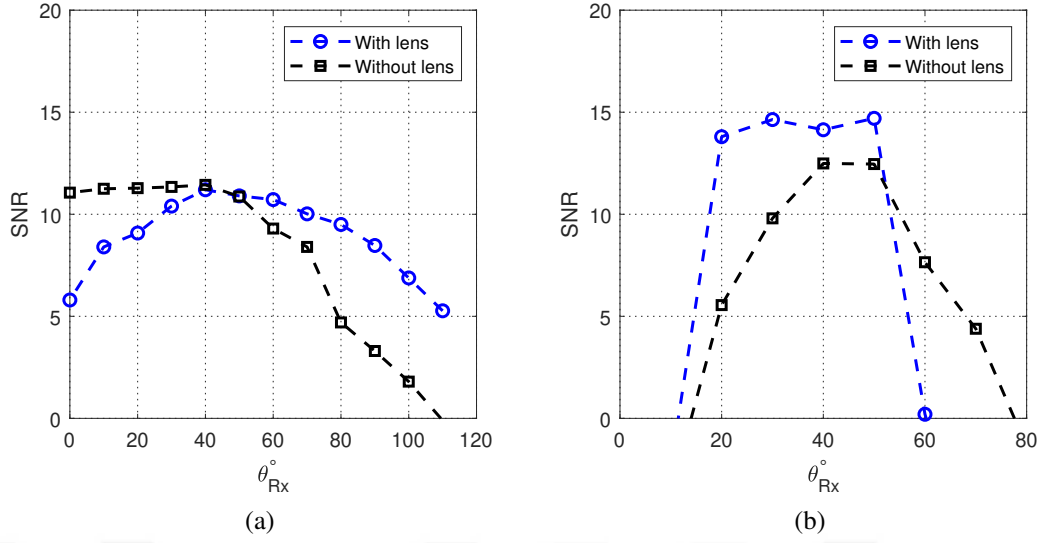


Figure 13: Receiver SNR versus reception angle for NLOS scenario at $d_w = 0.5$ m: (a) $\theta_{TX} = 60^\circ$ (b) $\theta_{TX} = 90^\circ$.

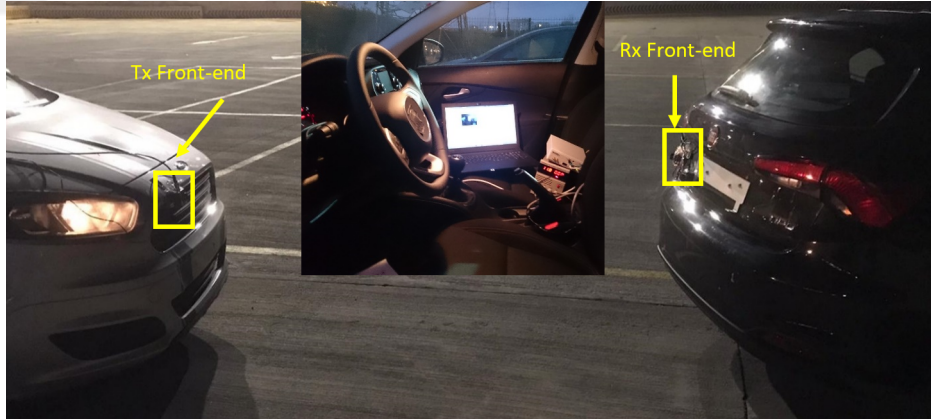
orientation, we assume $\theta_{TX} = 0^\circ$ (Figure 12.a) and $\theta_{TX} = 10^\circ$ (In Figure 12.b). As expected, it can be observed from Figure 12.a that the maximum SNR is observed at $\theta_{RX} = 0$ and SNR decreases as $|\theta_{RX}|$ increases. In Figure 12.b where we set $\theta_{TX} = 10^\circ$, the maximum SNR is obtained at $\theta_{RX} = 10$. In both cases, the employment of lens increases the received SNR by around 10 dB.

In Figure 13, we investigate the effect of transmitter orientation for the NLoS scenario. The transmitter and receiver are located at an equal distance from the reflecting surface with $d_w = 0.5$ m. We present SNR versus reception angle. It is observed that for $\theta_{TX} = 60^\circ$ (Figure 13.a), the received power is relatively low when the receiver orientation angle is less than $\theta_{RX} = 40^\circ$ assuming the employment of lens. Without a lens, PD provides a wider FOV which gains more power from the reflected rays even for values of θ_{RX} smaller than. At $\theta_{TX} = 90^\circ$, the receiver with lens starts to get signal at almost $\theta_{RX} = 20^\circ$. It reaches the maximum at 50° degrees, then it starts to degrade dramatically again. Without any lens, when the receive orientation angle is larger than $\theta_{RX} = 50^\circ$, PD receives more power from the reflected rays.

2.4 Selection of Proper Lens Combination

In this Section, we experimentally investigate the effect of using different lens combinations on vehicular VLC systems in outdoor environments. First, we obtain the optical channel coefficient of vehicular scenario based on non-sequential ray tracing approach in Zemax[®] similar to [74,75] and verify it through measurements. Then, we carry out measurements to obtain the effective channel coefficient which includes the effect of both frontends and propagation channel. Based on the estimated channel coefficients, we characterize the system performance in terms of SNR for all lens combinations under consideration. The effect of interference from other light sources and the effect of receiver orientation are further discussed.

As illustrated in Figure 14, the TX frontend was mounted at a height of 70 cm from the ground (same as front headlight). At the receiver side, a custom-design VLC frontend with 170-degree FOV [76] (mounted on the other vehicle at a height of 55 cm under the taillight) is used.



(a)

Figure 14: Selection of proper lens combination: Pictures taken from experiment.

To extend the communication distance as well as improve the received SNR, we investigate the use of different lens combinations at both transmitter and receiver side. Plano-convex lenses at transmitter side are used to collimate the light pattern. At the receiver side, plano-convex and biconvex lenses are selected to reduce the photodetector FOV and consequently enhance the receive SNR at further distances. As illustrated in

Figure 15, we consider four different combinations:

- **L0:** Tx and Rx frontends are without any lenses
- **L1:** A plano-convex 25.4 mm lens is attached to both Tx and Rx frontends
- **L2:** It is the same set-up as in L1 with an additional biconvex 75 mm lens as an outer lens at the Rx frontend
- **L3:** The transmitter employs a 25.4 plano-convex followed by an outer 35 mm bi-convex lens at the transmitter while the receiver uses an outer 200 mm bi-convex lens followed by 75 mm bi-convex lens at the receiver

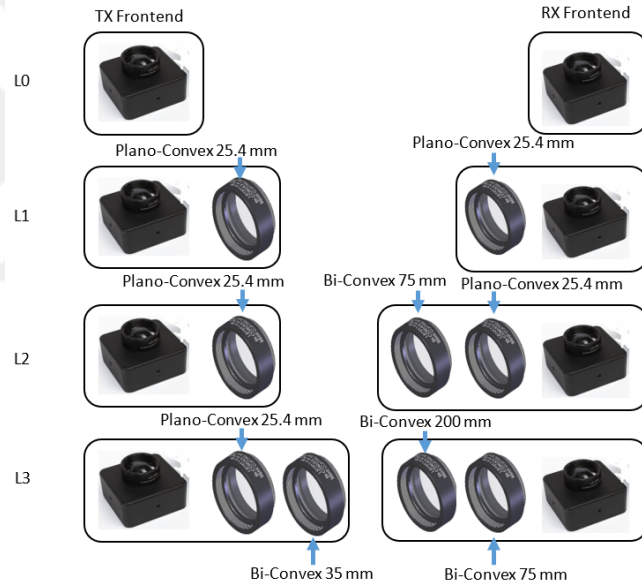


Figure 15: Lens combinations.

The measurements were taken in the Özyeğin University campus at the night time and in clear weather conditions. The scenarios under consideration are illustrated in Figure 16. In the first scenario, we consider two vehicles in the same lane separated by a distance of d . In the second scenario, we assume that the leading car is about to turn with angle θ between transmitter and receiver vehicles.

The effective channel coefficient (i.e., h_{eff}) includes the effects of both the propagation channel denoted by h_{opt} and system frontend, lenses and filters together represented

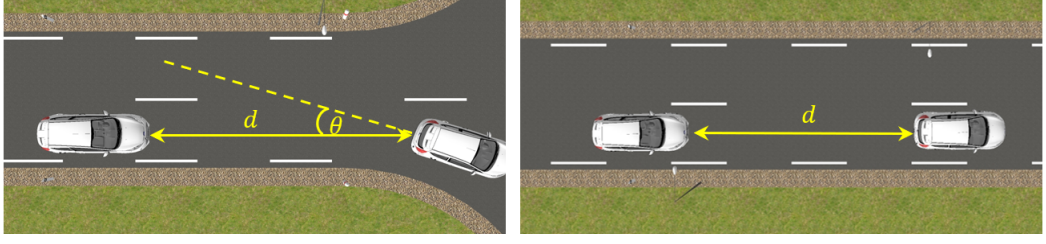


Figure 16: Scenarios under consideration.

by h_{sys} . To isolate these two effects, we first measured the received optical power using a power meter (Thorlabs S142C) and determined h_{opt} . In an effort to validate this measurement, we further conducted a simulation study using the optical design software OpticStudio[®] based on the methodology in [74] where the scenario under consideration was modeled as a 3D model and non-sequential ray tracing was used to obtain propagation path loss. The results give a very good match between the measured and simulated h_{opt} values as illustrated in Figure 17.

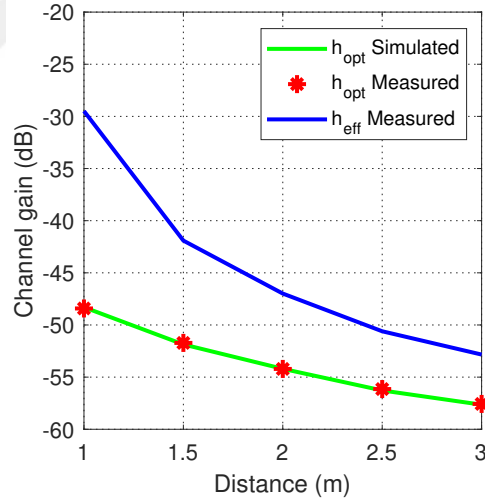


Figure 17: Comparison of optical and effective channel coefficients.

In Figure 18, we present the measured effective channel coefficient $\widehat{h_{\text{eff}}}$ with respect to distance for different lens combinations. We consider $\theta = 0^\circ$, i.e., perfect alignment between two vehicles. Curves associated with L1 lens set shows a sharp decay within 5 m while L2 gives a smoother decay because of the additional lens gain. In L2 combination, the additional lens reduces the FOV. Due to the different height (about 15 cm) between both Tx and Rx frontends, it degrades the system performance at distances

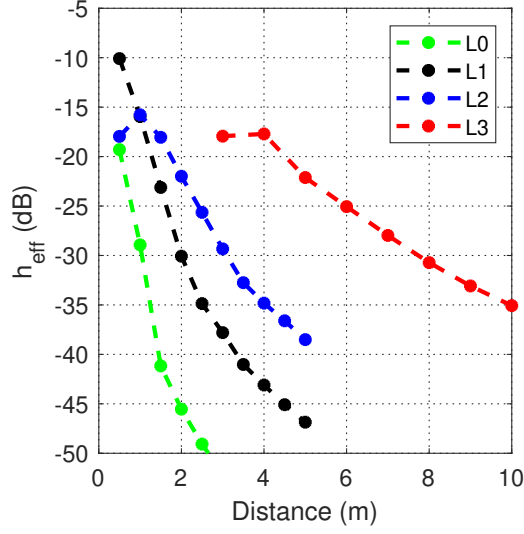


Figure 18: Effective channel coefficient for different lens combination at $\theta = 0^\circ$.

below 1 m. In L3 combination, the PD is saturated within the first 3 meters, however it is beneficial for longer distances and transmission span reaches to 10 m.

For $\theta = 10^\circ$, it is observed from Figure 18 that the effective channel coefficient for the case of L0 falls below 50 dB before the link distance of 2 m. Therefore, we did not include it for the case of $\theta = 10^\circ$. It is observed from Figure 19 that L1 and L3 follow the same decay curve as in $\theta = 0^\circ$. While L2 starts with a very low value at the distance of 0.5 m, it dramatically increases to peak around 1.5 m before it decays again.

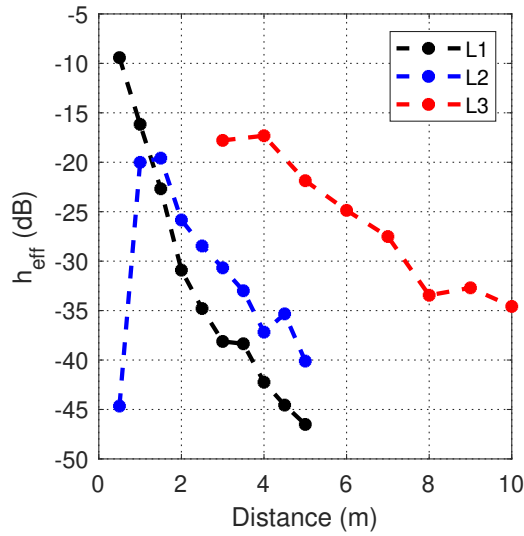


Figure 19: Effective channel coefficient for different lens combination at $\theta = 10^\circ$.

2.5 Conclusion

In this chapter, we experimentally evaluated the performance of different modulation techniques using USRP platforms. Through a measurement campaigns, for OFDM and OOK VLC systems, we measured the effective channel which includes the effects of both propagation channel and frontends. Then, we measure the BER and obtain the overall BER. Finally, we compare the experimental results with theoretical values to confirm our work. In addition, we investigate the performance considering different transmitter and receiver orientation angels. We further consider both LoS and NLoS scenarios with and without lens employment. As a comparison between OFDM and OOK VLC systems, it is observed that the implementation of OOK system has much lower complexity and lower overhead processing compared to OFDM systems. On the other hand, OFDM can support higher data rates. Due to the wide available license-free bandwidth offered by VLC, the lower complexity system (i.e., OOK) was selected for the rest of the dissertation. Moreover, we experimentally investigate the effect of using different lens combinations on vehicular VLC systems in outdoor environments. First, we measure the effective channel coefficient which includes the effect of both frontends and propagation channel. Then, based on the estimated channel coefficients, we characterize the vehicular system performance in terms of signal-to-noise ratio and quantify improvements from utilizing different lens combinations.

Overall, in this chapter we investigated the practical implementation aspects of different modulation techniques as well as the effect of the frontend. The next chapter focuses on investigating the effect of the background noises and the possibilities of extending the communications distances through hop employment.

CHAPTER III

MULTIHOP TRANSMISSION FOR END-TO-END CONNECTIVITY

3.1 *Introduction*

To satisfy the requirements of intelligent transportation system (ITS) applications [77], a reliable VLC link among vehicles is required [78]. The quality of vehicular VLC link depends on several factors, including visibility that limits the communication distance [79]. This necessitates the deployment of relay-assisted VLC systems to extend the transmission range. Such multihop systems are also desirable for emerging vehicle platoons to connect the follower vehicle with the leader vehicle [80].

Most of the current literature on multihop VLC are either simulation-based [81–83] or experimental works in controlled-lab environments [84–87]. To the best of the authors' knowledge, the performance of the multihop VLC system in a real outdoor environment has not been investigated yet. However, for having a precise characterization of the multihop vehicular VLC systems, it is necessary to take into account the effect of the outdoor environment.

In this chapter, we start by measuring the electrical SNR throughout a complete day (i.e., for 24 hours). Measurements demonstrate that SNR changes significantly during daytime due to ambient noise while it exhibits steady characteristics at nighttime. Based on this, we experimentally evaluate the performance of the vehicular multihop VLC system with two experiments, one in the daytime and one at nighttime. In our set-up, the source vehicle transmits through a low-beam LED while the destination vehicle receives the signal through a photodetector attached to its back. Our set-up is limited to two equipped vehicles which let us do the BER and SNR measurements for a single hop. To Characterize the presence of a multihop system, whenever the receiving vehicle

receives the bit stream it feeds the received bit stream back to the transmitting vehicle until making sure that the required number of hops is achieved.

The rest of the chapter is organized as follows. In Section 3.2, we illustrate the system model and the measurement campaign. In Section 3.3 we investigate the effect of background noise variation on the system performance. After that, we investigate the error rate for the under-consideration multihop system at different times (i.e. daytime and nighttime) for both individual hops and overall hops.

3.2 System Model

As shown in Figure 20, we consider a multihop vehicular VLC system where the source vehicle S communicates with a destination vehicle D through N relay vehicles denoted by R_j , $j = 1, 2, \dots, N$. Let d_j denote the length of the j^{th} hop, $j = 0, 1, \dots, N$. Therefore, end-to-end transmission distance becomes $d = \sum_{j=0}^N d_j$.

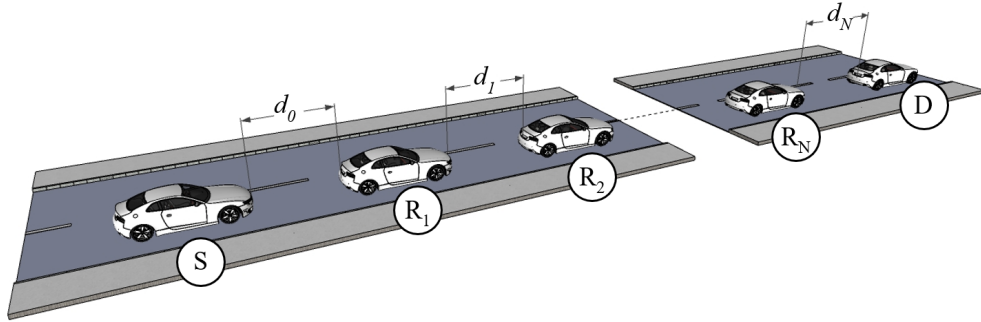


Figure 20: Vehicular multihop system.

One of the low-beam headlights of the transmit vehicle serves as wireless transmitter while the receiver vehicle is equipped with a photodetector as wireless receiver mounted at its back at the same height of the headlight. Let L_D denote the number of transmitted OOK symbols. The i^{th} transmitted symbol from the $(j-1)^{\text{th}}$ vehicle to the j^{th} vehicle is denoted as $S_j(i)$, $i = 1, \dots, L_D$. Due to the low output power of the generated symbol from the universal software radio peripheral (USRP), which is not enough for vehicular headlamp, the generated signal is further amplified with factor A . before adding a DC

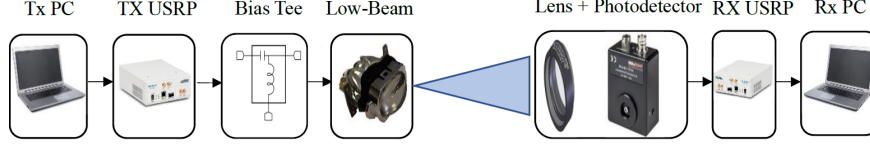


Figure 21: Block diagram of the experimental set-up of multihop vehicular VLC systems.

bias value of V_{DC} . The headlight's driving signal can be expressed as

$$x_j(t) = \left(A\sqrt{P} \sum_{i=1}^{L_D} S_j(i) \otimes p_T(t - iT_s) \right) + V_{DC}, \quad (3.1)$$

where P is average electrical transmit power per hop, $p_T(t)$ is transmit pulse shaping filter and T_s is pulse duration. The transmitted light propagates through an optical channel and received by the photodetector of destination vehicle. Let h_j , $j = 0, 1, \dots, N$ denote the “effective” channel coefficients incorporating the effects of both frontends and propagation channel of the j^{th} hop. Assuming that the pulse duration is much larger than the delay spread, the channel is considered as a single tap, which can be justified for ITS messages [88]. Consequently, the received signal can be expressed as

$$r_j(i) = \sqrt{P}h_j(i)S_j(i) + w_j, \quad (3.2)$$

where w_j is AWGN at j^{th} vehicle with zero mean and variance of $\sigma_{n_j}^2$.

The experimental set-up is illustrated in Figure 21. The digital signal waveform, generated using LabVIEW software, is transferred via Ethernet-cable to NI-2920 USRP platform for DAC process. For baseband signal generation, the USRP platform is modified with LFTX daughterboard [89]. Mini-circuits ZHL-32A+ amplifies the analogue signal then it is delivered to a custom-designed high-power bias-tee which can drive a headlight with a current up to 2.5 A. The bias-tee is used for adjusting the operating point (i.e., selecting the proper DC-bias value) to drive the vehicular low-beam headlamp.

At the receiver side, a Bi-convex 24.5 mm lens is used for enhancing the received signal. This lens focuses the light on the effective area of Thorlabs photodetector (APD130 A/M). The photodetector is feeding a NI-2920 USRP platform, modified with

Table 4: Experimental set-up parameters of multihop vehicular VLC system.

Average electrical transmitted power	P	[0, -2, ..., -10] dB at daytime, [-14, -16, ..., -24] dB at nighttime
Filter length of root-raised cosine (RCC) pulse shaping	L_{PS}	40
Hop distance	d_i	40 m
Length of channel estimation's training sequence	L_{CE}	80
Length of frame detection's training sequence	L_{FD}	80
Length of guard band	L_G	6
Number of frames	N_F	250
Payload	L_D	8000 bit/ frame
Roll-off factor	β	0.5
Sampling rate of NI-USRP platform	f_s	1×10^6 Sample/sec
Up-sampling factor	U	20

LFRX daughterboard, for ADC. LabVIEW software on the receive PC handles the rest of the online processing.

Our set-up is sufficient to make measurements for a single-hop between two vehicles. To experimentally characterize the DF multihop VLC transmission, the received bit sequence was fed back to the transmitting unit and processed $N + 1$ times. The experimental set-up parameters are listed in Table 4.

3.3 Effect of Outdoor Noises

The experiment is conducted at Ozyegin University campus on a semi-cloudy weather day.

First, we investigate the effect of the background noise on SNR. The transmit and receive vehicles are separated with a distance of 20 m. We take a full-day measurement to investigate the variation of background noise with respect to time. The average electrical transmit power P is fixed to 0 dB. The estimated receive electrical SNR with

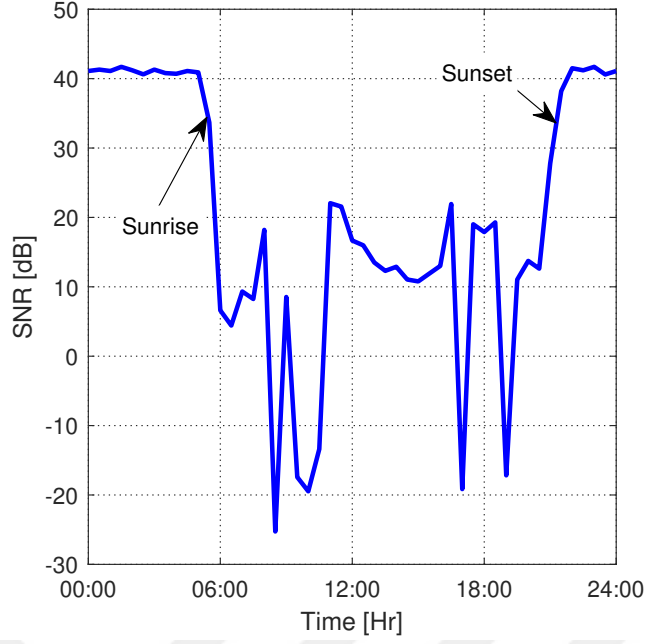


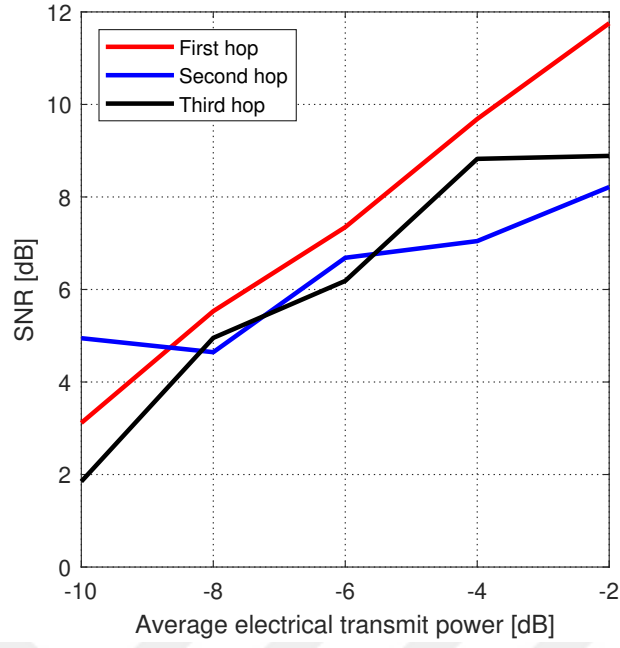
Figure 22: Measurements of the receive electrical SNR on 4th June 2020.

respect to time is illustrated in Figure 22. It is observed that there is a huge background noise variation. The reason behind SNR fluctuations during the daytime is the presence of clouds where sun can appear and disappear from time to time in such a semi-cloudy weather. It is also observed that the photodetector saturates at some time instants, see e.g., measurements at 10:00 am and 5:00 pm.

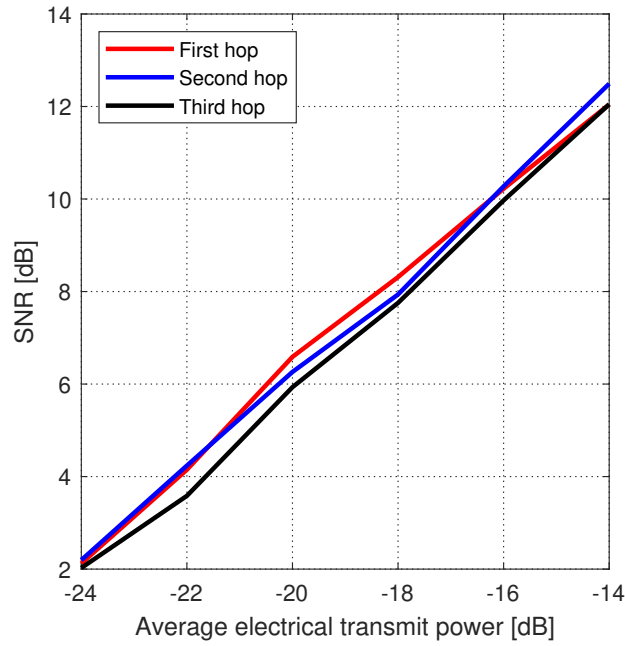
In Figure 23, we present the measured SNR for different average electrical transmit power during both day and nighttimes. For instance, at $P = -4$ dB, the received SNR for the first, second and third hops, respectively, are 9.73 dB, 7.11 dB and 8.56 dB. While SNR changes from hop to hop at daytime, it is observed that at nighttime it remains almost fixed. For instance, at $P = -14$ dB, the received SNR for the first, second and third hops are, respectively, 12.01 dB, 12.3 dB and 12.09 dB.

3.4 Error-Rate Analysis

In the following, we present the BER measurements. For these measurements, we transmit a number of frames of $N_F = 250$. Each frame contains a payload of $L_D = 8000$ data bits. The total number of transmitted bits is 2×10^6 bits. We calculate the experimental BER using LabVIEW by comparing received bit stream with the actual transmitted



(a)



(b)

Figure 23: Measurements of the receive electrical SNR for individual hops: (a) day-time and (b) nighttime.

ones. To experimentally characterize vehicular multihop system, the received bit stream at the receiving vehicle is fed back to the transmitting vehicle until making sure that the required number of hops is achieved. The theoretical BER is included as a benchmark. For DF serial relaying system, it is given by $BER = \sum_{k=0}^N \left(BER_k \prod_{j=k+1}^N (1 - 2BER_j) \right)$

[90], where $BER_j = Q(\sqrt{SNR_j})$ [68] is the j^{th} hop BER.

In Figure 24, we investigate the error rate performance for individual hops in a multihop system. Particularly, we consider three hop system and investigate the BER of the first, second and third hops by comparing the received bit stream at the first receiving vehicle with the original transmitted bit stream, comparing the received bit stream at the second receiving vehicle with the re-transmitted bit stream from the first receiving vehicle and comparing the received bit stream at the third receiving vehicle with the re-transmitted bit stream from the second receiving vehicle.

It is noted that the hop's BER performance at nighttime is almost constant. For example, consider an average electrical transmit power of $P = -18$ dB, the BER per hop is obtained as 2.36×10^{-2} , 2.87×10^{-2} and 2.54×10^{-2} for first, second and third hops. Unlike the BER performance at nighttime, which is almost the same, it is observed that the BER performance at daytime changes dramatically. For example, consider an average electrical transmit power of $P = -6$ dB, the BER per hop is obtained as 3.94×10^{-2} , 5.79×10^{-2} and 1.51×10^{-2} for first, second and third hops indicating relative change percentage for the second and third hops of 49.9% higher and 61.6% lower than the error rate of the first hop, respectively. As a comparison with the theoretical values, it is observed that, for single hop at a given BER of 1×10^{-2} , there is an average difference between the experimental and theoretical of 1 dB for nighttime compared to 1.25 dB for daytime.

In Figure 25, we consider single-hop, dual-hop and three-hop systems and present overall multihop system's BER performance. At nighttime, as expected, the dual-hop error rate is higher than the single-hop. Similarly, the three-hop has a higher error rate than both single- and dual- hops. Particularly, at the daytime, there is a 2 dB difference between dual and single-hop and 1 dB between dual and three-hop. At nighttime, there is a 0.5 dB difference between dual and single-hop and 0.5 dB between dual and three-hop. The difference between the experimental and theoretical values for dual and three hops, respectively, at nighttime is 1.3 dB and 1.61 dB compared to 1.43 dB and 2 dB

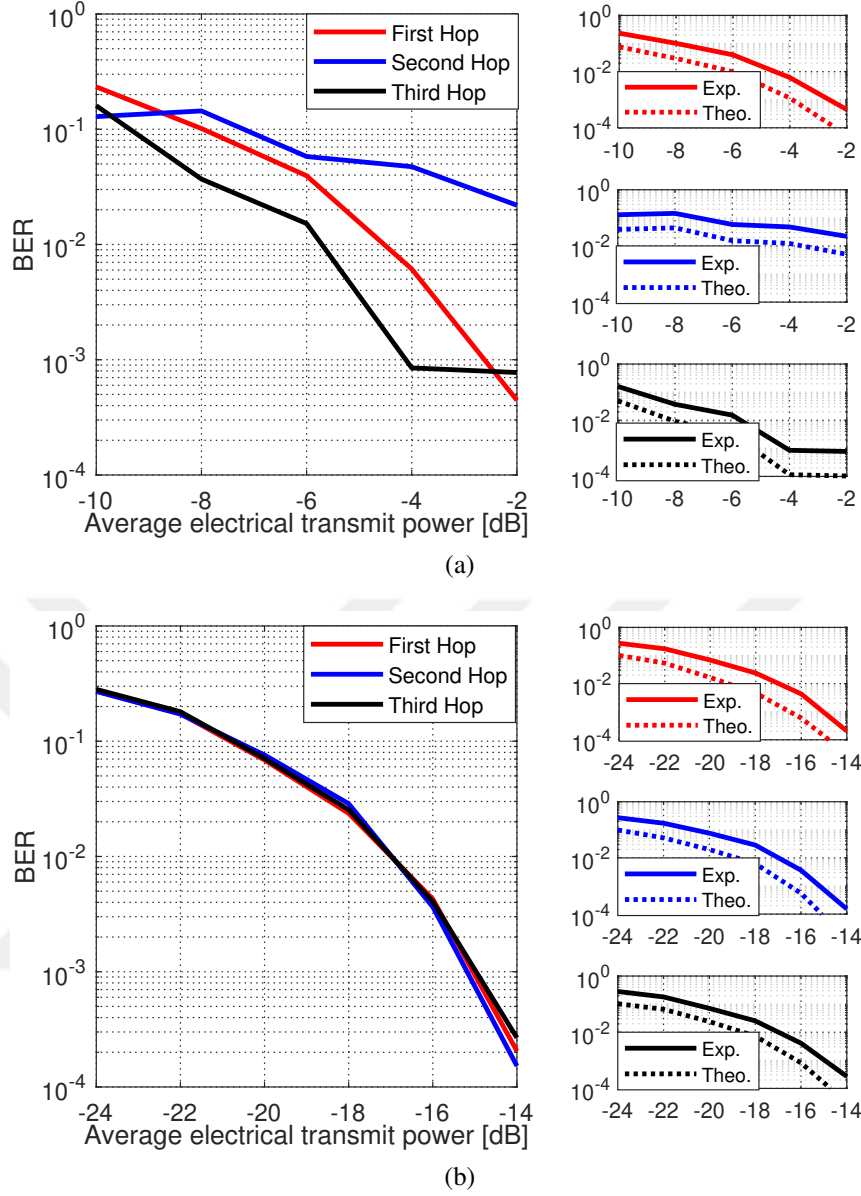


Figure 24: BER for individual hops (System with three hops is assumed): (a) daytime and (b) nighttime.

at the daytime. For BER of 1×10^{-2} , average electrical transmit power of -16 dB is required for three-hop system at nighttime. At daytime, the average required electrical transmit power for three-hop system jumps to -2 dB indicating extra needed power of 14 dB.

To further subpress the effect of the background noises and maintain a steady performance during day time, we employed an optical band pass filer. This filter passes the light wavelength between 450 nm and 550 nm while blocking the other wavelengths (i.e., background noises). These wavelengths were selected based on the wavelength

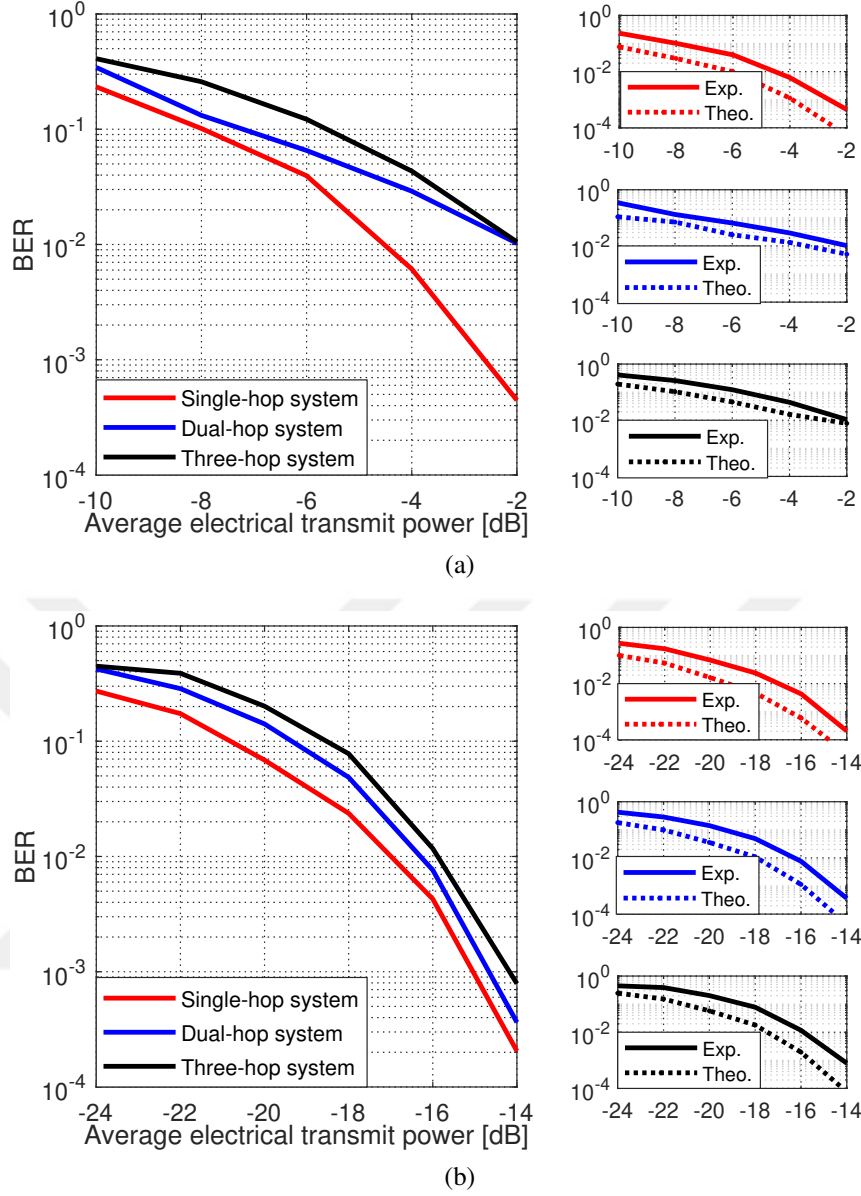


Figure 25: Overall multihop system's BER performances: (a) daytime and (b) night-time.

characteristics of the headlight. The effect of the filter employment is illustrated in Figure 26. As measurements indicates, a steady performance is observed with filter employment compared to the unstable behaviors of the filter unemployment. On the other hand, for night time where the background noises are absent, the performance of the filter unemployment case outperform the filter employment case with an average of around 8 dB.

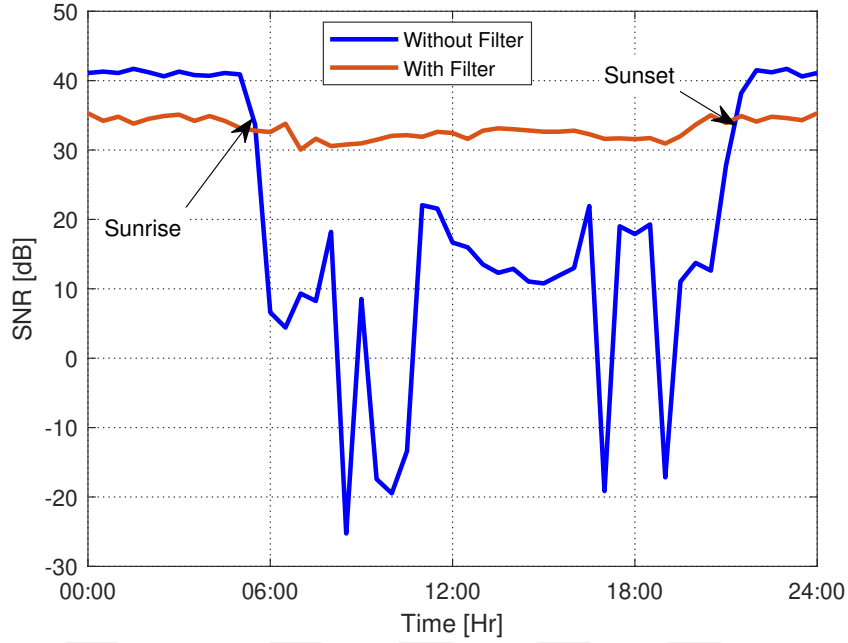


Figure 26: Measurements of the receive electrical SNR with and without filter employment on 4th June 2020.

3.5 Conclusion

The reliable vehicular VLC communication distance may be relatively short due to its dependence on many factors including the weather condition. This necessitates the use of relay-assisted systems to extend the transmission range. In this chapter, we investigate experimentally the performance of vehicular multihop VLC systems in outdoor environments. Particularly, we first examine the effect of ambient noise on SNR during a whole day. Then, we conduct two multihop experiments, i.e., one at daytime and one at nighttime. We measure the BER for each individual hop and then obtain the overall performance of the multihop system. We further compare experimental BER with theoretical BER to confirm our results. Furthermore, we stabilized the performance during the day time through employing an optical filter which, on the other hand, reduces the received SNR by an average of 8 dB.

To Conclude, the last chapters detailed the practical implementation and performance investigation only on static conditions. The next chapter investigates the effect of mobility as well as the performance in different road types.

CHAPTER IV

VEHICULAR VLC LINKS IN THE PRESENCE OF MOBILITY

4.1 *Introduction*

A critical concern in the practical implementation of vehicular VLC systems is the number of photodetectors and their locations on the vehicle. Majority of the literature on vehicular VLC is limited to the deployment of a single photodetector [91–98]. Employing a single receiver might result in an outage in the cases of imperfect alignment due to lateral shift, road curvatures, etc. [97, 98]. This motivates the use of multiple photodetectors in vehicular VLC systems. However, most of the initial experimental works in this area are mainly limited to some indoor tests under static channel conditions and/or the use of indoor-type LEDs [99, 100].

The existing experimental works deploy indoor-type LEDs which exhibit different characteristics from vehicular headlights with asymmetrical intensity distributions. Furthermore, a comprehensive performance analysis of vehicular VLC systems under mobility conditions has not been yet presented. To fill these research gaps, we first experimentally investigate the usage of multiple photodetectors for a vehicular VLC system with selection combining. In our experimental set-up, the source vehicle transmits through a low-beam LED headlight [101] while the destination vehicle receives the signal through two photodetectors located at its back and on the side. Based on the measurement campaign, we first identify the optimal orientation angle of the side photodetector. Then, we measure the SNR for each of the photodetectors individually as well as at the output of the selection combiner. To examine the mobility effect on the system performance, we consider tracks with straight and curved roads. Both day time and night time measurements are conducted. Through data fitting to experimental data, we obtain a Probability Density Function (PDF) to describe the instantaneous

SNR. Based on the PDF, we derive a closed-form expression for the BER and compare it with the measurements.

The rest of this chapter is organized as follows: In Section 4.2, we describe the system model and estimation steps. In Section Section 4.3, we present the measurement campaign results. In Section 4.4, we present the BER analysis and compare it with measurements. Finally, the chapter is concluded in Section 4.5.

4.2 System Model

We consider a vehicular VLC system with OOK. The experimental set-up is illustrated in Figure 27. One of the low-beam headlights of the source vehicle serves as the wireless transmitter. The destination vehicle is equipped with two photodetectors as wireless receivers. Both photodetectors are mounted at the same height of the low-beam headlight. While the first photodetector looks backward, the second one is oriented to have a deviation angle towards the outer side.

At the transmit vehicle, the digital waveform is generated by LabVIEW software and transferred through an Ethernet cable to a modified universal software radio peripheral (USRP NI 2920). The generated analog signal is then amplified by a Mini-circuits ZHL-32A+ amplifier and delivered to a custom-design bias-tee for adding a DC level. The bias-tee output signal is used to drive the low-beam headlamp. The 3dB-bandwidth of the transmitter frontend is measured as 1.2 MHz.

At the receiver side, two photodetectors (Thorlabs APD 130 A/M) were mounted on the back and side of the destination vehicle. To investigate the effect of FOV, we consider both cases with and without lens employment. The photodetector has a wide FOV of 80 degrees. The use of a plano-convex lens with a 25.4 mm focal length narrows down the FOV to 40 degrees. Photodetectors feed two synchronized USRPs for analogue-to-digital conversion. The rest of the online processing is handled by LabVIEW software on the receive PC.

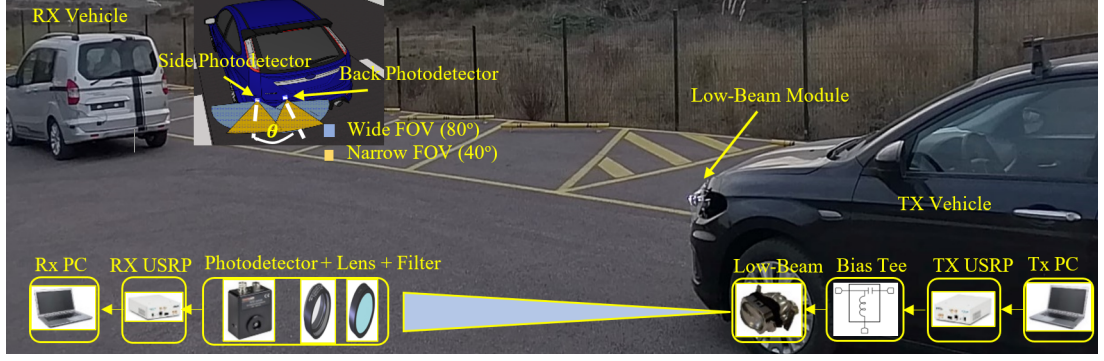


Figure 27: The experimental set-up.

The transmission frame consists of Synchronization Header (SHR) and a PHY payload. With a length of L_{SHR} , SHR is used for frame detection, DC-bias estimation and channel estimation purposes. In particular, a Pseudorandom Binary Sequence (PRBS) with good auto-correlation characteristics [102] should be selected for these purposes. SHR is then followed by data payload with a length of L_D . Therefore, the frame has a total of $L_{Frame} = L_{SHR} + L_D$ samples.

Let $S(i)$, $i = 0, 1, \dots, L_D - 1$ denote the i^{th} transmitted OOK data symbol. The continuous-time waveform that drives the LED headlight is expressed as

$$x(t) = \left(A \sum_{i=0}^{L_D-1} S(i) \delta(t - iT_s) \right) \otimes p_T(t) + V_{DC}, \quad (4.1)$$

where $p_T(t)$, T_s , A , V_{DC} and $\delta(\cdot)$ denote, respectively, transmit pulse shaping filter, pulse duration, amplification factor, DC bias and Dirac delta function. The received continuous-time waveform signal by the j^{th} photodetector (i.e., $j = 1$ is the back photodetector and $j = 2$ is the side photodetector) can be then written as

$$r_j(t) = \left(\sum_{i=0}^{L_D-1} S(i) \delta(t - iT_s) \right) \otimes h_j(t) + V_{Rj-DC} + w_j(t), \quad j = 1, 2, \quad (4.2)$$

where $V_{Rj-DC} = R_j V_{DC} \otimes l(t) \otimes h_{j-opt}(t) \otimes p_R(t)$ with R_j , $l(t)$, $h_{j-opt}(t)$ and $p_R(t)$, denoting, respectively, photodetector responsivity, headlight response, Channel Impulse Response (CIR) between the headlight and the j^{th} photodetector and receive pulse shaping filter. In (4.2), $w_j(t)$ is the additive white Gaussian noise with zero mean and noise variance of σ_{j-n}^2 and $h_j(t)$ denotes the effective CIR that incorporates the effects of

headlight response, optical CIR, pulse shaping filters, amplification factor and photodetector responsivity, i.e., $h_j(t) = AR_j p_T(t) \otimes l(t) \otimes h_{j-opt}(t) \otimes p_R(t)$.

At the receiver side, first symbol timing is performed to estimate the sampling point of the symbol using maximum output energy method [62]. Frame detection is then performed by correlating the received signal with the SHR sequence denoted by $\mathbf{T}_{SHR}$. The frame beginning is identified at the position of the autocorrelation's peak [102]. Let $r_j(i)$, $h_j(i)$ and $w_j(i)$ denote, respectively, the i^{th} , $i = 0, 1, \dots, L_{\text{Frame}} - 1$, sample of $r_j(t)$, $h_j(t)$ and $w_j(t)$. The sampled version of the received signal by the j^{th} photodetector can be written as

$$r_j(i) = S(i)h_j(i) + V_{Rj-DC} + w_j(i). \quad (4.3)$$

In Selection Combining (SC) technique, the photodetector that experiences the highest SNR is selected. Mathematically speaking, the selected signal can be expressed as

$$r_{\text{sel}}(i) = C r_1(i) + (1 - C) r_2(i), \quad (4.4)$$

where C is the combining coefficient and it is given as

$$C = \begin{cases} 1 & , \widehat{SNR}_1 \geq \widehat{SNR}_2 \\ 0 & , \text{Otherwise} \end{cases}, \quad (4.5)$$

where $\widehat{SNR}_j$ is the estimated SNR of the received signal by the j^{th} photodetector (See Appendix for the estimation steps). Data symbol can be then detected as

$$\widehat{S}(i) = \frac{r_{\text{sel}}(i) - \widehat{V_{Rsel-DC}}}{\widehat{h_{\text{sel}}}}, \quad i = 0, 1, \dots, L_D - 1, \quad (4.6)$$

where $\widehat{V_{Rsel-DC}}$ and $\widehat{h_{\text{sel}}}$ are, respectively, the estimated DC bias and estimated effective channel coefficient of the selected photodetector.

For SNR estimation, the training sequence consists of zeros and ones and could be

written as

$$\mathbf{T}_{\text{SHR}} = \begin{bmatrix} 1 & 1 & 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 0 & 0 \\ 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 1 & 1 & 0 & 0 \end{bmatrix}. \quad (4.7)$$

Let $\mathbf{I}_Z$ and $\mathbf{I}_O$ respectively denote the sets of locations for zeros and ones within $\mathbf{T}_{\text{SHR}}$. It can be readily checked from (4.7) that zeros and ones are located, respectively, at indices of $\mathbf{I}_Z = [4, 5, 8, 9, 10, 11, 12, 13, 16, 17, 18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28, 29, 30, 31, 32, 33, 34, 35, 38, 39, 44, 45, 46, 47, 48, 49, 50, 51, 52, 53, 62, 63, 66, 67, 72, 73, 74, 75, 78, 79]$ and $\mathbf{I}_O = [0, 1, 2, 3, 6, 7, 14, 15, 36, 37, 40, 41, 42, 43, 54, 55, 56, 57, 58, 59, 60, 61, 64, 65, 68, 69, 70, 71, 76, 77]$. Based on (4.3), the corresponding received samples for zeros and ones of the training symbols are, respectively, given by

$$r_Z(i_Z) = V_{Rj-DC} + w_j(i_Z), \quad i_Z \in \mathbf{I}_Z. \quad (4.8)$$

$$r_O(i_O) = h_j(i_O) + V_{Rj-DC} + w_j(i_O), \quad i_O \in \mathbf{I}_O. \quad (4.9)$$

Estimation of noise power and effective channel coefficient requires estimating the DC-bias V_{Rj-DC} and its subsequent removal. The DC bias is estimated by applying sample-mean method [103] to (4.8) as $\widehat{V_{Rj-DC}} = (1/L_Z) \sum_{i_Z \in \mathbf{I}_Z} r_Z(i_Z)$. After removing the DC-bias, the noise power at the j^{th} photodetector can be estimated from (4.8) as

$$\widehat{\sigma_{j-n}^2} = \frac{1}{L_Z} \sum_{i_Z \in \mathbf{I}_Z} \left| r_Z(i_Z) - \widehat{V_{Rj-DC}} \right|^2. \quad (4.10)$$

Similarly, the effective channel coefficient can be estimated from (4.9) as

$$\widehat{h_j} = \frac{1}{L_O} \sum_{i_O \in \mathbf{I}_O} \left(r_O(i_O) - \widehat{V_{Rj-DC}} \right). \quad (4.11)$$

Utilizing (4.10) and (4.11), the estimated SNR of the j^{th} photodetector can be then written as

$$\widehat{SNR_j} = \frac{\widehat{h_j}^2}{\widehat{\sigma_{j-n}^2}}. \quad (4.12)$$

The estimated SNR at the output of the selection combiner is then expressed as

$$\widehat{SNR_{\text{sel}}} = C \widehat{SNR_1} + (1 - C) \widehat{SNR_2}. \quad (4.13)$$

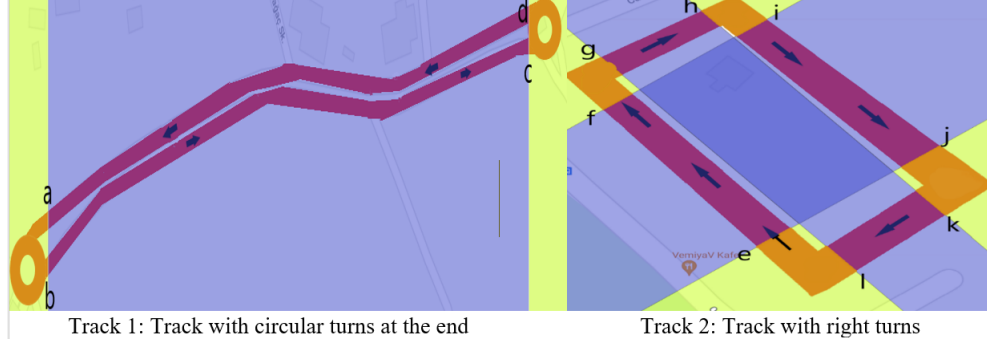


Figure 28: Two tracks on the university campus where the measurements are taken.

4.3 Measurement Campaign and Analysis

In our measurement campaign, unless otherwise stated, we use roll-off factor of $\beta = 0.5$, up-sampling factor of $U = 20$, USRP sampling rate of $f_s = 1 \times 10^6$ sample/sec. The SHR is selected as in (4.7) with length of $L_{\text{SHR}} = 80$. It consists of $L_Z = 50$ zeros and $L_O = 30$ ones. The frame payload is equal to $L_D = 4000$ and the total number of frames is $N_F = 1800$.

To investigate the effect of mobility and road type, we measure the SNRs in two different tracks. As illustrated in Figure 28, one track involves circular turns at the end (Track 1) and the other has right turns (Track 2). Both tracks include nearly straight roads (See blue highlighted parts) as well as curved roads (See yellow highlighted parts). The straight roads are indicated between the following locations; $b \rightarrow c$, $d \rightarrow a$, $e \rightarrow f$, $g \rightarrow h$, $i \rightarrow j$ and $k \rightarrow l$. The curved roads are marked between the following locations; $a \rightarrow b$, $c \rightarrow d$, $f \rightarrow g$, $h \rightarrow i$, $j \rightarrow k$ and $l \rightarrow e$. The measurements are made at both day and night time in clear weather conditions.

A critical design issue is the orientation of side photodetector. While vehicles move in a straight road, the back photodetector is expected to be the primary receptor since it would collect most of the power. However, in a curved road, to ensure reception in all cases, the presence of a side photodetector with a properly selected orientation angle is critical. To select the optimal orientation, we initially conducted the experiment in Track 1 for different values of orientation angle (denoted by θ) considering both cases (i.e., with and without lens employment). The connection is assumed to be lost if the

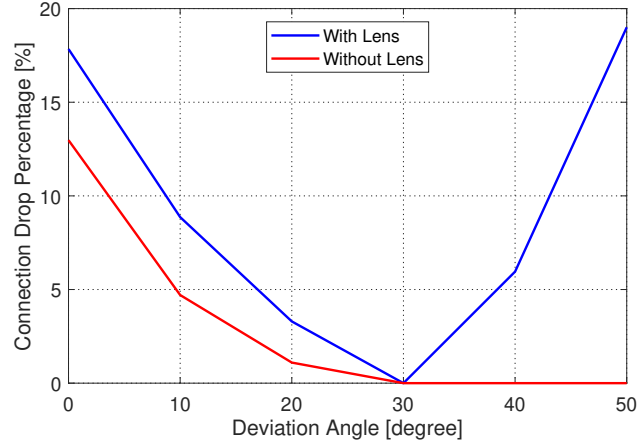


Figure 29: Connection drop percentage for different deviation angle.

received SNR drops below 10 dB (which gives a BER around 10^{-3} for OOK modulation). The percentage of connection drop for different deviation angles is illustrated in Figure 29. Without lens employment, connection drop takes its highest value at 0 degree and starts to decrease until it reaches zero at $\theta = 30^\circ$. A reliable connection is maintained as long as $\theta \geq 30^\circ$. With lens employment, it is observed that the connection drop percentage has a high value at $\theta = 0^\circ$ and gradually improves until it reaches zero at $\theta = 30^\circ$. After that, it starts to degrade dramatically. Based on these observations, we choose the orientation angle of $\theta = 30^\circ$ for side photodetector in the rest of the measurements.

Here, we present the measured SNRs over a time duration of 30 minutes which is the travel time to complete the tracks for 20 round trips. For each track, we consider four cases: (a) Day time measurements without lens employment, (b) day time measurements with lens employment, (c) night time measurements without lens employment and (d) night time measurements with lens employment.

In Figure 30, we present the received SNR measured during day time with lens employment in both tracks for a single trip. The yellow and blue highlighted parts in this figure correspond to the yellow and blue parts of the tracks in Figure 28. In Figure 30.a (For Track 1), a robust link with no connection drop is observed. During the whole track. The same robust behaviour is observed for all considered cases. The measured time-average SNR for different cases are listed in Table 5. It is further observed that

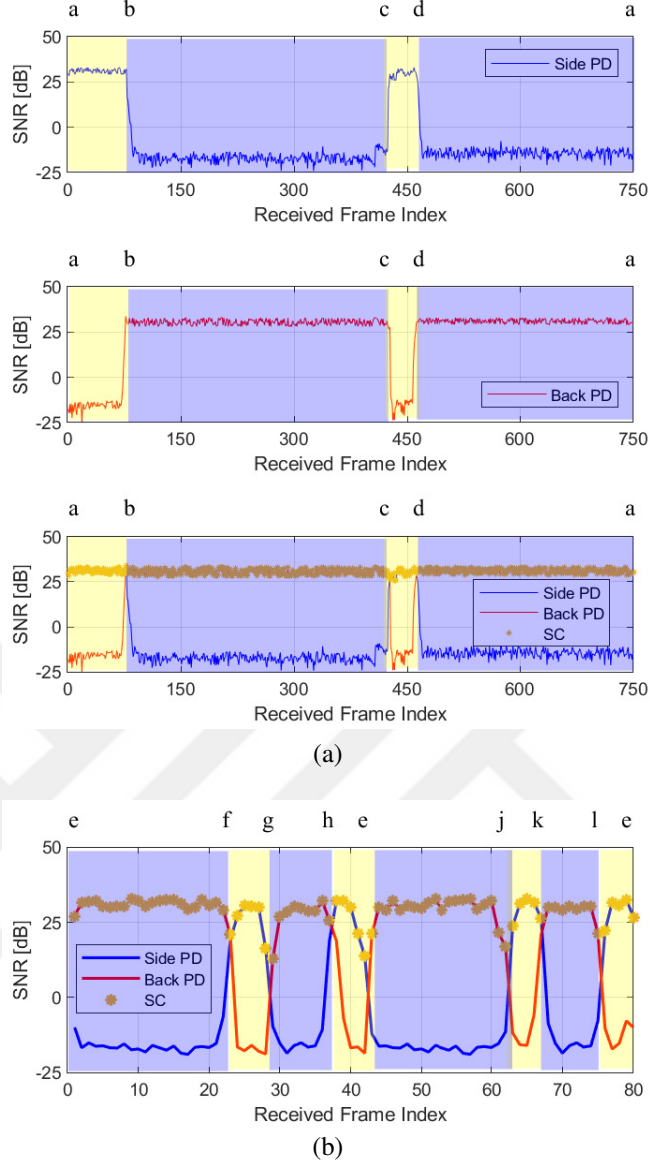


Figure 30: Received SNR at (a) Track 1 and (b) Track 2. Day time measurements with lens employment are assumed.

the back photodetector is always selected in straight roads (See blue highlighted parts). This makes the use of back photodetector sufficient for straight roads. On the other hand, the side photodetector is always selected at curved roads (See yellow highlighted parts). This highlights the contribution of the side photodetector and its importance to maintain the VLC link during curvatures.

In Figure 30.b, we present the received SNR measured at Track 2 for a single trip. The yellow and blue highlighted parts in this figure correspond to the yellow and blue parts of the Track 2. Similar to earlier observations, by symmetry, it is observed that

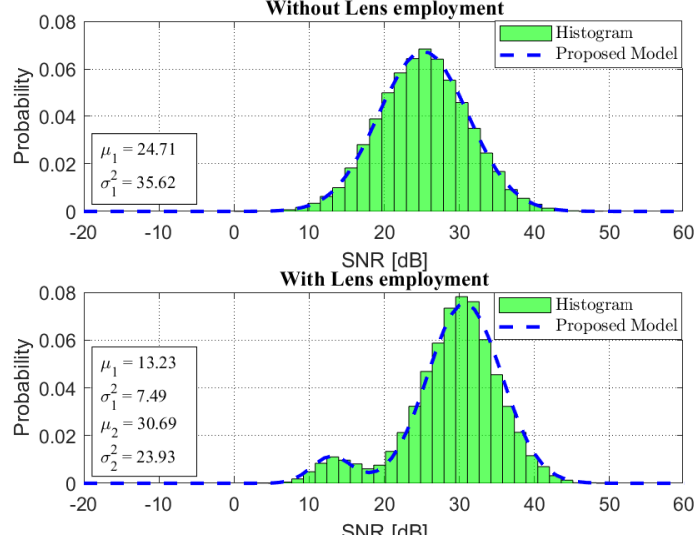


Figure 31: Histograms of measured SNR at the output of the selection combiner. Day time measurements are assumed.

the side photodetector is necessary during turns (i.e., curved roads). This makes the employment of both back and side photodetector necessary in tracks that involve both straight and curved roads in order to maintain link reliability in the presence of mobility.

Based on the received SNR values, we further investigate the maximum achievable distance between two vehicles. We determine the maximum distance such that the received SNR remains more than 10 dB which yields $\text{BER} \approx 10^{-3}$ for OOK modulation. Our results demonstrate that, with lens employment, the maximum achievable distance is 50 meters at night time and reduces to 42 meters at day time due to higher background noises. If there is no lens, achievable distances further reduce to 39 m and 29 m, respectively, at night time and day time.

In Figure 31, we present the histograms of the measured SNRs at the output of selection combiner respectively with and without lens employment. Day time measurements are considered in this figure. Similar observations are made for night time measurements.

It is observed that the PDF has either one or two peaks. Based on the MATLAB built-in function “dfittool” [104], weighted Gaussian PDF is found to be the best fit.

Table 5: Comparison between measured and analytical values

	Track	Case (Lens)	Measured SNR (Time- average) [dB]	Measured BER	Weighting Coefficients		Analytical BER
					F1	F2	
Day	1	Without	24.71	1.031×10^{-4}	0	1	6.741×10^{-5}
		With	28.89	4.616×10^{-5}	0.1	0.9	3.518×10^{-6}
	2	Without	23.12	9.667×10^{-5}	0	1	1.652×10^{-5}
		With	33.17	0	0.1	0.9	7.533×10^{-8}
Night	1	Without	38.42	0	0	1	4.099×10^{-11}
		With	43.26	0	0.1	0.9	9.308×10^{-14}
	2	Without	36.82	0	0	1	4.136×10^{-11}
		With	42.36	0	0.1	0.9	4.502×10^{-12}

This PDF is given by

$$f_x(x) = \sum_{i=1}^2 \frac{F_i}{\sqrt{2\pi\sigma_i^2}} \exp\left(-\frac{(\mu_i - x)^2}{2\sigma_i^2}\right), \quad (4.14)$$

where i is the number of observed peaks and $x = 10\log_{10}(\widehat{SNR}_{\text{sel}})$ with $\widehat{SNR}_{\text{sel}}$ denoting the estimated SNR at the output of the selection combiner (See Appendix). In (4.14), F_i , $i = 1, 2$ are weighting coefficients determined based on data fitting and are listed in Table 5.

4.4 BER Analysis and Measurements

In this section, we derive a closed-form expression for the BER and compare with measurements. Under the assumption of OOK signal, the conditional BER (Conditioned on the instantaneous SNR) is given by [68]

$$BER_c = Q\left(\sqrt{10^{x/10}}\right). \quad (4.15)$$

Taking an expectation with respect to the PDF in (4.14), we obtain the average BER as

$$BER = \sum_{i=1}^2 \frac{F_i}{\sqrt{2\pi\sigma_i^2}} \int_{-\infty}^{\infty} Q\left(\sqrt{10^{x/10}}\right) \exp\left(-\frac{(\mu_i - x)^2}{2\sigma_i^2}\right) dx. \quad (4.16)$$

Applying variable change $y_i = (x - \mu_i)/\sqrt{2}\sigma_i$, we could write (4.16) as

$$BER = \sum_{i=1}^2 \frac{F_i}{\sqrt{\pi}} \int_{-\infty}^{\infty} \exp(-y_i^2) Q\left(\sqrt{10^{\frac{(\sqrt{2}\sigma_i y_i + \mu_i)}{10}}}\right) dy_i. \quad (4.17)$$

Integration in (4.17) is in the form of $\int_{-\infty}^{\infty} g(y_i) f(y_i) dy_i$, where $g(y_i) = \exp(-y_i^2)$ is absolutely integrable function and $f(y_i)$ is a smooth-continuous function. Using Gauss-Hermite rule, we can express (4.17) in the form of a finite summation as [105, Eq. (25.4.46)]

$$BER \approx \sum_{i=1}^2 \sum_{k=1}^l \frac{F_i w_{k,l}}{\sqrt{\pi}} Q\left(\sqrt{10^{(\sqrt{2}\sigma_i y_{k,l} + \mu_i)/10}}\right), \quad (4.18)$$

where $y_{k,l}$, $k = 1, 2, \dots, l$ is set of roots defined by $H_l(x) = 0$ with $H_l(x)$ denoting the physicists Hermite polynomial. In (4.18), $w_{k,l}$, $k = 1, 2, \dots, l$ is the corresponding weight to $x_{k,l}$.

To further confirm the accuracy of truncated Gauss Hermite summation we investigate the required number of summed terms in our considered scenarios. Particularly, to further clarify the accuracy of this approximation, we compare exact BER through numerical integration and the derived BER for different order of approximation (l). As can be seen in Figure 32, almost perfect match is observed between the analytical results and the derived closed form expression by just using few terms. As expected, the number of needed terms (i.e., approximation order) to have a good degree of approximation is larger for lower BER compared to high BER values. However, number of required terms in all cases under consideration was $l = 15$ (i.e., less than $l = 20$ as reported in [106] for a good degree of approximation). This confirms the validity of our derived closed-form BER expression.

It can be therefore concluded that an order of $l = 15$ is enough for practical BER values under consideration. In the revised version, due to the limitation on the number of figures, we added a footnote to explain the accuracy of this approximation.

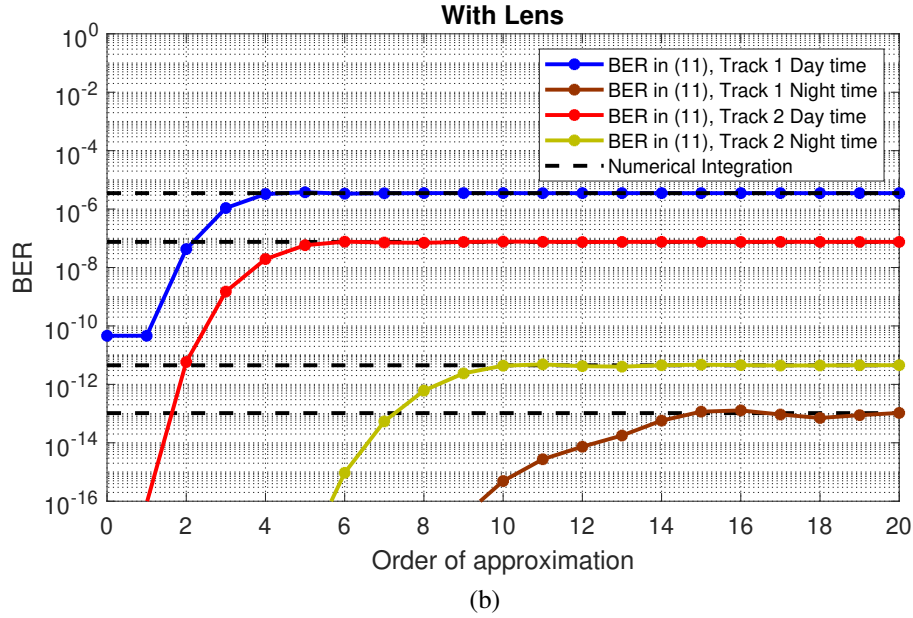
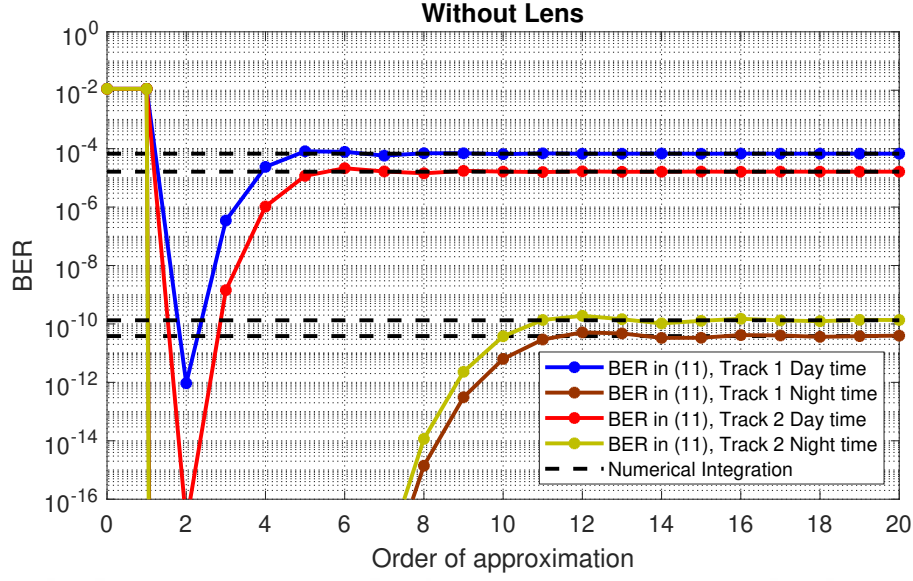
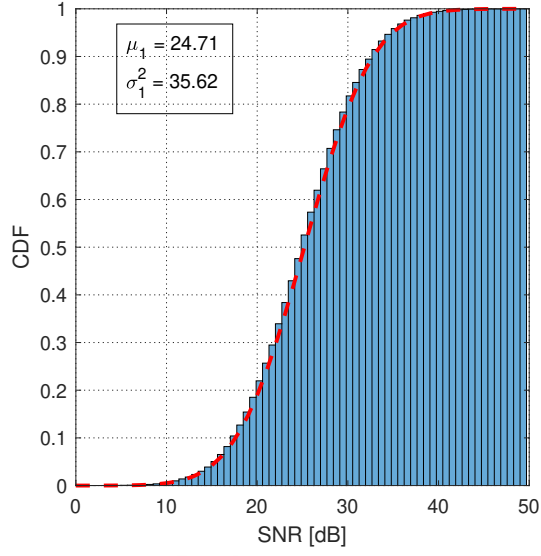
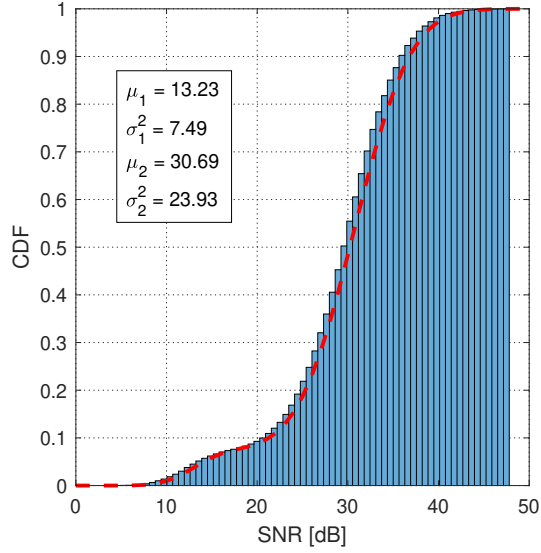


Figure 32: BER Comparison between numerical and closed-form expression with different orders.

Another point is confirming the accuracy of the proposed PDF. In Figure 33, we present the CDF for the experimental data and CDF of the weighted Gaussian PDF. Then For comparison purposes, we use the well-known Kolmogorov–Smirnov statistic [107] as the goodness-of-fit. This quantifies a vertical distance between approximate and exact CDFs. From Figure 33, it can be observed that these CDFs match very well and we find out that the vertical difference is small enough. This good match confirms our findings. In the revised manuscript, we included a footnote emphasizing that the



(a)



(b)

Figure 33: BER Comparison between numerical and closed-form expression with different orders.

approximated PDF accurately describes the histogram of the experimental data.

To verify the accuracy of the derived BER expression, we further conduct BER measurements. For each case, we transmit 1800 frames each of which contains a payload of 4000 bits (i.e., a total of 7.2×10^6 bits). The BER measurements are performed in real-time using LabVIEW by comparing the bits of each received frame with the actual transmitted bits. For comparison purposes, we also calculate the BER from (4.18) for

the measured time-average SNR values (See Table 5). As expected, there is some difference between analytical and measured BER values since the analytical expression in (4.18) does not take into account the practical effects such as channel estimation and frame detection. For a fixed BER value, the associated losses are found to be at most 1.1 dB.

4.5 Conclusion

A critical concern to establish a reliable VLC link between two moving vehicles is the number of photodetectors and their locations on the destination vehicle. This is needed to maintain link reliability in different road shapes and environmental conditions. In this chapter, we experimentally investigated the usage of dual photodetectors for a vehicular VLC system with selection combining to enhance signal reception in mobile conditions. We measured the electrical receive SNR for each photodetector as well as at the output of the selection combiner in two different tracks involving straight and curved roads. Through data fitting to experimental data, we obtained a PDF to describe the instantaneous SNR. Based on the PDF, we derived a closed-form expression for the average BER in the form of a finite summation and compared it with the measurements.

In conclusion, previous chapters ignored the effect of vehicle dynamics (vertical oscillation of the vehicles), which is expected to have a dramatic effect due to the asymmetrical nature of the vehicular light sources. In the next chapter, we experimentally model both the optical channel for asymmetrical headlight as well as the vertical oscillation of the vehicles to investigate the vertical oscillation effect on the vehicular VLC link performance.

CHAPTER V

VEHICULAR VLC LINKS IN THE PRESENCE OF OSCILLATION

5.1 *Introduction*

In this chapter, we focus on a special case of Truck-to-Truck VLC communication systems. Among all vehicles, heavy vehicles (i.e., Trucks) experience a huge vertical oscillation in mobility conditions. This oscillation is expected to have a huge impact on degrading the VLC link performance. This further motivates to mitigate this degrading effect in order to satisfy a specific QoS

In Section 5.2, we experimentally measure the optical channel for V2V considering a low-beam headlight and propose a closed-form path loss expression through data fitting. In Section 5.3, we illustrate the experimental set-up used for measuring the modelling of the truck's vertical oscillation. In Section 5.4, we obtain a closed-form BER expression for the truck-to-truck VLC system and investigate the effect of different system and channel parameters on the error rate performance. In Section 5.5, we present our numerical results. In an attempt to mitigate the degrading effects of the vertical oscillation, we further investigate the proper selection of photodetector location in order to minimize the error rate performance.

5.2 *Vehicular VLC Channel Modelling*

As in any communication system, accurate channel modelling is critical for designing and analyzing V2V communication systems. Different V2V optical VLC channel models are available through simulation methods [108–118], analytical works [119–121] or experimental measurements [122, 123]. In those models, either high beam [108, 109] or

low-beam [119–123] were considered as light sources. While high beam utilization allows further illumination/communication distances, there are regulations [124] banning the usage of high beam headlights in some cases [125]. These limitations motivate us to use low-beam headlights for vehicular VLC connectivity.

Literature has limited work for modelling vehicular optical channel using low-beam [119–123]. [119–121] presents analytical efforts through Lambertian channel model. Due to the asymmetrical pattern of the headlight, Lambertian channel model could not be justified for vehicular VLC channel. In [122], with low-beam headlight, measurements were taken and then simplified to Lambertian model. In [123], a model based on measurements of low-beam tungsten-halogen bulbs was introduced. Since LED-based headlight patterns are different than those of the traditional halogen light, this model is somewhat unrealistic. For benchmarking, we compared our measurements with the available low-beam models listed in Table 6.

In this section, we considered the use of dual headlights of the vehicle. We further considered the relative height between both headlights of the transmit vehicle and the photodetector of the receive vehicle. Let z_1 and z_2 respectively represent the height of the low-beam headlights (i.e., transmitters) of the follower vehicle and the photodetector (i.e., receiver) of the lead vehicle. Thus, the vertical displacement between the optical frontends of transmitter and receiver vehicles is $z = z_1 - z_2$. The headlights of the transmitter vehicle are separated from each other by a distance of d_s . The generated white light from the j^{th} headlight, $j = 1, 2$, propagates through the optical channel. The illumination pattern of the utilized low-beam headlamp was measured using a Goniophotometer [127] and provided in Figure 34. It is obvious from this figure that it has a very narrow beamwidth compared to the conventional Lambertian pattern. At the receiver side, the receiver effective area's diameter (i.e., D_R) is controlled with an iris diaphragm followed by a power meter [128] located at the back of the receiver vehicle.

We conducted a measurement campaign to measure the received optical power and then obtain the path loss through normalizing the measured values by the transmit

Table 6: Comparison of the available vehicular channel models

Ref.	Proposed Model	Modelling Approach	Source Type	KPI	Observation
[108]	A model based on ray-tracing simulator	Zemax Simulator	High beam	Maximum achievable distance	The model did not consider lateral distances
[109]	A model based on ray-tracing simulator	Zemax Simulator	High beam	Maximum achievable distance	Realistic model where distance, lateral shift, weather condition and received aperture are considered
[119, 120]	Lambertian	Analytical/MATLAB	Low-beam	Channel impulse response	Lambertian channel model could not be justified for asymmetrical pattern of the vehicular LED headlight
[121]	Lambertian	Analytical/MATLAB	Low-beam	Received optical power versus distance	Lambertian channel model could not be justified for asymmetrical pattern of the vehicular LED headlight
[122]	Lambertian	Experiment	Low-beam	Optical radiation pattern	Lambertian channel model could not be justified for asymmetrical pattern of the vehicular LED headlight
[123]	A model based on the measured headlamp beam pattern	Analytical/MATLAB	Low-beam	Received optical power versus distance	Measurements of low-beam pattern were for tungsten-halogen bulb [126] which has different pattern than vehicular LED modules. Besides, VLC is not supported by such light bulbs

power. During the measurements, we consider different inter-vehicles distances starting from $d = 10$ m up to $d = 50$ m with 5 m intervals. Different lateral shifts up to

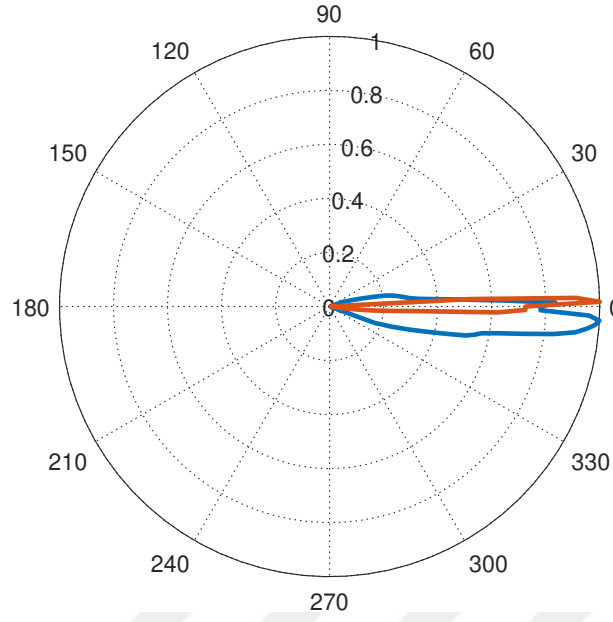


Figure 34: The measured illumination pattern of the utilized low-beam headlamp.

Table 7: Experimental set-up parameters for optical channel modelling.

Average electrical transmit power	P_{te}	28 W
Electrical to optical conversion efficiency	η	0.9
Inter-vehicle distance	d	10, 15, ... 50 m
Lateral distance	d_h	0, 1, ... 2 m
Measurement's center wavelength (LED peak's wavelength)	λ	450 nm
Power meter effective area's diameter	D_R	6, 8, ... 10 mm
Power meter responsivity (at 450 nm)	ρ	2.5 mA/W
Vertical displacement	Z	-0.5, -0.4, ... 0.5 m

$d_h = 2$ m are also considered. We further consider possible vertical displacements in the range of $-0.5 \text{ m} \leq z \leq 0.5 \text{ m}$. The positive sign of z indicates that the photodetector is higher than the headlamps while the negative sign indicates that the photodetector is lower than the headlamps. The experimental parameters are summarized in Table 7.

The measurements were carried out at night time to ensure accurate optical power measurements. Based on data fitting to the experimental measurements, the path loss can be described as

$$PL [\text{dB}] \approx 10 \log_{10} \left(\underbrace{\left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right) \exp \left(- (20(1 - \text{sign}(z)) + 10) \left(\frac{|z|}{d} \right) \right)}_h \right). \quad (5.1)$$

where $d_j = ((2j-3)d_s/2 + d_h)$ and h represents the channel coefficient between transmitter and receiver.

As benchmarks, we consider different path loss models proposed earlier in the literature:

1) Path loss model for low-beam headlight based on the assumption of Lambertian profile: The works in [129–135] considered Lambertian model for low-beam headlight. Noting that we have two headlights acting as wireless transmitters, we can write this path loss formula as

$$PL_{\text{Lambertian}} [\text{dB}] = 10 \log_{10} \left(\frac{1}{2} \sum_{j=1}^2 \left(\frac{(\pi D_R^2/4)(m+1)}{2\pi d^2} \cos^m(\theta_j) \cos(\phi_j) \right) \right). \quad (5.2)$$

where $m = -1/\log_2(\cos(\theta_{1/2}))$ is the Lambertian emission order with $\theta_{1/2}$ denoting the half-angle of the optical transmitter. In (5.2), θ_j and ϕ_j , respectively, denotes the irradiance and incident angle between the j^{th} headlamp and the photodetector.

2) Path loss formula for high beam headlight based on the assumption of asymmetrical pattern: In [136], a path loss model was proposed for high beam headlights through ray tracing for different weather conditions. It is given by

$$PL [\text{dB}] = 10 \log_{10} \left(\frac{1}{2} \sum_{j=1}^2 \left(\frac{D_R (d/\sqrt{d^2+d_{h,j}^2})^{1/\varepsilon}}{\zeta \sqrt{d^2+d_{h,j}^2}} \right)^2 \dots \exp \left(-c \sqrt{d^2+d_h^2} \left(\frac{D_R}{\zeta \sqrt{d^2+d_{h,j}^2}} \right)^{\varepsilon/2} \right) \right) \quad (5.3)$$

where $\zeta = 0.1585 \text{ rad}$ and $\varepsilon = 0.0175$ are corrections coefficients for clear weather [136]. In (5.3), $d_{h,j} = d_h \pm d_s/2$ and c is the extinction coefficient.

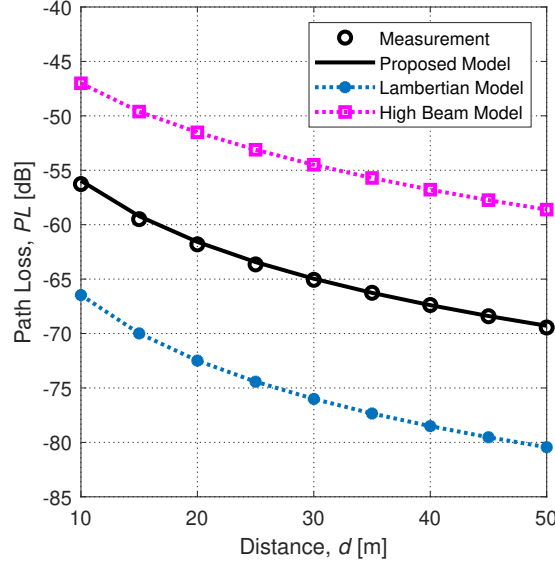


Figure 35: Comparison of the proposed path loss expression and other benchmarks.

The comparison of our measurement results with the first benchmark will illustrate the impact of asymmetrical radiation pattern of the vehicular headlights on the path loss compared to the Lambertian profile. Comparing with the second benchmark will elaborate the different characteristics of low and high beam headlights.

In Figure 35, we present our path loss model in (5.1) versus inter-vehicle distance along with the measurements and the above benchmarks. We consider perfect alignment between vehicles in terms of both lateral and vertical shifts (i.e., $d_h = 0$ m and $z = 0$ m). We further assume a receiver aperture diameter of $D_R = 10$ mm. It is observed that our proposed path loss expression in (5.1) maintains a good fit with measurements. The first benchmark in (5.2) building on Lambertian model underestimates the received power and lower power values are achieved as a result of Lambertian unlike the more focused pattern of low-beam headlights. On the other hand, the formula in (5.3) developed for high beam highlight shows that much higher received powers are available as expected. This confirms the further illumination/communication distances that could be achieved through employing high beam headlights compared to low-beam headlights.

In Figure 36, we investigate the effect of vertical displacement between the headlamps and the photodetector. The highest value of channel coefficient is obtained when the photodetector's height is exactly equal to the height of the headlamp, i.e., $z = 0$ m. In

the presence of vertical shifts, whether headlamps are higher or lower than the photodetector, the value of the channel coefficient decreases. In particular, if the photodetector is higher than the headlamps, the deterioration is more significant. For example, at an inter-vehicular distance of $d = 10$ m, the channel coefficient is -56.05 dB for $z = 0$ m. This value decreases to -64.51 dB and -58.18 dB, respectively, if the photodetector is higher and lower than the headlamps by the same absolute vertical distance of 40 cm (i.e., $z = \pm 0.4$ m) indicating a percentage change of 15.91 % and 3.80 %.

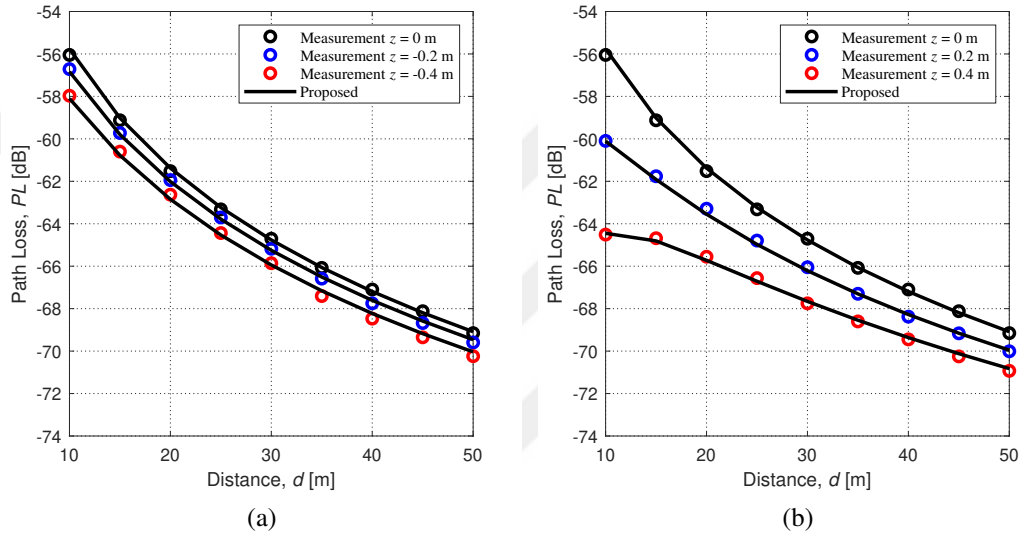


Figure 36: Comparison of the proposed path loss expression for vehicle's low-beam headlamps with measured data ($D_R = 10$ mm and $d_h = 0$ m are assumed): (a) Photodetector is lower than headlights and (b) photodetector is higher than headlights.

In Figure 37, we investigate the effect of the receiver aperture diameter. We assume perfect alignment in both vertical and lateral displacements (i.e., $z = 0$ m and $d_h = 0$ m). It is observed that the channel coefficient value decreases as the receiver aperture diameter decreases. For instance, assuming a receiver aperture diameter of $D_R = 8$ mm, the channel coefficient is measured as -57.26 dB, -66.06 dB and -70.51 dB, respectively, for $d = 10$ m, $d = 30$ m and $d = 50$ m. These drop, respectively, to -60.04 dB, -68.7 dB and -73.08 dB for $D_R = 6$ mm indicating relative change percentage of 4.86 %, 3.99 % and 3.64 %.

In Figure 38, we investigate the effect of lateral displacement between the vehicles. The highest value of channel coefficient is obtained for the perfect alignment case. In

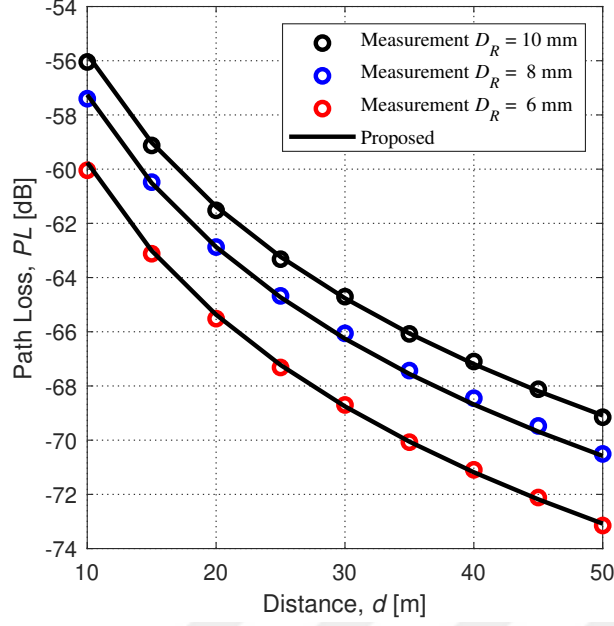


Figure 37: Comparison of the proposed path loss expression for vehicle's low-beam headlamps with measured data for different receiver aperture diameter ($z = 0$ m and $d_h = 0$ m are assumed).

the presence of lateral shifts, the channel coefficient decreases. For instance, at inter-vehicular distance of $d = 10$ m, $d = 30$ m and $d = 50$ m, respectively, the measured channel coefficients are -56.05 dB, -64.71 dB and -69.15 dB assuming zero lateral shift (i.e., $d_h = 0$ m). The channel coefficients drop, respectively, to -56.36 dB, -65.06 dB and -69.45 dB for $d_h = 1$ m and further drop to -57.69 dB, -65.65 dB and -69.67 dB for $d_h = 2$ m. It can be further seen that the effect of lateral displacement decreases as the inter-vehicular distance increases. For example, the percentage change in channel coefficient when the lateral displacement is 2 meters is 0.75 % for $d = 50$ m. This percentage increases to 2.93 % for $d = 10$ m. This occurs as the light is more collimated at short distances and any change in the lateral shift between vehicles has a higher impact compared to longer distances at which the light is already spread.

5.3 Vehicular Vertical Oscillation Modelling

In the presence of mobility, trucks suffer from vertical oscillation [137]. This oscillation depends on many factors such as road type, vehicle speed and vehicle size [138].

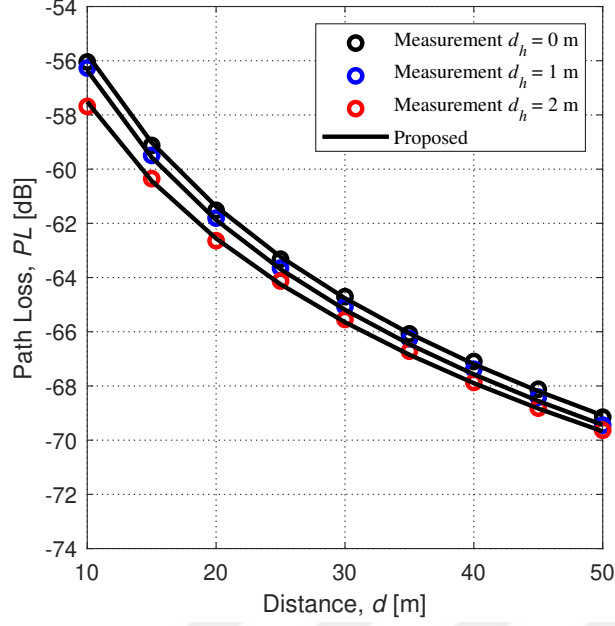
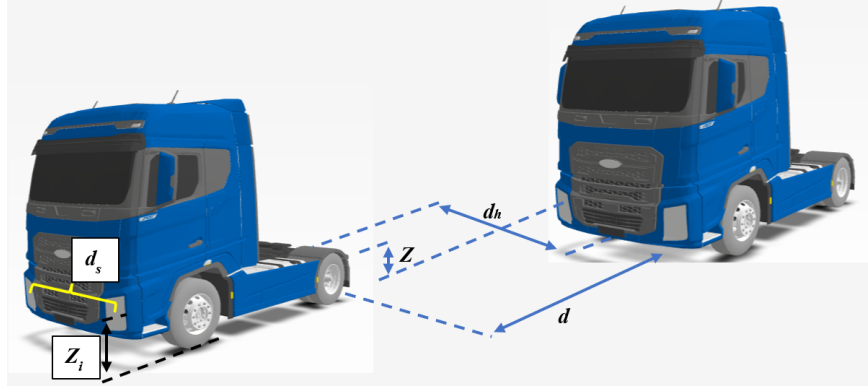


Figure 38: Comparison between the proposed path loss expression for vehicle's low-beam headlamps and the measured data for different lateral distances ($D_R = 10$ mm and $z = 0$ m are assumed).

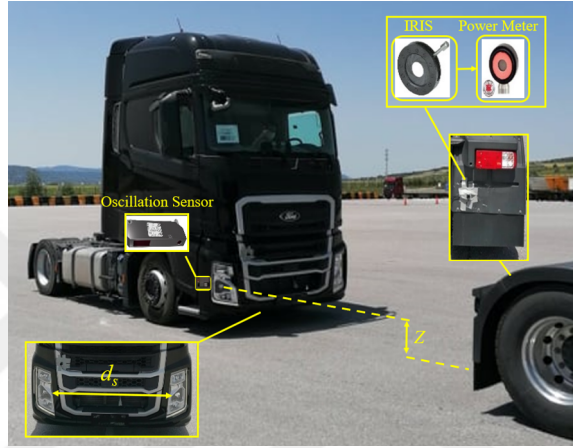
Among vehicles, heavy vehicles such as trucks were found to produce the most perceptible oscillation [139, 140]. We have conducted a measurement campaign to model the truck's vertical oscillation. A Waycon laser displacement transducer [141] was placed at the same level of the headlamp of Ford FMAX truck (See Figure 39). This transducer reads the vertical distance to the road pavement in real time manner. During this measurement campaign, the vertical displacement was measured on a smooth road considering different speeds. In order to model the vertical oscillation of a truck at a specific speed, the vehicle accelerates until reaching the desired speed. After reaching the desired speed, we wait a few seconds before starting the measurement to make sure that the oscillation due to the acceleration is not included in any of the recorded samples. In our measurements, we considered speeds from $v = 30$ km/h up to $v = 90$ km/h.

As illustrated in Figure 40, the histogram of the measured vertical displacement can be well-described by Gaussian distribution for different speeds under consideration. The PDF of the vertical oscillation of the i^{th} truck can be therefore expressed as

$$f_{z_i}(z_i) = \frac{1}{\sqrt{2\pi\sigma_i^2}} \exp\left(-\frac{(z_i - \mu_i)^2}{2\sigma_i^2}\right), \quad i \in \{1, 2\}, \quad (5.4)$$



(a)



(b)

Figure 39: (a) V2V system under consideration (b) experimental set-up.

where σ_i^2 is the variance of vertical oscillation. In (5.4), μ_i is the mean of the vertical oscillation, which represents the distance from the optical frontend (i.e., headlamps or photodetector) to the road surface. Vertical oscillation leads to a change in the photodetector/headlamps' heights. Therefore, to model its effect on the communication performance, the distribution of $z = z_1 - z_2$ is needed. Utilizing the fact that the distribution of the difference between two Gaussian random variables is also Gaussian [142], the PDF of vertical oscillation could be written as

$$f_z(z) = \frac{1}{\sqrt{2\pi\sigma_V^2}} \exp\left(-\frac{(z-\mu)^2}{2\sigma_V^2}\right), \quad (5.5)$$

where the mean and variance of the relative vertical displacement are, respectively, $\mu = \mu_1 - \mu_2$ and $\sigma_V^2 = \sigma_1^2 + \sigma_2^2$. In (5.5), μ also represents the relative height between the light sources of the transmit truck and the photodetector of the receive truck (i.e., the photodetector's vertical location relative to the headlamps). Positive value of μ means

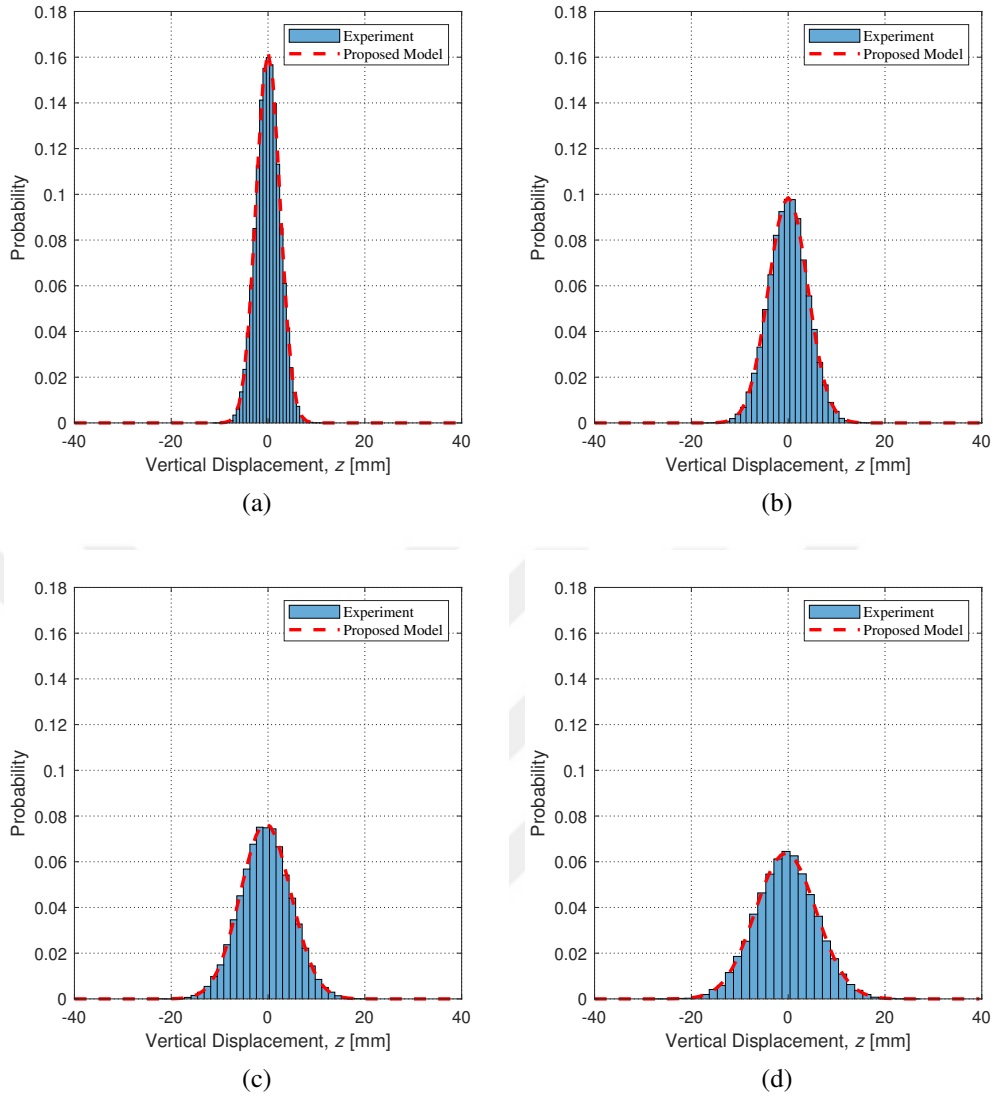


Figure 40: Measured vertical displacement of Ford FMAX truck at: (a) 30 km/h, (b) 50 km/h, (c) 70 km/h and (d) 90 km/h.

that the photodetector is higher than the headlights while the negative values indicates that it has lower level compared to the headlight.

It is observed from Figure 41 that the variance of the vertical oscillation (i.e., σ_i^2) increases linearly with the speed. Such observation is consistent with the study in [143]. For example, consider a truck speed of $v = 30$ km/h, the variance of vertical oscillation is found as $\sigma_i^2 = 6.25$. This increases to $\sigma_i^2 = 17.2$, $\sigma_i^2 = 27.3$ and $\sigma_i^2 = 39.1$, respectively, for $v = 50$ km/h, $v = 70$ km/h and $v = 90$ km/h. Using curve fitting [104] to the measurement data, the relation between the variance of the vertical oscillation (σ_i^2) and

truck's speed (v) can be expressed as

$$\sigma_i^2(v) \approx 0.55v - 11.45, \quad 30 \leq v \leq 90. \quad (5.6)$$

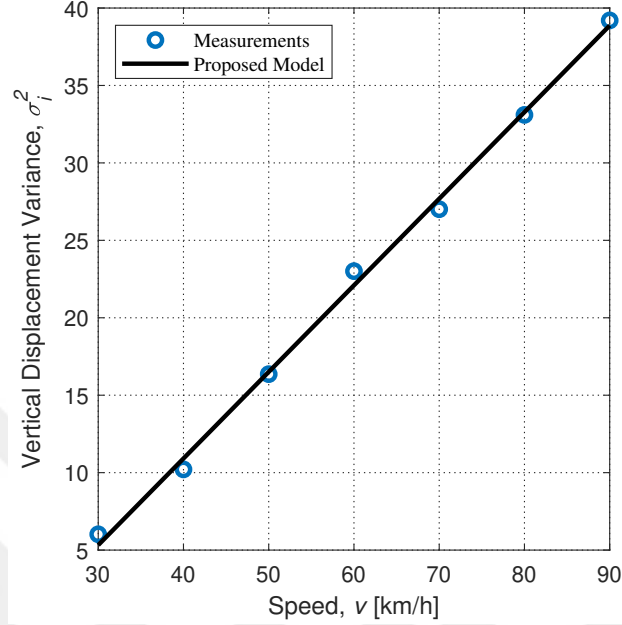


Figure 41: Relation between the vertical oscillation variance and the truck speed.

5.4 Error Rate Performance

We presented a closed-form expression for the path loss in Section 5.2 and a statistical model for vertical oscillation in Section 5.3. In this section, based on these new expressions, we derive a BER expression for truck-to-truck VLC communication system considering Intensity Modulation direct-detection (IM/DD). Let x denote the transmitted unipolar M -ary pulse amplitude modulation (M -PAM) symbol. The received signal can be expressed as

$$y = \sqrt{P_{te}} \eta h r G_{PD} x + w, \quad (5.7)$$

where P_{te} , η , r , and G_{PD} are, respectively, average electrical transmit power, electro-optical conversion efficiency of the headlight, opto-electrical responsivity of the photodetector and transimpedance gain. In (5.7), w denotes Additive White Gaussian Noise (AWGN) term with zero mean and noise variance of σ_n^2 , i.e., $w \sim \mathcal{N}(0, \sigma_n^2)$.

The conditional BER (conditioned on the statistical distribution of the vertical oscillation denoted by z) could be written in the form of (5.8) [144]

$$\begin{aligned} BER(z) &= F \operatorname{erfc} \left(\sqrt{C\gamma} h \right) \\ &= F \operatorname{erfc} \left(\sqrt{C\gamma} \sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \exp \left(-(20(1 - \operatorname{sign}(z)) + 10) \left(\frac{|z|}{d} \right) \right) \right). \end{aligned} \quad (5.8)$$

In (5.8), $\operatorname{erfc}(\cdot)$ is the complementary error function, $F = (M-1)/(M \log_2(M))$, $C = 3/(2(M-1)(2M-1))$, and $\gamma = P_{\text{te}} \eta^2 r^2 G_{PD}^2 / \sigma_n^2$ is the transmit SNR.

By averaging the conditional BER over the PDF of the vertical oscillation in (5.5), we obtain the average BER as

$$\begin{aligned} \overline{BER} &= \frac{F}{\sqrt{2\pi\sigma_V^2}} \int_{-\infty}^{\infty} \exp \left(-\frac{(z-\mu)^2}{2\sigma_V^2} \right) \dots \\ &\operatorname{erfc} \left(\sqrt{C\gamma} \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right) \exp \left(-(20(1 - \operatorname{sign}(z)) + 10) \left(\frac{|z|}{d} \right) \right) \right) dz. \end{aligned} \quad (5.9)$$

The integration in (5.9) can be re-written as

$$\begin{aligned} \overline{BER} &= \frac{F}{\sqrt{2\pi\sigma_V^2}} \sum_{k=1}^2 \int_0^{\infty} \exp \left(-\frac{(z-\mu_k)^2}{2\sigma_V^2} \right) \dots \\ &\operatorname{erfc} \left(\sqrt{C\gamma} \exp \left(\frac{40k-90}{d} z \right) \sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right) dz. \end{aligned} \quad (5.10)$$

In (5.10), $\mu_k = (2k-3)\mu$. Using the approximation of $\operatorname{erfc}(x) \approx a_1 \exp(b_1 x^2) + a_2 \exp(b_2 x^2)$ [145] where $a_1 = 0.3017$; $a_2 = 0.4389$; $b_1 = -1.0510$ and $b_2 = -2.1020$, (5.10) can be approximated as

$$\begin{aligned} \overline{BER} &\approx \sum_{k=1}^2 \sum_{l=1}^2 \frac{a_l F}{\sqrt{2\pi\sigma_V^2}} \int_0^{\infty} \exp \left(-\frac{(z-\mu_k)^2}{2\sigma_V^2} \right) \dots \\ &\exp \left(b_l \left(C\gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \exp \left(2z \frac{40k-90}{d} \right) \right) \right) dz. \end{aligned} \quad (5.11)$$

Using the variable change of $x = (z - \mu_k) / \sqrt{2\sigma_V^2}$, we have

$$\overline{BER} \approx \sum_{k=1}^2 \sum_{l=1}^2 \frac{a_l F}{\sqrt{\pi}} \int_0^{\infty} \exp(\xi_{k,l}(x)) dx, \quad (5.12)$$

where $\xi_{k,l}(x)$ in (5.12) is given as

$$\xi_{k,l}(x) = b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \dots \exp \left(2\mu_k \frac{40k-90}{d} \right) \exp \left(2\sqrt{2\sigma_V^2} x \frac{40k-90}{d} \right) - x^2. \quad (5.13)$$

Utilizing Laplace method of integration [146], (5.12) is obtained as

$$\overline{BER} \approx \sum_{k=1}^2 \sum_{l=1}^2 \frac{a_l F \sqrt{2}}{\sqrt{|\xi''_{k,l}(x_{\circ,k,l})|}} \exp(\xi_{k,l}(x_{\circ,k,l})), \quad (5.14)$$

where $\xi''_{k,l}(x)$ is the second derivative of $\xi_{k,l}(x_{\circ,k,l})$ and can be expressed

$$\xi''_{k,l}(x) = 8 \left(\frac{40k-90}{d} \right)^2 \sigma_V^2 b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \dots \exp \left(2\mu_k \frac{40k-90}{d} \right) \exp \left(2\sqrt{2\sigma_V^2} x \frac{40k-90}{d} \right) - 2. \quad (5.15)$$

$x_{\circ,k,l}$ in (14) is the solution of $\xi'_{k,l}(x) = 0$ with $\xi'_{k,l}(x)$ denoting first derivative of $\xi_{k,l}(x)$ and can be readily found as

$$x_{\circ,k,l} = -\frac{1}{2C_k \sqrt{2\sigma_V^2}} \dots W \left(-4 \left(\frac{40k-90}{d} \right)^2 \sigma_V^2 b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \exp \left(2\mu_k \frac{40k-90}{d} \right) \right). \quad (5.16)$$

In (5.16), $W(\cdot)$ is the Lambert W function [147]. Replacing (5.13), (5.15) and (5.16) within (5.14), the final BER expression can be written as

$$\begin{aligned} \overline{BER} \approx & \sum_{k=1}^2 \sum_{l=1}^2 a_l F \sqrt{2} \exp \left(-\frac{1}{8C_k^2 \sigma_V^2} W^2 \left(-4C_k^2 \sigma_V^2 b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \exp \left(2\mu_k \frac{40k-90}{d} \right) \right) \right) \\ & \exp \left(b_l C \gamma A^2 \exp(2C_k \mu_k) \exp \left(-W \left(-4C_k^2 \sigma_V^2 b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \exp \left(2\mu_k \frac{40k-90}{d} \right) \right) \right) \right) \\ & \times \frac{1}{\sqrt{\left| 8C_k^2 \sigma_V^2 b_l C \gamma A^2 \exp(2C_k \mu_k) \exp \left(-W \left(-4C_k^2 \sigma_V^2 b_l C \gamma \left(\sum_{j=1}^2 \left(\frac{4D_R}{3(d+d_j^2)} \right)^2 \right)^2 \exp \left(2\mu_k \frac{40k-90}{d} \right) \right) \right) - 2 \right|}}. \end{aligned} \quad (5.17)$$

5.5 Results and Discussion

In this section, we present the numerical results and validate the preciseness of derived BER expression. Unless otherwise stated, we assume modulation order of $M = 4$, noise

variance of $\sigma_n^2 = -63\text{dB}$ [101], photodetector aperture diameter of $D_R = 10\text{ mm}$, photodetector responsivity of $r = 0.9\text{ A/W}$, transimpedance gain of $G_{PD} = 33.49\text{ dB}$ [148], lateral displacement of $d_h = 0\text{ m}$, average electrical transmit power of $P_{te} = 28\text{ W}$ and electro-optical conversion efficiency of $\eta = 0.9$. We further consider a minimum intervehicular distance of 10 m for safety measures [149] and truck speed of $v = 50\text{ km/h}$.

In Figure 42, we assume both trucks are exactly behind each other without lateral shift (i.e., $d_h = 0\text{ m}$). We further assume that the height of the photodetector is exactly equal to the height of the headlamps (i.e., $z = 0\text{ m}$) and plot the derived approximate BER expression in (5.17) along with the exact expression in (5.9) in which the integration was numerically calculated. An ideal case where the two trucks follow each other without any vertical oscillation is also included as a benchmark. It is observed that the derived BER expression provides excellent match with numerical calculations confirming the preciseness of our derivation. As expected, the best performance is obtained in the absence of the vertical oscillation. For example, at $d = 20\text{ m}$, the BER of the ideal case is 4.66×10^{-6} and reduces to 2.48×10^{-3} in the presence of vertical oscillation.

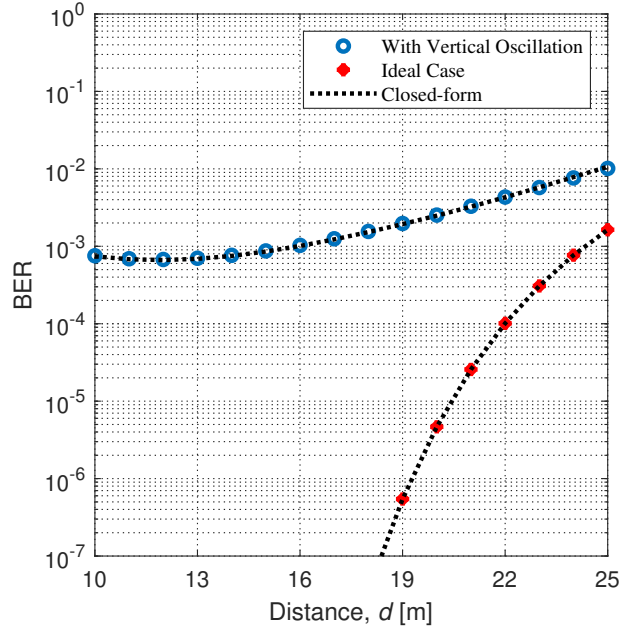


Figure 42: Numerical BER and closed-form for $\mu = 0\text{ m}$ ($D_R = 10\text{ mm}$, $d_h = 0\text{ m}$ and $v = 50\text{ km/h}$ are assumed).

In the following, we investigate the effects of relative height between the frontends

(denoted by μ), lateral shift between the trucks and truck speed on the system performance. We further optimize the photodetector location and investigate the resulting enhancement in error rate performance.

Effect of relative height: In Figure 43, we investigate the effect of relative height between the light sources of the transmit truck and the photodetector of the receive truck on the BER performance. We assume both trucks are exactly behind each other (i.e., $d_h = 0$ m). We consider relative height of $\mu = \pm 0.1$ m and present BER for $v = 30$ km/h and $v = 50$ km/h, respectively, in Figure 43.a and 43.b. The case of $\mu = 0$ m is also included as a benchmark. In the absence of vertical oscillation, the highest value of the channel coefficient is obtained when both the transmitter and the receiver are at the same height (i.e., $\mu = 0$ m) whereas, in the presence of vertical oscillation, the performance at $\mu = 0$ m becomes no longer the best. In particular, the performance gets better for negative values of μ (i.e., photodetector is lower than the headlights). For example, at $d = 10$ m and $v = 30$ km/h, the BER is 9.15×10^{-6} for $\mu = 0$ m. This increases to 7.54×10^{-4} for $\mu = 0.1$ m. On the other hand, for $\mu = -0.1$ m, BER decreases to 2.77×10^{-7} . It can be further observed that the BER changes with the truck's speed. For example, at $v = 50$ km/h (see Figure 43.b), the BER for $\mu = 0.1$ m, $\mu = 0$ m and $\mu = -0.1$ m become, respectively, 3.02×10^{-3} , 7.54×10^{-4} and 1.43×10^{-4} at $d = 10$ m. This is due to the fact that at higher speed, the vertical oscillation variance is higher compared to lower speed. This changes the average received optical power at different relative height.

Effect of lateral shift: In Figure 44, we investigate the effect of lateral shift between transmit and receive trucks on the BER performance considering $d_h = 1$ m and $d_h = 2$ m. The case of $d_h = 0$ m is also included as a benchmark. It is observed that the presence of lateral shift between trucks degrades the BER performance. For instance, at $d = 10$ m, the BER is obtained as 7.54×10^{-4} assuming perfect alignment (i.e., $d_h = 0$ m). This increases to 9.88×10^{-4} and 2.38×10^{-3} , respectively, for $d_h = 1$ m and $d_h = 2$ m indicating relative change percentage of 31.03 % and 215.64 %. These percentages drop to 27.27 % and 204.87 % for $d = 20$ m. This indicates that as the inter-truck

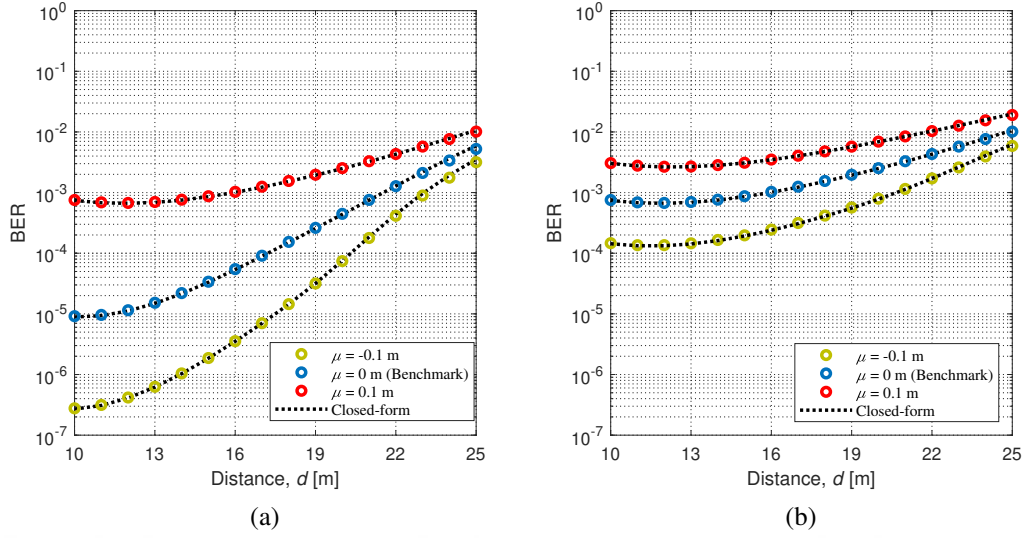


Figure 43: Effect of relative mean on error rate performance ($D_R = 10$ mm and $d_h = 0$ m are assumed) at: (a) $v = 30$ km/h and (b) $v = 50$ km/h.

distance increases, the effect of lateral shift on BER decreases because the vehicular light pattern is more collimated at short distances compared to longer distances.

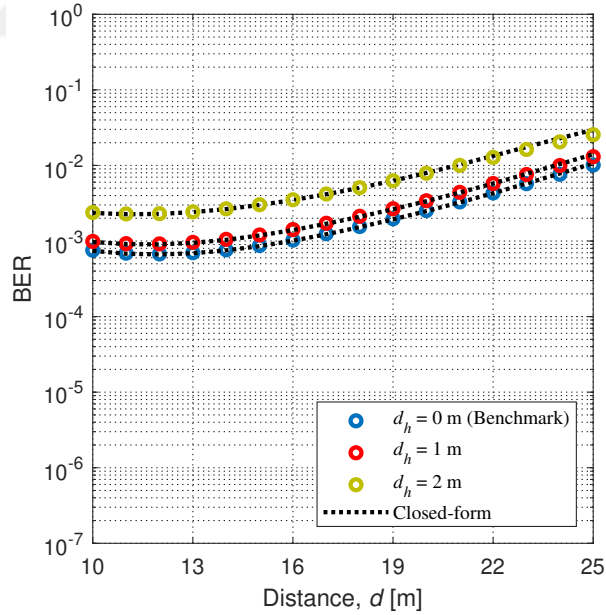


Figure 44: Effect of lateral shift on error rate performance ($D_R = 10$ mm, $\mu = 0$ m and $v = 50$ km/h are assumed).

Effect of speed: In Figure 45, we investigate the BER performance for different truck speeds. When speed increases, the vertical oscillation variance increases and

effectively degrading the BER. For example, at $d = 10$ m, BER is obtained as 1.73×10^{-6} for $v = 30$ km/h. This increases to 5.79×10^{-4} , 2.85×10^{-3} and 6.36×10^{-3} , respectively, for $v = 50$ km/h, $v = 70$ km/h and $v = 90$ km/h. Furthermore, it is noticed that the effect of speed on the error rate performance at longer distances has lower effect compared to that of shorter distance as the light is already spread. For example, BER values of 1.73×10^{-6} and 6.36×10^{-3} are observed at speeds of $v = 30$ km/h and $v = 90$ km/h, respectively, at $d = 10$ m. The BER values are found as 1.74×10^{-5} and 6.18×10^{-3} at $d = 15$ m and 9.76×10^{-4} and 1.11×10^{-2} at $d = 20$ m.

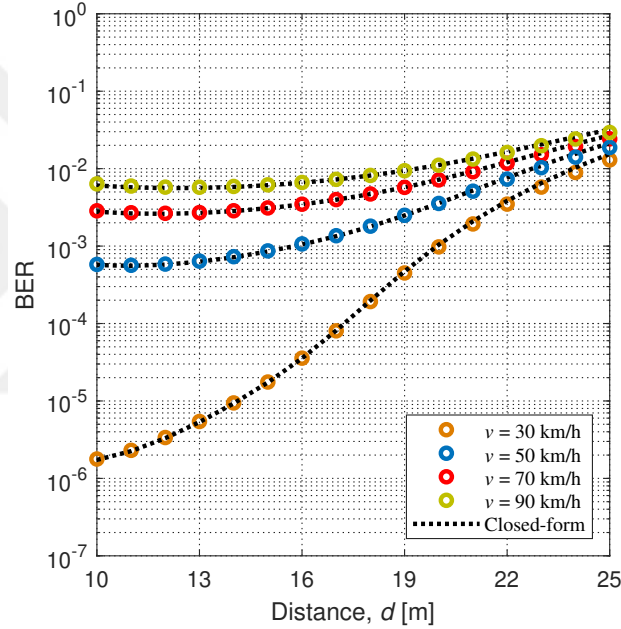


Figure 45: Effect of truck's speed on error rate performance ($D_R = 10$ mm, $d_h = 0$ m and $\mu = -0.1$ m are assumed).

Effect of photodetector location: The presence of vertical oscillation as a result of mobility changes the average received optical power. This motivates to locate the photodetector at a position where the highest average optical power is received in the presence of mobility. In Figure 46.a, at a fixed speed of $v = 50$ km/h, we plot the BER against different relative vertical location values. As discussed earlier, when μ is positive, the average error rate is high. Whereas it is observed that the average error rate reduces for negative μ values till a specific point then it starts to degrade. For instance, at $d = 10$ m, a minimum BER of 6.45×10^{-7} is observed at $\mu = -0.48$ m. As the

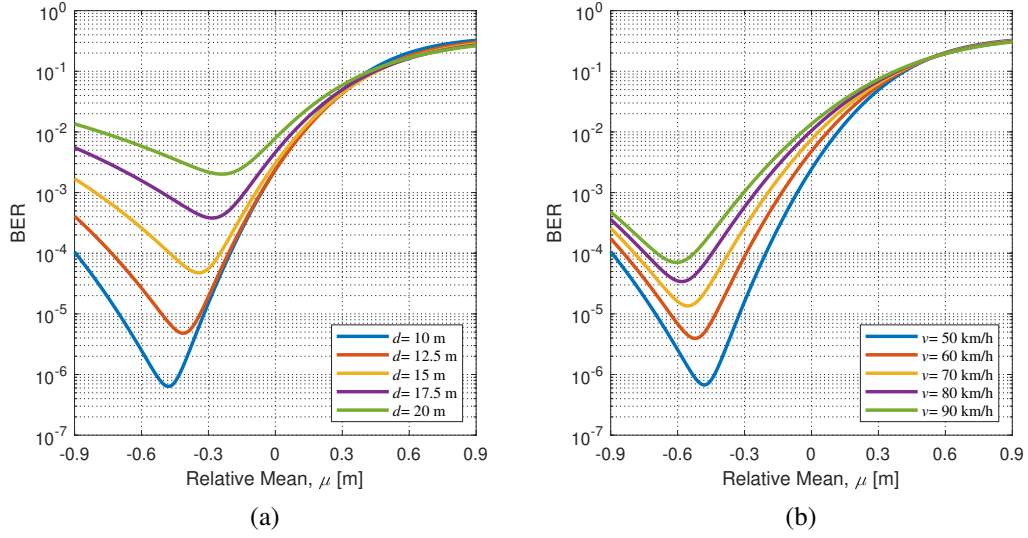


Figure 46: BER versus photodetector's relative vertical location for: (a) Different inter-trucks distances ($D_R = 10$ mm, $d_h = 0$ m and $v = 50$ km/h are assumed) and (b) different truck speed values ($d = 10$ m and $D_R = 10$ mm and $d_h = 0$ m are assumed).

distance increases, the minimum BER further decreases. For example, at $d = 15$ m and $d = 20$ m, minimum BER of 4.71×10^{-5} and 2.02×10^{-3} are observed at $\mu = -0.36$ m and $\mu = -0.22$ m. The BER performance starts to degrade after the minimum-BER point as any vertical displacement will cause reduction in the average received power.

To investigate the effect of the truck speed on selecting the photodetector vertical location, we assume a fixed inter-truck distance of $d = 10$ m and plot the achieved BER versus different relative vertical location values (see Figure 46.b). As the speed increases, the vertical oscillation variance increases and the vertical location at which the BER is minimized increases as well. For instance, at $v = 50$ km/h, a minimum BER of 6.45×10^{-7} is found at a photodetector vertical location of $\mu = -0.48$ m. This vertical location increases to $\mu = -0.56$ m and $\mu = -0.59$ m, respectively, for speed values of $v = 70$ km/h and $v = 90$ km/h with obtained minimum BER of 1.35×10^{-5} and 7.05×10^{-5} .

To find the optimal vertical location of the photodetector on the receive truck, we formulate an optimization problem that minimizes the BER. Under the constraints of $D_R \leq 10$ mm, $d_h \leq 2$ m and $|\mu| \leq \mu_{\max}$ ($\mu_{\max} = 0.95$ m is limited by the height of the

truck's headlight with respect to the ground), the optimal location could be obtained as

$$\mu_{Opt} = \arg \min_{\mu \in \Omega} \{ \overline{BER} \} , \quad (5.18)$$

where Ω is the set of μ values. Using MATLAB optimization toolbox [150], we find $\mu_{Opt} = -0.43$ m (i.e., below the headlights). In Figure 47, we present the BER performance for a photodetector located at the optimal location. It is observed that the performance becomes close to the performance of the ideal case (i.e., Trucks follow each other without any vertical oscillation).

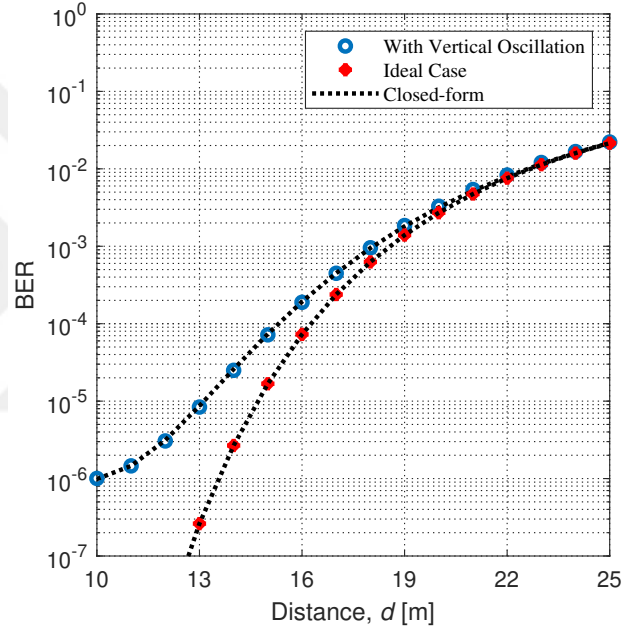


Figure 47: Error rate performance at the optimal vertical location on the receive vehicle of $\mu_{Opt} = -0.43$ m ($D_R = 10$ mm, $d_h = 0$ m and $v = 50$ km/h are assumed).

5.6 Conclusion

For the special case of truck-to-truck V2V system, we experimentally investigate the system VLC performance. In this chapter, we considered two heavy vehicles (e.g., trucks) on a road following each other. Low-beam headlights of the follower truck were acting as wireless transmitters and the lead truck was equipped with a photodetector at its back. We first proposed an accurate path loss model based on experimental measurements. The proposed model takes into account the presence of lateral shift between

the trucks as well as the fact that headlamps and photodetector can be placed at different heights. Furthermore, we modeled the truck vertical oscillation experimentally at different speeds. Utilizing these models, we derived a BER expression for VLC-based truck-to-truck communication system and investigated the effect of several system and channel parameters on the performance. In an attempt to mitigate the degrading effects of the vertical oscillation, we further investigated the proper selection of photodetector location in order to minimize the error rate performance.

Previous chapters investigated the performance of the vehicular VLC system in a real outdoor environment on mobility conditions. To further benchmark with the RF vehicular connectivity and have a robust V2X connectivity, a hybrid (VLC/RF) system needs to be implemented. In the next chapter, not only we practically implement a hybrid system but also we propose a switching algorithm to maintain the needed QoS.

CHAPTER VI

HYBRID VLC/RF SYSTEM

6.1 Introduction

In this chapter, we compare two potential candidates for vehicular connectivity (i.e., VLC and RF) in realistic driving scenarios. Furthermore, in order to satisfy a specific QoS, we propose a switching algorithm for hybrid vehicular VLC/RF systems. As a benchmark, we considered the deployment of some available algorithms of the litterateur. There are two main critical concerns for the experimental deployment of the switching algorithms: The first concern is to make both technologies (mmW and VLC) support almost similar data rates. Thus, for this concern, a new frontend design has been developed. The second concern is extracting the data rate from different both chipsets in a real-time manner which is achieved through API and Nating processes. After satisfying those concerns, the practical implementation of switching algorithms will be deployable.

The chapter is organized as follows: In Section 6.2, we validate the new custom-designed frontend used for VLC system to be able to support high data rates. Section 6.3 focuses on the measurement campaign of mmW links for vehicular connectivity. Section 6.4 Explains the experimental set-up and measurements for real-driving scenarios. In Section 6.5 we detail the proposed switching algorithm and benchmark with different algorithms. Finally, the chapter is concluded in In Section 6.6.

6.2 VLC frontend Design and Validation

A critical concern is maintaining similar illumination pattern as headlight. Thus, we re-designed the headlight's Printed Circuit Board (PCB) and placed the new LEDs in the same place behind the optics. The new designed PCB is illustrated in Figure 48.



Figure 48: The new designed PCB for the vehicular headlight.

Moreover, in the VLC frontend, to improve the performance, equalization is employed. The bandwidth comparison with and without equalizer is illustrated in Figure 49. It is observed that the frontend with equalization has a more stable response compared to the old design. Moreover, it provides higher signal levels. This new custom-designed frontend has a 10-dB bandwidth of 50 MHz compared to 1.7 MHz of the old design.

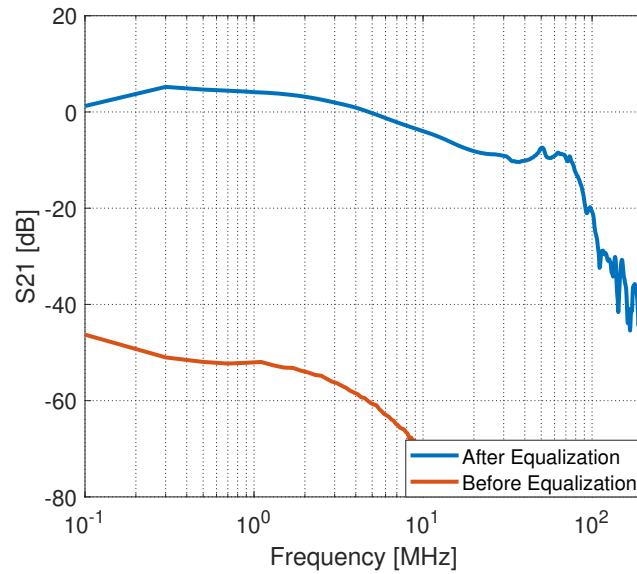


Figure 49: VLC system bandwidth before and after equalization.

6.3 mmW Measurement Campaign on Static Conditions

Initial measurements of the mmW channel/performance have been evaluated for the V2V scenario. The performance of two mmW transceiver modules (represents follower/leader vehicles) has been initially investigated considering different intervehicle distances, lateral distances between both modules. We considered the use of Siklu 600T modules as mmW transceivers. Each module was attached to a tripod that represents a vehicle. On static conditions, we measured the Received Signal Strength Indication (RSSI), Carrier to Interference Noise Ratio (CINR) as well as the datarate.

In this measurement campaign, we considered conducting the measurements in dens area. As a benchmark, we further considered the measurements of open area as well. The experimental environment is illustrated in Figure 50.



Figure 50: Experimental environment: (a) Dens area and (b) open area.

The measured mmW link performance of the dens environment for different lateral distances is illustrated in Figure 51. In all measurements, the interference noise is observed to have a huge impact on the mmW link performance. This fluctuating performance is due to the reflection caused by the surrounding objects such as buildings' walls, bicycles, ...etc. This might give a chance for VLC to outperform the mmW link in similar short distances inside the city where mmW will suffer from destructive reflections.

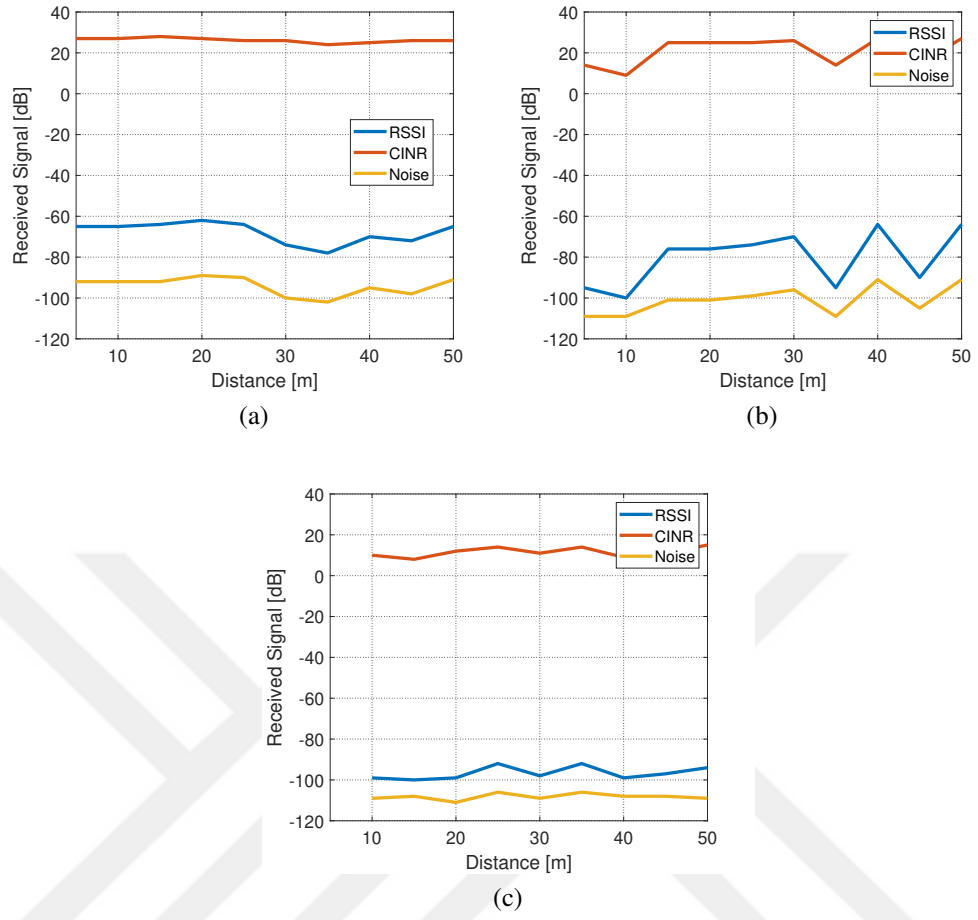


Figure 51: Received signal at dens area for different intervehicle distances of (a) 0 m (b) 1 m and (c) 2 m.

To confirm the degrading effect of mmW link reflection, we conducted another measurement campaign in an open area. The performance results this benchmark area is illustrated in Figure 52. Unlike the observed significant interference in the presence of the surrounding object, a steady performance is observed. This gives a clear benchmark for mmW measurements.

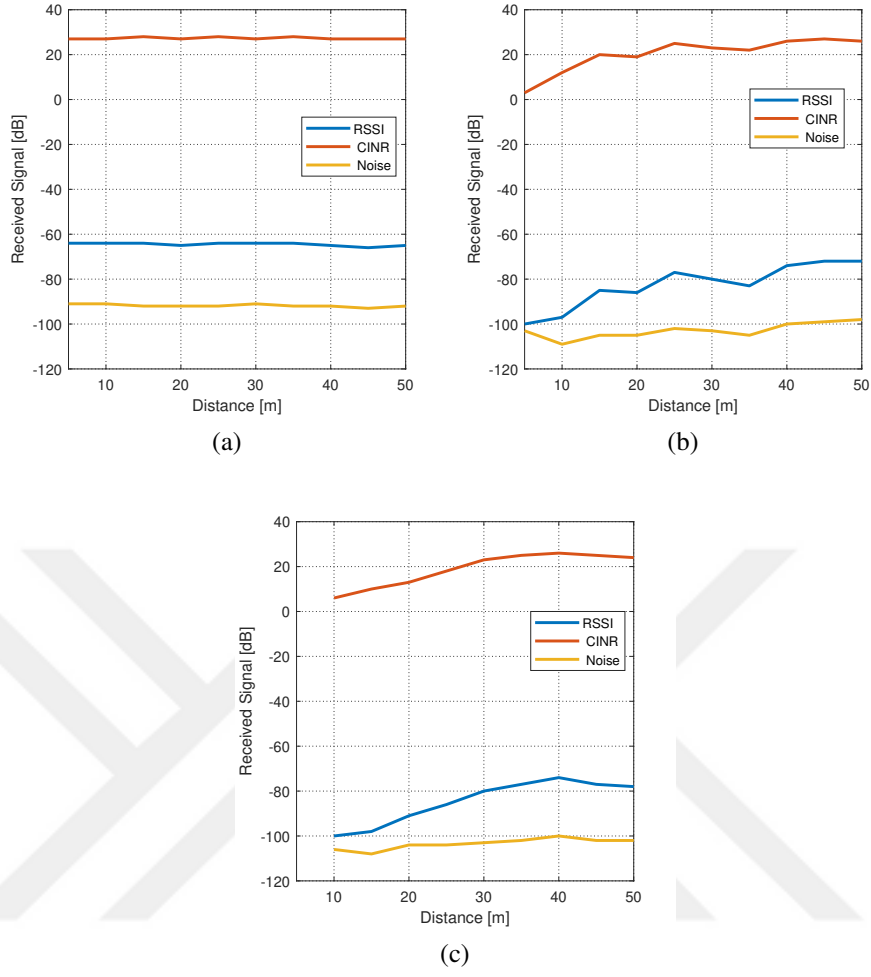


Figure 52: Received signal at open area for different intervehicle distances of (a) 0 m (b) 1 m and (c) 2 m.

A comparison in the achieved data rates of both campaigns is illustrated in Figure 53. A similar fluctuating performance in the data rates is observed in the dens area compared to a steady performance at the open area.

6.4 Measurement Results of the Hybrid VLC/RF on Mobility Conditions

In this section, we consider the hybrid use of VLC with RF counterparts. In particular, we employ mmW modules and compare with VLC in mobility conditions. An API is developed to extract the data rates of the mmW link (i.e., Siklu 600T modules) in a real-time manner.

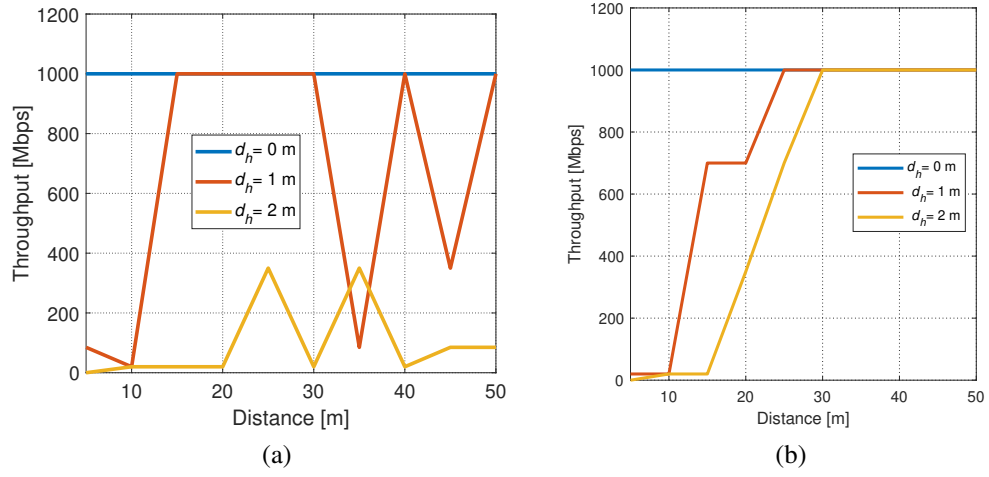


Figure 53: mmW link data rates at (a) Dens area and (b) open area.



Figure 54: Experimental set-up of hybrid system.

As illustrated in Figure 54, each vehicles is equipped with VLC and mmW transceivers. These vehicles, follower and leader, are following each others in the city road. We considered two types of road; main road (See Red highlighted roads in Figure 55) and side roads (Black highlighted roads). The side roads has less surrounding vehicles while the main road has nearby vehicles (See Figure 56).

For a fair comparison between mmW and VLC systems, due to the difference in the operating bandwidths (i.e., mmW module has a 500 MHz bandwidth compared to 50

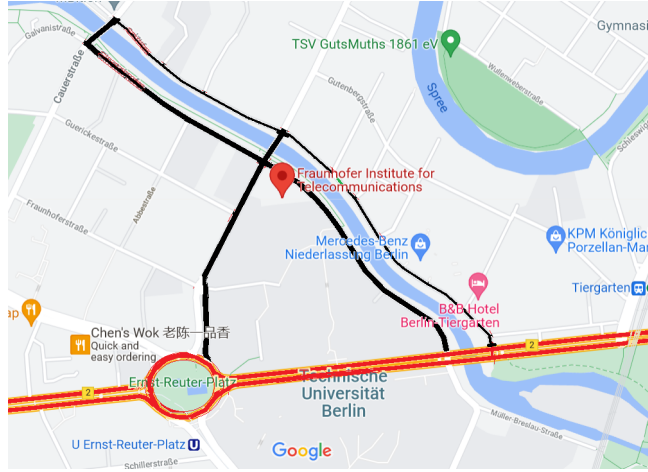


Figure 55: Under-consideration Roads.



(a)

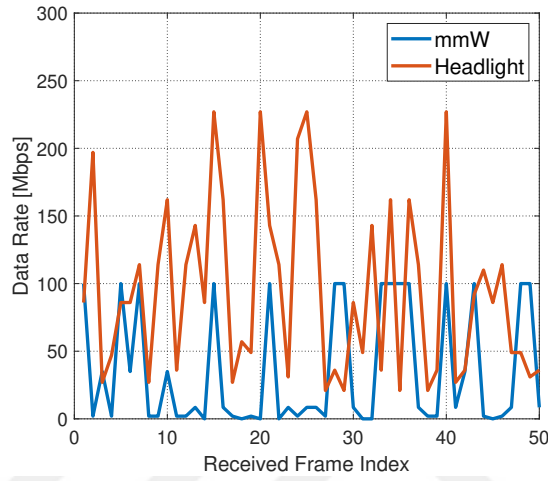


(b)

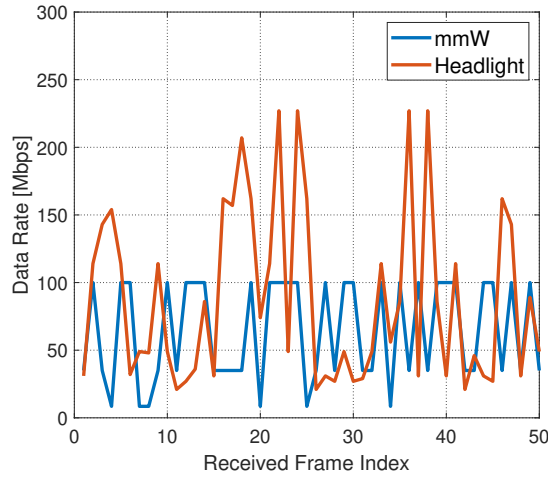
Figure 56: Environment of the experimental campaign: (a) Side road and (b) main road.

MHz bandwidth for VLC system), the received data rates are normalized accordingly. The measurements are conducted in streets nearby HHI's building in Berlin during the ESR secondment. We equip two vehicles with both VLC and mmW links. These two

vehicles follow each other in a realistic driving scenario. On mobility conditions, the normalized received data rates for each road under-consideration are illustrated in Figure 57. Higher data rates for VLC link are observed. Moreover, mmW link sometimes gives a very low data rate as a result of the signal reflections from the nearby vehicles' bodies. In addition, mmW causes EMI to nearby RF systems. This makes VLC relatively a more suitable connectivity technology for the scenario under consideration.



(a)



(b)

Figure 57: Comparison between VLC and mmW systems on mobility conditions for: (a) Main roads and (b) side roads.

6.5 Switching Algorithm Deployment

In this section, we propose a new switching Algorithm that maintain the link reliability. As a benchmark, we considered some Algorithms available in the literature such as Hard switching (i.e., Algorithm 1) as well as hysteresis Algorithm (Algorithm 2) to avoid the ping-pong effect.

The hard switching which selects the link that has the highest data rate. It could be written as in Algorithm 1. The flowchart that illustrates the experimental deployment of the hard switching Algorithm is illustrated in Figure 58.

Algorithm 1 Best Connection Algorithm

Input: Rb_{VLC} and Rb_{mmW}

Output: Channel dedication

1 Initialization:

2 While calculate Rb_{VLC} and Rb_{mmW} ;

if $Rb_{VLC} > Rb_{mmW}$ Then

Dedicate VLC channel for reception;

wait for the determined time interval;

else

Dedicate mmW channel for reception;

wait for the determined time interval;

end if

end while

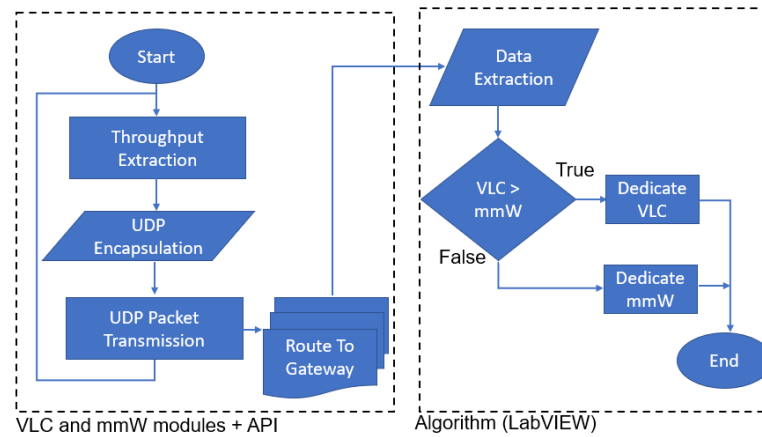


Figure 58: Flow chart of the hard switching Algorithm.

While, the hysteresis Algorithm defines a minimum threshold for the VLC link and it accepts that the VLC link might has a lower than mmW link conditioned to be higher than the threshold. This Algorithm could be written as in Algorithm 2. The flowchart that illustrates the experimental deployment of the Hysteresis Algorithm is illustrated in Figure 59.

Algorithm 2 Hysteresis Algorithm

Input: Rb_{VLC} , $Rb - mmW$ and γ_{min}

Output: Channel dedication

1 Initialization:

2 While calculate Rb_{VLC} and $Rb - mmW$;

if $Rb_{VLC} > Rb - mmW$ and $Rb_{VLC} > \gamma_{min}$

Then Dedicate VLC channel for reception;

wait for the determined time interval;

else

Dedicate mmW channel for reception;

wait for the determined time interval;

end if

end while

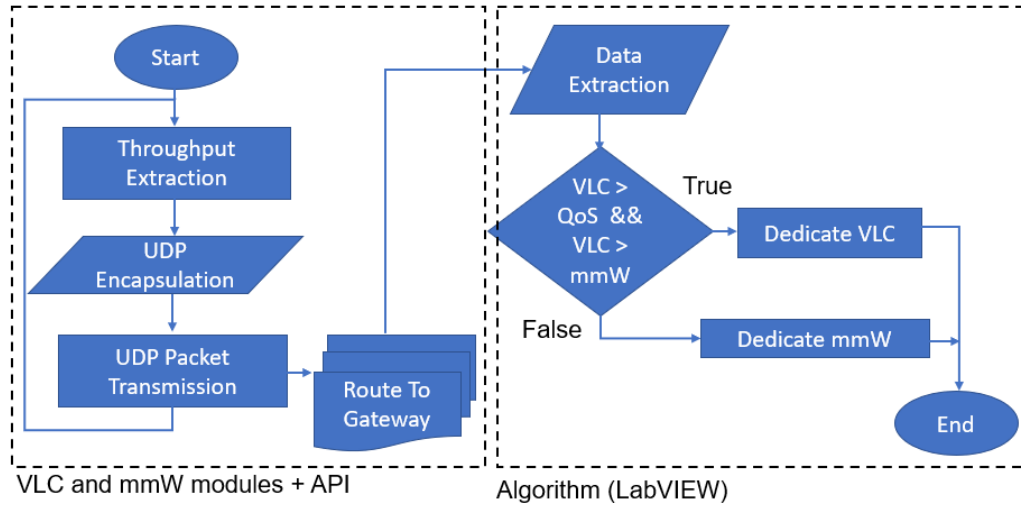


Figure 59: Flow chart of the Hysteresis Algorithm.

On the other hand, the proposed Algorithm activates the mmW link only when it is needed. This reduces the EMI to other vehicles and enhance the power efficiency. The proposed Algorithm is written as in Algorithm 3. The flowchart that illustrates the

experimental deployment of the proposed Algorithm is illustrated in Figure 60.

Algorithm 3 Proposed Algorithm

Input: Rb_{VLC} , $Rb - mmW$, γ_{min} , ROC_{th} and TM

Output: Channel dedication

1 Initialization:

2 While calculate Rb_{VLC} , $Rb - mmW$, ROC and $T - on - VLC$;

if $ROC > ROC_{th}$ Then

 Activate mmW;

 Wait for the determined interval;

else if $T - on - VLC > TM$ and $Rb_{VLC} > \gamma_{min}$

 Dedicate VLC;

 Hibernate mmW modules;

 Wait for the determined interval;

else

 Activate mmW;

end if

end while

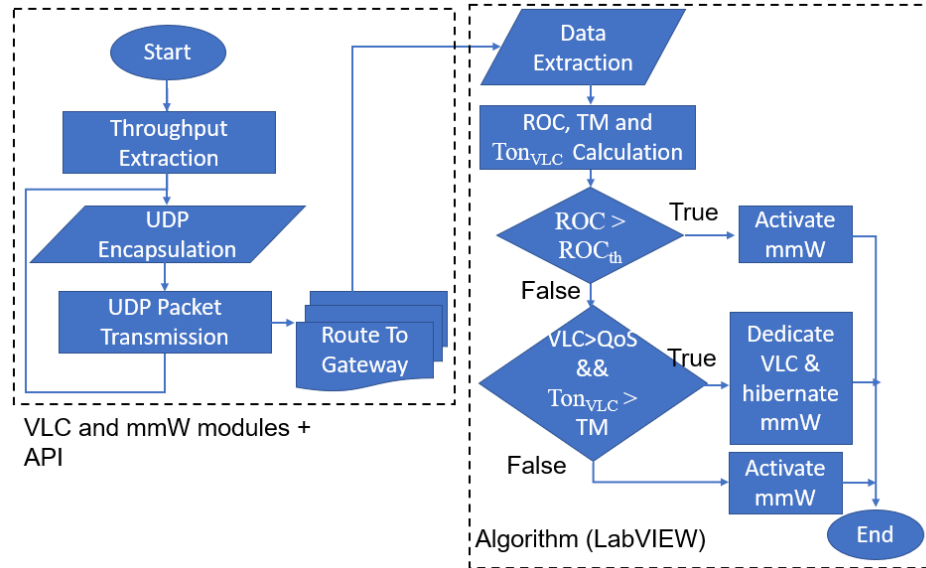
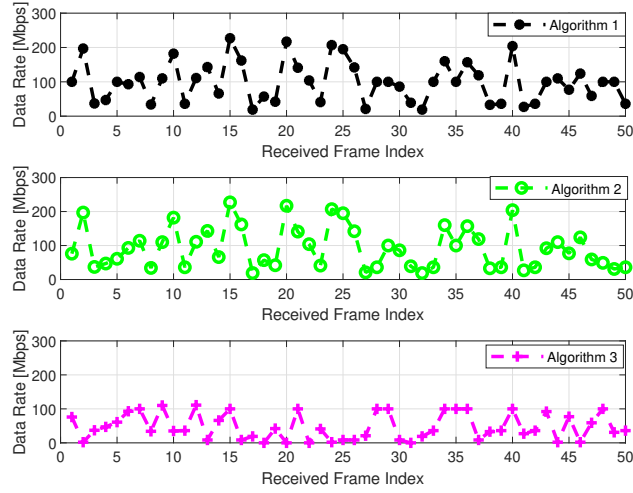
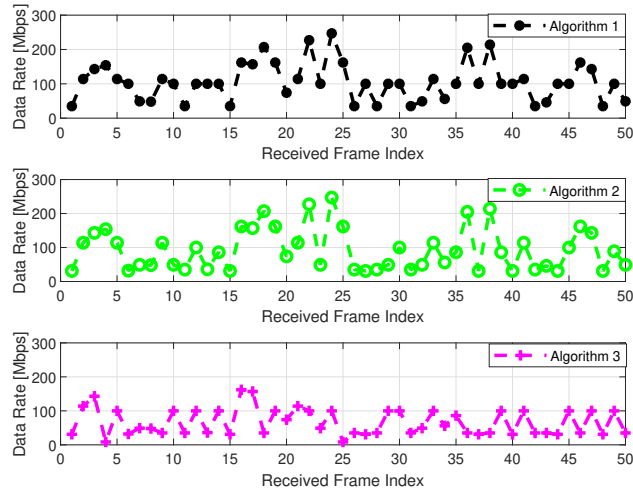


Figure 60: Flow chart of the proposed Algorithm.

During the measurements, we deployed the aforementioned algorithms and observed the performance. The performance of the main roads and side roads is illustrated in Figure 61. It could be seen that the proposed hard switching Algorithm (Algorithm-I) provides the highest data rates. While the second Algorithm gives a relatively lower



(a)



(b)

Figure 61: Data rates of different Algorithms at: (a) Main road and (b) side road.

data rates but it avoids the ping-pong effect. On the other hand, the proposed algorithm gives the lowest data rates but it supports data rates sufficient for satisfying the needed QoS.

In addition, the proposed algorithm deactivates the mmW link in the presence of VLC link. This gives a higher power efficiency for the proposed Algorithms compared to other Algorithms under-consideration (See Figure 62).

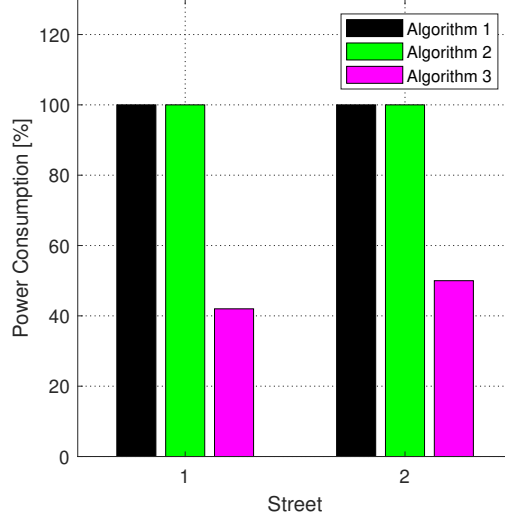


Figure 62: Power efficiency of different switching Algorithms.

6.6 Conclusion

In this chapter, we considered the use of dual connectivity approaches (VLC and mmW) for V2V applications. For a fair comparison, we first developed a custom-designed frontend to increase the VLC link data rates. Then we deployed a specific API to make both technologies operate in a real-time manner. Measurement demonstrated that VLC link is a relatively more suitable connectivity technology for the scenario under consideration. Furthermore, our proposed switching algorithm has high power efficiency and satisfies the needed QoS to support reliable vehicular communications.

CHAPTER VII

CONCLUSIONS AND FUTURE WORK

In this chapter, we review the entire dissertation objectives and results. Then we emphasize the contributions of this dissertation. Finally, we point out some potential topics for future work.

In this dissertation, we presented a comprehensive performance analysis of vehicular VLC systems under mobility conditions. We first investigated the baseband processing and the experimental implementation of different modulation techniques for VLC systems. Then We focused on developing a VLC frontend that supports high power transmission of vehicular LEDs and experimentally characterized impairments of the outdoor environment. A critical aspect of the vehicular VLC system is the channel model with its asymmetrical light sources, thus, we experimentally proposed a pathloss model through a measurement campaign. Counting on these, we investigated the vehicular VLC link performance in the presence of mobility for passenger vehicles as well as for heavy trucks. For benchmarking purposes, we further implemented a hybrid system (VLC/ RF) for comparison purposes as well as deployed a switching algorithm to satisfy a specific (QoS).

The contributions of this dissertation can be summarized as follows:

In Chapter 2, we experimentally evaluated the performance of modulation techniques using USRP platform. We illustrated the experimental set-up and detailed the measurement procedures. We further examined the effective channel that includes the effects of both the propagation channel and frontends. Furthermore, we measured the BER and validate the system results through a comparison with the theoretical values. In addition, we investigated the performance considering different transmitter and receiver orientation angles, LoS, NLoS scenarios. Furthermore, we identified the proper

lens combination to be used for real-time driving scenarios considering both increasing the communication distance and the angular limitations.

In Chapter 3, the experimental impairments of vehicular VLC systems have been investigated. Firstly, we experimentally investigated the effect of the background noise variation (i.e., sunlight) on the vehicular VLC link. In addition, to extend the communication distance, we characterized the multihop vehicular VLC links in both daytime and nighttime. We further measured the BER for each individual hop and then obtain the overall performance of the multihop system.

Chapter 4 of the thesis focused on investigating the effect of mobility. Particularly, we employed multiple photodetectors on the receiving vehicle and identified each position/ orientation through a measurement campaign. We then considered different roads (straight/ curved) and the lens employment/ unemployment cases. We further considered daytime and nighttime measurements.

In Chapter 5 of the thesis, we considered a special case for truck-to-truck VLC communication. We proposed a 3D channel model that considers longitudinal, lateral and vertical distances. Then, we experimentally modelled the truck's vertical oscillation. We then investigated different channel and system parameters on the link performance. Furthermore, in an effort to mitigate the effect of the vertical oscillation, we optimized the photodetector location on the receive vehicle/truck.

In Chapter 6, we compared with RF technology. First, a new frontend for VLC system was designed/validated to support a high data rate (almost similar to RF for fair comparison purposes). Additionally, for hybrid system deployment, both technologies must be able to connect with each other's, This is achieved through an API designed specifically for this purpose. We further proposed a switching algorithm to maintain a specific QoS and benchmarked with different algorithms in the literature.

In addition to the aforementioned chapters, the following research points could be further considered as future works:

In Chapter 5, results demonstrated that the effect of the photodetector location on

the receiving vehicle has a tremendous effect on the system performance. For a special case (Truck-to-Truck VLC), we optimized the photodetector location. This direction might be beneficiary to consider different types of vehicles and becomes a generic optimization work.

To get the maximum benefit of the VLC system, the bottleneck (i.e., Tx frontend) has to be further developed. As illustrated in Chapter 6, we were able to achieve a 10 dB bandwidth of 50 MHz for the white light headlight. This could be further extended to increase the data rate as well as the system capacity.

Since the demand for high data rate vehicular applications is continuously increasing, the integration of VLC with other emerging V2X solutions (i.e., DSRC, CV2X or mmW) becomes strongly required. In Chapter 6, we already experimentally deployed a switching Algorithm to maintain a specific QoS. Further improvement might be a research topic for future vehicular connectivity applications.

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VITA

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