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BROADBAND OPTICAL WIRELESS COMMUNICATIONS

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Electrical Engineering

by

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ABSTRACT

As Light-Emitting Diode (LED)'s increasingly displace incandescent lighting over the next few years, general applications of Indoor Optical Wireless Communications (IOWC) technology are expected to include wireless internet access, broadcast from LED signage, and machine-tomachine communications. This dissertation explores several fundamental research topics of IOWC.

In this dissertation, the author develops a simulation method to generate IOWC channel models by tracking light reflections. The method is further optimized by investigating the contribution of each order of reflections and proposing a calibration method. Based on the channel models, the author reveals that sources' Half-Power Angles (HPA), receivers' Field-Of Views (FOV), sources layout, and the power distribution among sources are the significant factors impacting IOWC system quality.

The dissertation also investigates the applications of advanced communication techniques to improve IOWC performance. The author starts by performing bit-error rate (BER) distributions and outage probability estimations using intensity modulation / direct detection (IM/DD) with additive white Gaussian noise (AWGN). Next, various orthogonal frequency-division multiplexing (OFDM) schemes are applied to mitigate the multipath effect in IOWC and decrease clipping noise. Precoding approaches are also used to reduce peak-to-average power ratio (PAPR) in OFDM systems. In addition, the multiple-input and multiple-output (MIMO) methods are explored to increase system reliability and/or band efficiency.

Finally, this dissertation extends its topics to specific industrial applications of IOWC. By customizing the channel simulation for airplane cabins, the visible light propagation features in

this environment are investigated. The results confirmed the effectiveness of applying IOWC to airplane cabin wireless communications. An efficient algorithm to apply IOWC to indoor navigations is also developed.

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Chapter 1

Introduction

1.1 Motivation

High speed communication has found an important role in people's daily lives [1]. The Wireless Home Link (WHL) is becoming more and more a reality. In the next decades, wireless communication will play a significant role in electronic devices interconnections.

Visible Light Communications (VLC) is an emerging wireless communication technology based on white Light-Emitted Diode (LED). The LED is generally considered as the next generation light source and may replace the universal incandescent bulb and fluorescent bulb in home and work places, because of its advantages such as long lifetime, low power consumption, small size and being environment friendly. Moreover, LED has a high response sensitivity to support high speed communication. In VLC, LED takes both communication and illumination duties. We modulate the LED by user data to create illumination and communication dual functional "Base Station Light", while human can hardly sense the flickers due to high modulation speed of hundreds of megabit per second. A typical VLC system is demonstrated by Fig. 1-1.

Comparing with conventional Radio Frequency (RF) wireless communication, the VLC has many advantages, such as high data rate, energy saving, secure transmission, and lack of electromagnetic interference and most importantly, spectral regulation [1][2][3][4][5][6][7][8]. First, by some novel technique, VLC is able to provide 513Mbps transmission, according to [8].

Second, light source combines illumination and communication. Third, light is confined in the room and this offers physical-layer security. Fourth, VLC causes no interference in electromagnetic sensitive environments and provides an effective wireless solution to these environments, such as hospitals, aircrafts and others. Finally, and most importantly, VLC offers sufficient bandwidth resources, free from regulation. Currently, bandwidth for wireless communication is exhausted in the microwave range and is strictly controlled by the Federal Communications Commission (FCC).



Fig. 1-1: Typical VLC system (from Chinese Academy of Science)

1.2 Objective

Recognizing the importance of VLC, there has been much research on this topic which involves numerous aspects of VLC. References [5] and [9] provided imaging diversity receivers for indoor optical wireless communication. Reference [10] proposed a novel VLC modulation scheme called Pulse Dual Slope Modulation (PDSM) which solved intra-frame and inter-frame flickers. Reference [11] designed spread spectrum code for VLC to improve system performance. Reference [12] integrated dimming control with VLC. The dimming control made users feel psychologically comfortable when they use the VLC.

Though these investigations have made substantial contribution to VLC, they mostly work on the single source scenarios, rather than multiple sources. In order to fill the gap, this work focuses on the interference among sources when plural light sources are applied. Since most rooms install plural light sources on the ceiling, plural sources produce light coverage areas overlapping, introducing Inter-symbol Interference (ISI). The ISI degrades system performance.

This thesis consists of several fundamental researches of multiple sources VLC. Source layout is one of the most important factors that affect light footprints overlapping and thus producing ISI. It determines the pattern and extent of the lights overlapping. We will explore the VLC performance in several conventional household layouts and investigate the impact of these layouts to VLC. Multiple sources increase multipath distortion. As orthogonal frequency-division multiplexing (OFDM) is proved to be effective in reducing multipath-involved ISI, we will investigate this modulation scheme for VLC applications. Multiple-input and multiple-output (MIMO) techniques will also be included as they provide either reliability improvement or bandwidth efficiency increase. Based on these investigations, we will further explore VLC performance in real applications, such as aircraft cabin wireless communications and indoor navigations.

1.3 Organization

In this thesis, the Lambertian emitting pattern of LED and the diffusion features in indoor environment are studied. Based on the theory, light pulses are traced to establish a MIMO indoor wireless channel model on a specific sources layout. Next, test data is generated to simulate BER distribution in a room and the outage probability is calculated. After that, various OFDM modulation schemes are applied to mitigate multipath distortion. Furthermore, the performance improvement is expected when MIMO techniques are applied. After that, VLC performances in specific applications are investigated, including aircraft onboard wireless communications and indoor navigations. An efficient VLC based positioning algorithm is proposed.

The rest of the thesis is organized as follows. Chapter 2 will describe indoor optical wireless channel and modeling method. Chapter 3 will provide detailed analysis of model accuracy and propose a calibration method. Chapter 4 will introduce bit-error rate (BER) and outage probability simulation results; impact factors to optical wireless communication systems will be investigated. Chapter 5 will explore several OFDM approaches to improve VLC performance. Chapter 6 will investigate the VLC improvement by applying novel MIMO techniques. Chapter 7 will validate the VLC solutions to airplane wireless communications. Chapter 8 will propose a novel positioning algorithm using VLC. Chapter 9 will provide some thoughts about future work and summarize the thesis.

Chapter 2

Indoor Optical Wireless Model

2.1 Introduction

Modeling indoor wireless channel is fundamentally tracing the light pulse, which experiences multiple Lambertian reflections. A light pulse is emitted at the source and propagates to all directions, while the directional power distribution obeys Lambertian law. When a part of the light beam arrives at a point on room surface, it is reflected in Lambertian pattern and this point works as the secondary light source. The process continues until the light reaches the receiver / user. Since the successive captured lights at the receiver come from the same pulse, but experience different optical paths, the overall receiver response demonstrates the impulse response from the source to the receiver. In most cases, there are plural sources and receivers; the indoor wireless channel is described by an impulse response matrix.

2.2 Single Input Single Output (SISO) model

Indoor optical wireless channel characteristic by computer simulations was first presented by Barry [13]. The authors set up a simulation tool to evaluate multipath impulse response of optical indoor channel.

Barry's room model divides the room surfaces including ceiling and floor into many grid elements. Each element is called a reflector and is assigned a reflection coefficient. The light pulse travels from a source to all the reflectors in a Line-of-Sight (LOS) Lambertian pattern. When light arrives at one of these, light intensity will be decreased by the multiplying reflection factor. Then, this element will be seen as a secondary Lambertian source emitting light pulse to all other reflectors. This procedure continues until light arrives at the system receiver.



Fig. 2-1: Line-of-Sight Lambertian transmission

As shown in Fig. 2-1, the impulse response of LOS Lambertian transmission between elements is expressed as:

$$h^{(0)}(t;S,R) \approx \frac{n+1}{2\pi} \frac{\cos^{n}(\phi)\cos(\theta)A_{R}}{R^{2}} rect(\theta/FOV)\delta(t-R/c) \quad (2-1)$$

where superscript 0 means light travels from source to receiver directly without passing through objects (line-of-sight (LOS) is required); *S* and *R* represent source and receiver parameters set respectively; *n* is the mode number of the radiation lobe that specifies the source directionality; ϕ is the angle between source orientation vector and the vector pointing from source to receiver; θ is the angle between receiver orientation vector and the vector pointing from receiver to source; A_R is receiver area; *FOV* is the Field-of-View (FOV) of the receiver; *R* is the distance between the source and receiver; *c* is the speed of light. When light reaches the destination reflector, the transmission continues and the destination reflector becomes the starting source of next transmission. Therefore, the total impulse response can be calculated recursively. Let $h^{(k)}(t;S,R)$ denote the response of light undergoing exactly *k* reflections.

$$h^{(k)}(t; S, R) = \int_{S}^{\cdot} h^{(0)}(t; S, E_R) \otimes h^{(k-1)}(t; E_S, R) \, ds \quad (2-2)$$

where E_R is the parameter set of a reflector as a receiver; E_S is the parameter set of a reflector as a transmitter. The integration carries on all reflectors. As a result, the overall impulse response is the infinite sum of the impulse responses undergoing all possible number of reflections.

$$h(t; S, R) = \sum_{k=0}^{\infty} h^{(k)}(t; S, R) \quad (2-3)$$

2.3 Multiple Input Multiple Output (MIMO) model

In most cases, where there are plural light sources and users. It is necessary to extend a MIMO indoor wireless channel model from the SISO model.

Kavehard and Alqudah developed the MIMO indoor optical wireless channel simulation based on diffuse-transmission configuration [3]. This method transforms impulse response into a matrix form.

In the MIMO model, considering the transmitter and receiver, the transfer function between any two points is divided into four components. Suppose there are J sources, N surface elements and M receivers, the four components are: the transfer function between a source and surface element ($F_{J\times N}$), the transfer function between surface elements ($\Phi_{N\times N}$), the transfer function between surface element and receiver ($G_{N\times M}$), and the direct transfer function between source and receiver ($D_{J\times M}$).

2.3.1 Source Profile ($F_{J \times N}$)

 $F_{J\times N}$ is a J by N matrix.

Each entry of the matrix f_{sk} is the transfer function between a source s and element k as

$$f_{sk} = \frac{n+1}{2\pi} \frac{\cos^{n}(\phi_{sk})\cos(\theta_{sk})A_{k}}{R_{sk}^{2}} \delta(t - \frac{R_{sk}}{c})u(\frac{\pi}{2} - \theta_{sk}) \quad (2-5)$$

where *n* is the Lambertian order of the source; ϕ_{sk} is the emitting angle from source *s* to element *k*; θ_{sk} is the incident angle from source *s* to element *k*; R_{sk} is the distance between source *s* and element *k*; A_k is the area of the element; *c* is speed of light; *u*() is unit step function.

2.3.2 Environment Matrix ($\Phi_{N \times N}$ **)**

 $\Phi_{N \times N}$ is a N by N matrix. In matrix format, considering up to n reflections, it is expressed as

$$\Phi_{N \times N} = \begin{cases} I_{N \times N} + \Psi_{N \times N} + \Psi_{N \times N}^{2} + \Psi_{N \times N}^{3} + \dots + \Psi_{N \times N}^{n-1}, & n \ge 2\\ I_{N \times N}, & n = 1 \end{cases}$$
(2-6)

where $I_{N \times N}$ is the N by N identity matrix, and $\Psi_{N \times N}$ is given by

$$\Psi_{N\times N} = \begin{bmatrix} \psi_{11} \ \psi_{12} \ \cdots \ \cdots \ \psi_{1N} \\ \psi_{21} \ \psi_{22} \ \cdots \ \cdots \ \psi_{2N} \\ \cdots \ \cdots \ \cdots \ \cdots \\ \psi_{N1} \ \psi_{N2} \ \cdots \ \cdots \ \psi_{NN} \end{bmatrix}$$
(2-7)

 ψ_{ik} represents the transfer function between two elements *i* and *k* as

$$\psi_{ik} = \begin{cases} 0, & i = k \\ \frac{\rho_i \cos(\phi_{ik}) \cos(\theta_{ik}) A_k}{\pi R_{ik}^2} \delta(t - \frac{R_{ik}}{c}) u(\frac{\pi}{2} - \theta_{ik}), & i \neq k \end{cases}$$
(2-8)

where ρ_i is the reflection coefficient of element *i*; ϕ_{ik} is the emitting angle from element *i* to element *k*; θ_{ik} is the incident angle from element *i* to element *k*; R_{ik} is the distance between element *i* and element *k*; A_k is the area of the element; *c* is speed of light.

2.3.3 Receiver Profile ($G_{N \times M}$)

 $G_{N\!\times\!\!M}$ is a N by M matrix. It is represented as

$$G_{N\times M} = \begin{bmatrix} g_{11} & g_{12} & \cdots & \cdots & g_{1M} \\ g_{21} & g_{22} & \cdots & \cdots & g_{2M} \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ g_{N1} & g_{N2} & \cdots & \cdots & g_{NM} \end{bmatrix}$$
(2-9)

The entry g_{ir} is the transfer function from element *i* to receiver *k* as

$$g_{ir} = \frac{\rho_i \cos(\phi_{ir}) \cos(\theta_{ir}) A_r}{\pi R_{ir}^2} \delta(t - \frac{R_{ir}}{c}) u(FOV_r - \theta_{ir}) \quad (2-10)$$

where ρ_i is the reflection index of element *i*; ϕ_{ir} is the emitting angle from element *i* to receiver *r*; θ_{ir} is the incident angle from element *i* to receiver *r*; R_{ir} is the distance between element *i* and receiver *r*; A_r is the area of the receiver; *c* is speed of light; FOV_r is the FOV of the receiver *r*.

2.3.4 Direct Response Matrix $(D_{J \times M})$

 $D_{J \times M}$ is a J by M matrix. It is represented as

The entry d_{ir} is the transfer function from element s to receiver r as

$$d_{sr} = \frac{n+1}{2\pi} \frac{\cos^n(\phi_{sr})\cos(\theta_{sr})A_r}{R_{sr}^2} \delta(t - \frac{R_{sr}}{c})u(FOV_r - \theta_{sr}) \quad (2-12)$$

where *n* is the Lambertian order of source *s*; ϕ_{sr} is the emitting angle from source *s* to receiver *r*; θ_{sr} is the incident angle from source *s* to receiver *r*; R_{sr} is the distance between element *s* and receiver *r*; A_r is the area of the receiver; *c* is speed of light; FOV_r is the FOV of the receiver *r*.

2.3.5 Total Response $(H_{J \times M})$

The total impulse response matrix of the MIMO system is given by:

$$H = D_{J \times M} + F_{J \times N} \otimes \Phi_{N \times N} \otimes G_{N \times M}$$
 (2-13)

where \otimes indicates matrix convolution.

2.4 Typical Impulse Response Distortions

In most locations, the impulse response has insignificant impulse spread. However, two kinds of locations exhibiting considerable impulse distortion are found. The first kind is the room corner (0.2, 0.1); the second kind is the overlapping area of light footprints (2.2, 2.4).

Two typical types of distorted impulse responses from the entries of H are demonstrated in Fig. 2-2. At the room corner, the spread of impulse response comes from multipath effect of light reflections. In the corner area, the receiver captures reflected lights, which experience different reflection paths, and cause the decreasing tail in impulse response. What is worth mentioning is that the tail has much lower power compared with the peak.

In the overlapping areas, several equally high sharp peaks in the impulse response are observed. The reason for the multiple peaks is that lights from different sources enter receiver via different LOS paths, the time difference of the arrivals causes the multiple peaks. This is an important influencing factor to VLC.



Fig. 2-2: (a) multi-path effect (b) inter-source interference

2.5 Conclusions

In this chapter, the approaches to establish indoor optical wireless models are investigated. These approaches are based on tracing light reflections in the environment. The most efficient method is proposed by Kavehrad and Alquadah. In their method, the entire impulse response matrix is divided into several stages. Each stage provides a group of parameters. This method substantially saves simulation time when we calculate the impulse response matrix for multiple receiver locations is calculated. Once the environmental matrix $\Phi_{N\times N}$ is obtained at one location, it is saved and reused for other locations.

Chapter 3

Indoor Optical Wireless Channel Model Error Analyses and Calibration

3.1 Introduction

This chapter analyzes the impact of high order light reflections on Indoor Optical Wireless (IOW) channel models. Based on the authors' findings, a calibration method is proposed to reduce model errors. Channel models are generated by tracing and adding up diffuse light reflections and sequential sub-reflections along its traveling path. As computation complexity increases significantly with the number of reflection orders considered, researchers conventionally take the contribution of a first few orders, most commonly three, to represent the complete channel. Discarded high-order reflections bring no significant performance difference to low-speed systems; however, major contemporary IOW research institutions focus on highspeed Gbps communications where their impact is no longer negligible. Root-Mean-Square (RMS) delay-spread, for instance, is severely underestimated by neglecting high-order reflections. We simulate an IOW system in an ordinary 6m*6m*3m lab room and calculate the contributions of each order of reflections at 841 locations. It shows the RMS delay-spread estimation using first three orders is underestimated by 15.3% on the average and 26.6% as the maximum. To limit error within half a symbol period, 1Gbps and 10Gbps systems tolerate underestimations up to 13.7% and 1.4%, respectively. These must be achieved by applying first 5 and 9 orders. To keep the computation efficiency of low order reflection models and improve their accuracies, we propose a statistical calibration method. It reduces average model error of first three reflection orders from 15.7% to 4.3%. After calibration, numbers of orders required by 1Gbps and 10Gps systems are individually reduced to 3 and 7.

3.2 Historical Review of Indoor Optical Wireless Channel Research and Error Analyses

Indoor Optical Wireless Communication (OWC) has been studied extensively in recent years. It is recognized as a strong candidate technology for the next generation high-speed wireless networks. Comparing with conventional Radio Frequency (RF) wireless communications, it offers significant advantages [14][15][16][17]. First, the visible, infrared and ultraviolet spectral regions offer virtually unlimited bandwidth and are not regulated. Second, light is confined in rooms because it is not able to penetrate opaque barriers, such as walls, ceilings and floors. This feature enables band reuse between neighboring rooms and provides physical layer communications security. Third, OWC neither generates nor is susceptible to electromagnetic interferences. This technology, therefore, can be widely applied to sensitive environments including hospitals, aircrafts, mines, power plants and others.

Realizing the substantial potentials of OWC, many research institutions are being sponsored in this area by government and industry. In the United States, the Pennsylvania State University and Georgia Institute of Technology lead the NSF Centre on Optical Wireless Applications (COWA). They collaborate with industrial leaders to evaluate the potentials of the interdisciplinary research center activities, in providing leadership to develop new generation of environment-friendly and extremely wideband optical wireless technology applications [18]. The Center for Ubiquitous Communication by Light (UC-Light) is established by University of California, Riverside. This center focuses on white Light-emitting Diodes (LEDs) for wireless information sharing and retrieving [19]. In Europe, the hOME Giga Access (OMEGA) project started on 2008. It will develop a user-friendly home area network capable of delivering high-bandwidth services and content at a transmission speed of one Gigabit per second [20]. In Japan, the Visible Light Communication Consortium (VLCC) was founded in 2003. It is aiming to publicize and standardize the visible light communication technology, which has been discussed and evaluated in various industry fields [21].

In OWC, the channel model is one of the most important research subjects, since it indicates the transmission capacity and communication performance. The first channel characteristics model study for infrared was made by Gfeller and Bapst in 1979 [22]. Later, a recursive simulation method for diffusion indoor optical wireless channel was developed by Barry and Kahn in 1993 [9]. Not long after, Alqudah and Kavehrad invented the method to simulate Multiple-Input and Multiple-Output characteristics of indoor optical wireless link [3]. Though these methods substantially increase simulation efficiency, the computation still consumes considerable time and the complexity significantly increases with number of reflection orders concerned. To make sure the model can be simulated in a reasonable amount of time, researchers use channel models considering only first a few bounces, most commonly three, to represent the complete model. This approximation has been generally accepted for decades, because of the insignificant performance difference in relatively low speed transmissions at that time. As present leading research institutions are moving to Gbps high speed transmissions, this approximation no longer holds. Consequently, there is an urgent demand for explorations on higher order reflections and involved model errors. This chapter analyzes the impact of highorder reflections on channel models. Based on our research, we summarize the general rules of model error distributions, with bounce order increase. A calibration method is developed to reduce errors and at the same time maintain computation efficiency.

The rest of the chapter is organized as follows: in section 3.3, we discuss the methods to create indoor optical wireless channel model; in section 3.4, we demonstrate the simulation results and discuss the impact of high-order reflections; then, we propose a calibration method to low-order channel models in section 3.5 and draw conclusions in section 3.6.

3.3 Indoor Optical Wireless Channel Model Analyses

3.3.1 Indoor Optical Wireless Channel Overview

A typical indoor optical wireless communication system is demonstrated by Fig. 3-1. The transmitter is placed in the center of the room and generates multiple diffusion spots on the ceiling. Receivers spread in the room and receive data from the spots [23][24].



Fig. 3-1: A Typical Indoor Optical Wireless Communication System

Indoor optical wireless channels are characterized by impulse responses (IRs). In this chapter, we consider multiple spots sending data on the ceiling as spatially diverse transmitters. The IRs are generated correspondingly when the receiver is placed at different locations in the room. Assuming there are M spots and N receiver locations, the channel characteristic profile of this communication system is represented by an IR matrix $H_{M\times N}(t)$ as:

The element $h_{ij}(t)$ indicates the impulse response from the i - th spot to the j - th location. As multiple spots are applied for spatial diversity, assuming equal gain combining is applied, the total impulse response when the receiver is placed at the j - th location is:

$$h_i(t) = \sum_{i=1}^M h_{ii}(t)$$
 (3-2)

3.3.2 Channel Model Simulation

The principle of channel simulation is partitioning a room into numerous reflectors and tracing the light diffusive reflections from them. In this simulation, we place the laser transmitter in the center of the room, pointing upwards. By dividing the beam, it generates multiple diffusion spots on the ceiling. The diffusions follow Lambertian pattern and light travels to all reflectors through Line-Of-Sight (LOS) paths. When light arrives at a reflector, after reflection loss, the reflector becomes a secondary diffusion spot and retransmits light to other reflectors in the same manner. These diffusions keep repeating in the room, with intensity decrease from propagation attenuations and surface absorptions. By collecting the lights from all diffusions at the receiver,

we get the total IR through diffusion paths. It is apparent that a diffusion channel is fundamentally a combination of a large amount of scaled and delayed LOS channels.

The LOS impulse response from one transmitter *i* to receiver $j g_{ij}^{(0)}(t)$ is given by [9]:



$$g_{ij}^{(0)}(t) \approx \frac{1}{\pi} \frac{\cos(\phi_{ij})\cos(\theta_{ij})A_{Rj}}{R_{ij}^2} \delta\left(t - \frac{R_{ij}}{c}\right) rect\left(\frac{\theta_{ij}}{FOV_j}\right) \quad (3-3)$$

Fig. 3-2: LOS Transmission

As shown in Fig. 3-2, the superscript (0) indicates that it is a LOS IR, because 0 reflections are experienced from *i* to *j*; ϕ_{ij} is the angle between the vector pointing from transmitter to receiver and the transmitter normal; θ_{ij} is the angle between the vector pointing from receiver to transmitter and the receiver normal; A_{Rj} is the effective receiver area; R_{ij} is the distance between the two; *c* is the speed of light; *FOV_j* is the Field-Of-View (FOV) of the receiver.

Previous research successfully indicates that the IR from one point *i* to another *j* experiencing exact *k* bounces can be calculated from the IRs experiencing exact k - 1 bounces and LOS IRs, as
$$g_{ij}^{(k)}(t) \approx \sum_{l=0}^{Q} g_{il}^{(0)}(t) \otimes g_{lj}^{(k-1)}(t)$$

= $\frac{1}{\pi} \sum_{l=0}^{Q} \frac{\rho_l \cos(\phi_{il}) \cos(\theta_{il}) A_{Rl}}{R_{il}^2} rect \left(\frac{\theta_{il}}{FOV_l}\right) g_{lj}^{(k-1)} \left(t - \frac{R_{il}}{c}\right)$ (3-4)

 ρ_l is the reflectivity of reflector *l*; *Q* is the total number of reflectors in the model. By recursively using (3-4), we can generate the IRs experiencing arbitrary number of reflections from LOS IRs.

In this way, the accumulated IR concerning first k orders reflections $h_{ij}^{(k)}(t)$ is the sum of the IRs that experience exact l orders (l = 1, 2, 3, ..., k).

$$h_{ij}^{(k)}(t) = \sum_{l=0}^{k} g_{ij}^{(l)}(t)$$
 (3-5)

In [9], the authors show the runtime to compute $g_{ij}^{(k)}(t)$ is roughly exponential in k. At that time, it required approximately 24 hours to calculate k = 3 bounces of IR with 2776 elements. Though in recent decades the simulation method has been optimized and the computation capacity has been upgraded, the computational complexity is still considerably high and increases significantly with reflection orders. In a practical scenario, people still have to use only first a few orders, most often three, to approximate the complete IR $h_{ij}(t)$.

3.3.3 Channel Features

Signal-to-Noise Ratio (SNR) and Inter-Symbol Interference (ISI) are essential factors determining communication performance. The former can be estimated by received power at the receiver while the latter can be indicated by average delay and delay-spread. We therefore focus our research on power attenuation, average delay and delay-spread. For power issues, we shall keep in mind the differences between OWC systems and RF. The received optical power is proportional to the IR magnitude. Since optical power is linearly converted to the amplitude of

electrical/signal current, the received electrical/signal power is proportional to the square of IR magnitude.

On the power side, to exclude the influence of ISI, we assume the transmitters are sending a unity amplitude square impulse x(t) in a symbol period. For the received optical power at location *j*, it is calculated by

$$P_{oj} = \frac{1}{T} \int_0^T x(t)^* h_{ij}(t) dt \quad (3-6)$$

For the received signal power at location *j*, it is calculated by

$$P_{sj} = \frac{1}{T} \int_0^T (x(t)^* h_{ij}(t))^2 dt \quad (3-7)$$

On the ISI side, the average delay is the first moment of the power delay profile with respect to the first arriving path, defined as [25]:

$$\mu_{j} = \int_{-\infty}^{\infty} t h_{ij}^{2}(t) / h_{ij}^{2}(t) dt \quad (3-8)$$

The Root-Mean-Square (RMS) delay-spread is the square root of the second central moment of a power delay profile as:

$$s_{j} = \sqrt{\int_{-\infty}^{\infty} (t - \mu_{j})^{2} h_{ij}^{2}(t) / h_{ij}^{2}(t) dt} \quad (3-9)$$

Delay-spread is a good measure of multipath distortion and indicates potential ISI. Previous research shows maximum transmission rate over a wireless channel is determined by the inverse of its delay-spread, given that no diversity or equalization applied [26][27].

3.4 Simulation and Results

We, at first, choose three typical test locations to analyze the contribution of each order of reflections to IR. The three locations are (unit: m): A (0, 0, 0.9) representing a point at the room

corner, where severe diffusions are experienced; B (0, 3, 0.9) representing a point near a wall but away from corners, where medium diffusions are experienced; C (3, 3, 0.9) representing a point at the center of the room, where weak diffusions are experienced. Next, we demonstrate the estimation accuracy distributions all over the room. The IOW system has 9 transmitting spots generated from one laser source for spatially diverse purpose and the simulation parameters are given in Table 3-1.

Room size,	6m*6m*3m
length*width*height (unit: m)	
Laser source location (unit: m)	(3, 3, 0.5)
Diffusion transmitting spots	(1.5, 1.5, 3) (1.5, 3, 3) (1.5, 4.5, 3)
location (unit: m)	(3, 1.5, 3) (3, 3, 3) (3, 4.5, 3)
	(4.5, 1.5, 3) (4.5, 3, 3) (4.5, 4.5, 3)
Transmission power at each spot	1
(unit: W)	
Reflection coefficients	0.9, 0.7, 0.1
(ceiling, wall, floor)	
Reflection elements size	0.2*0.2
(unit: m*m)	
Receiver FOV	60
(unit: degrees)	
Receiver Aperture Area	1e-4
(unit: m²)	
Time resolution (unit: ns)	0.66

TABLE 3-1: ROOM MODEL SIMULATION PARAMETERS

3.4.1 High-order Reflection's Impact to IR

Fig. 3-3, Fig. 3-4, and Fig. 3-5 show the contributions of each order of reflections to the IRs at test locations A, B and C. As given in (3-5), the total IR is the sum. In all three locations, the 0-th reflection (LOS) contributes the major impulses of the IRs, which contain most of the optical power. From the 3-rd order reflections, the shapes of individual order IRs are similar; nevertheless, they attenuate and temporally spread out as order increases. Comparing with low order reflections, high order reflections contribute less amplitude but more delay to IR. Unlike received power which is only related to IR amplitude, delay-spread is jointly determined by IR amplitude and delay. High-order reflections, therefore, obviously make more significant impact to delay-spread than received power. In other words, delay-spread estimation should converge slower than power with reflection orders increase.



Fig. 3-3: Impulse response of each order of reflections at location 1



Fig. 3-4: Impulse response of each order of reflections at location 2



Fig. 3-5: Impulse response of each order of reflections at location 3

3.4.2 Model Accuracy Analyses

To further explore the convergence differences, we plot the estimation accuracy of received optical power, received signal power, average delay and RMS delay-spread in Fig. 3-6 to Fig. 3-8. We use the results applying first 20 orders of reflections as references of accurate estimation. Accuracy is defined by the ratio of estimated value to referred accurate value.



Fig. 3-6: Channel characteristics estimation accuracy at location 1



Fig. 3-7: Channel characteristics estimation accuracy at location 2



Fig. 3-8: Channel characteristics estimation accuracy at location 3

The figures show that signal power and average delay converge fast and shall not be noticeably impacted by truncating them to first two or three orders; however, the received optical power and the RMS delay-spread converge substantially slower. For a Gbps high-speed transmission system, delay-spread deserves particular attention, because it dominantly determines ISI. The maximum delay-spread we observed all over the room is 3.66ns. To make sure any delay-spread model error is smaller than half of a symbol period, we need the delay-spread estimation accuracy higher than $1 - \frac{1e-9}{2\times 3.66e-9} \times 100\% = 86.3\%$ for 1Gbps systems and $1 - \frac{1e-10}{2\times 3.66e-9} \times 100\% = 98.6\%$ for 10Gbps systems. As the figures show, estimation accuracies for the three locations by first three orders are only 73.4%, 82.9% and 90.7%, respectively. If we only use first three orders reflections to create the model, only location C meets the accuracy needed for 1Gbps and none of them satisfy 10Gbps.

3.4.3 Delay-spread Spatial Distributions

As IOW projects are to provide full coverage and mobility, it is necessary to extend channel model analysis from the three test points to the entire area of the room. Our research shows high-order reflections make impacts differently at different locations. The spatial distribution is explored by simulating 841 channels, representing every piece of 0.2m×0.2m area of a 6m×6m×3m room. As we demonstrated that delay-spread experiences most severe impact from discarded high-order reflections, we utilize delay-spread estimation accuracy to indicate model accuracy as the worst case. Its contours for each additional order of reflections considered are shown in Fig. 3-9 to Fig. 3-16, respectively. Generally, the shapes of the contours are similar as number of reflections increases: high accuracy areas are near the center of the room and the

accuracy decreases when approaching the corners of the room, where multi-path effect is much richer; we also observe that dense contours exist at the corner of the room, which indicates a sharp decrease of model accuracy. We are able to get the required reflection orders for specific transmission rate from the contours. For instance, to ensure the accuracy of the entire room is above the need for 1Gbps data rate, we need 5 orders (86.3% above accuracy guaranteed) of reflections and 9 orders (98.6% above accuracy guaranteed) for 10Gbps.



Fig. 3-9: Delay-spread model accuracy contour considering first 2 orders reflections (%)



Fig. 3-10: Delay-spread model accuracy contour considering first 3 orders reflections (%)



Contour of Delay Spread Estimation Accuracy with first4 orders reflections

Fig. 3-11: Delay-spread model accuracy contour considering first 4 orders reflections (%)





Fig. 3-12: Delay-spread model accuracy contour considering first 5 orders reflections (%)



Contour of Delay Spread Estimation Accuracy with first6 orders reflections

Fig. 3-13: Delay-spread model accuracy contour considering first 6 orders reflections (%)



Fig. 3-14: Delay-spread model accuracy contour considering first 7 orders reflections (%)



Contour of Delay Spread Estimation Accuracy with first8 orders reflections

Fig. 3-15: Delay-spread model accuracy contour considering first 8 orders reflections (%)



Fig. 3-16: Delay-spread model accuracy contour considering first 9 orders reflections (%)

3.5 Discussion on the Model Error and Calibration Method

As discussed in section 3.2, the computational complexity increases considerably with number of reflection orders included. It is reasonable that many researchers neglect high-order reflections to enable the model computation be conducted in a practical amount of time. As researchers are working on faster OWC systems than ever before, this sacrifice of accuracy for efficiency results in more and more significant performance errors. A reasonable and feasible solution to keep both accuracy and efficiency is applying calibration. We propose a calibration method based on the statistic data of the model accuracy curves. The model accuracy curves of all 841 locations are drawn in Fig. 3-17. By averaging them, we obtain the general calibration curve c[k]. It provides the correction value for each order of reflections. For the delay-spread calculated from first k orders of reflections, the value can be calibrated by:

$$\tilde{s}_j^{(k)} = s_j^{(k)} / c[k] \quad (3-10)$$

where $\tilde{s}_{j}^{(k)}$ and $s_{j}^{(k)}$ are the post-calibrated and pre-calibrated delay-spread from first *k* orders of reflections, respectively.

IABLE 3-2: CALIBRATION VALUES FOR CHANNEL MODEL						
Orders	1	2	3	4	5	6
Calibration	60.53	74.53	84.71	91.77	95.52	97.67
(%)						
Orders	7	8	9	10	11	12
Calibration	98.79	99.38	99.69	99.84	99.92	99.96
(%)						

The calibration value for each order of reflections is given in Table 3-2

We apply the RMS model error to compare the performance before and after calibration as in Fig. 3-17. These are calculated by (3-11) and (3-12), respectively:

$$e^{(k)} = \frac{1}{N} \sum_{j=1}^{N} \left(\frac{s_j^{(k)} - s_j}{s_j} \right)^2 \quad (3-11)$$
$$\tilde{e}^{(k)} = \frac{1}{N} \sum_{j=1}^{N} \left(\frac{\tilde{s}_j^{(k)} - s_j}{s_j} \right)^2 \quad (3-12)$$

where *N* is the total number of channels tested; $e^{(k)}$ is the pre-calibrated estimation error; $\tilde{e}^{(k)}$ is the post-calibrated estimation error; s_j is the reference of accurate estimation. As we can see, there is a substantial decrease in delay-spread model error after calibration. In the model applying first three orders of reflections, the average RMS error drops from 15.7% to 4.3%. We draw the contour for the calibrated model accuracy as in Fig. 3-19 and Fig. 3-20. It can be discovered that the order number of reflections needed for 1Gbps systems reduces from 5 to 3 and from 9 to 7 for 10Gbps systems.

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Fig. 3-17: Channel estimation accuracies for all 841 locations and average



Comparison of Delay-Spread Estimation RMS Error Before and After Calibration

Fig. 3-18: Comparison of average RMS model error before and after calibration



Fig. 3-19: Calibrated delay-spread model accuracy contour considering first 3 orders reflections (%)



Contour of Delay Spread Estimation Accuracy with first7 orders reflections

Fig. 3-20: Calibrated delay-spread model accuracy contour considering first 7 orders reflections (%)

3.6 Conclusions

This chapter extensively discusses the impact of high-order light reflections on IOW channel model error. The results indicate that conventional channel models based on a first few, most commonly three, orders of reflections are not able to provide sufficient model accuracy for contemporary research over Gbps high speed IOW systems. The spatial distribution of model error and the impact of each additional order are explored in details. Based on their findings, the authors develop a calibration approach. It successfully mitigates RMS delay-spread error and keeps efficiency simultaneously.

Chapter 4 Bit Error Rate (BER) Distribution and Outage

4.1 Introduction

BER distribution and outage is a critical measure of Optical Wireless Communications (OWC) performance. It is impacted by source lay-out, room structure, receivers' locations and other factors. In the communications process, sources continuously emit light pulses in Lambertian pattern. Due to the room structure, a transmitted pulse undergoes multiple consecutive reflections on the surfaces, until it arrives at the receiver or attenuates to a negligible energy. Since plural light sources, in particular layout, are commonly used in a room for providing sufficient illumination, the indoor optical wireless channel is a spatially diverse channel to a specific user in the room. Due to the multi-path effect by reflections, and the intersource interference, the channel causes temporal distortion of the impulse. As the distortion strongly relies on the receiver location, the system will exhibit significantly different performance (BER) when the receiver is placed at different locations. In this chapter, we consider all possible users' locations in the room and apply MIMO channel model to simulate BER distribution and outage.

4.2 Indoor Visible Light Communication System

The simulation assumes that there are multiple light sources on the room ceiling in a specific layout format. All these sources broadcast the same information. The floor is divided into many elements, while each of the elements is a possible receiver location, representing a possible user. For a particular receiver location, we generate the impulse response from the sources to the

receiver at that location. Then, run a data simulation based on the obtained impulse response to calculate BER. The simulation is carried for Intensity Modulation / Direct Detection (IM/DD). We repeat the process at other locations and obtain the BER of all possible user locations. Based on the BER data collected, we calculate the percentage of the room area that does not meet the outage requirement as the outage probability.

In practice, for mathematical convenience, we change the order slightly by using Kavehrad's method, as referred to in Chapter 2. We gather all the impulse responses at all locations at first in a Matrix H(t) as step 1; then we run the data simulations together to get the BERs in a Matrix form as step 2; and calculate the BER outage as step 3. The sequential change leads to the same result as we desire.

4.2.1 Indoor Wireless Channel Modeling

The theories of the modeling method are presented in detail in Chapter 2. We list the parameters in our simulation in Table 4-1.

The parameters are chosen for the following reasons, respectively. For convenience, we set the transmission power of each light source to 680lm, which equals to 1 Watt at 555 nm [28]. Receiver FOV is set to a large value as 60 degrees to enlarge receiving area as well as reducing blocking probability. LED half power angle is set to 60 degrees as the Lambertian mode number is equal to 1, which is the common value of most commercial LEDs [4]. Reflection coefficients for ceiling, walls and floor are from Kavehrad's previous research [3]. The room dimension is selected as that of an ordinary room.

TABLE 4-1. WIRELESS CHANNEL MODELTA	NAMETERS
Transmission power (lm)	680
FOV (degree)	60
LED half power angle (degree)	60
Reflection coefficients	
(ceiling, wall, floor)	0.9, 0.5, 0.1
Room dimension (m*m*m)	6*6*3
Reflection element size (m*m)	0.1*0.1
Time resolution (ns)	0.33

TADLE A 1. WIDELESS CHANNEL MODEL DADAMETERS

Suppose there are *n* possible receiver locations. After the modeling, the impulse response matrix is obtained as $H_{n \times 1}(t)$. $H_{n \times 1}(t)$ is an $n \times 1$ matrix, for which each entry represents the impulse response from the sources to one location.

4.2.2 Communication Simulation

The communication arguments used are demonstrated in Table 4-2. Though it is higher than the maximum modulation rate of most available commercial LEDs, the computer simulation bit rate is set to 600Mbps for two reasons. First, we focus our research on channel and it is reasonable to idealize the source; second, by some modern coding and equalization approaches, the available transmission rate of commercial LED is approaching the value that we use [8]. The layouts of 4, 6, 9 sources are typical. d[m] is the test data sequence. s(t) is the transmission waveform, which is generated by OOK modulation of d[m]. $r_{n \times 1}(t)$ is the receiving waveform vector, while each entry is the received waveform at a specific location. Each entry of $r_{n \times 1}(t)$ comes from the convolution of s(t) and corresponding entry of $H_{n \times 1}(t)$ plus noise. It can be presented as

$$r_{n \times 1}(t) = H_{n \times 1}(t) \otimes s(t) + n(t)$$
 (4-1)

where \otimes means convolution and n(t) represents the noise.

Bit rate (Mbps)	600
noise (dbmW)	-105
Test data length	1e6

4.2.3 BER Outage Evaluation

Recovering data $\hat{d}_{n\times 1}[m]$ from $r_{n\times 1}(t)$ by sampling and decision making, we compare it with $d_{n\times 1}[m]$ to obtain error vector $e_{n\times 1}[m]$

$$e_{n \times 1}[m] = XOR(d_{n \times 1}[m], d_{n \times 1}[m])$$
 (4-2)

where *XOR* stands for Exclusive-OR to each pair of entries. Each entry of $e_{n\times 1}[m]$ is a "1-0" sequence where a "1" represents an error. Counting the number of "1"s in each entry of $e_{n\times 1}[m]$ and dividing it by the length of corresponding entry, we obtain the BER vector $ber_{n\times 1}$. Each entry of $ber_{n\times 1}$ is the BER of corresponding location. We set an acceptance threshold for BER for

indoor wireless communication, for example, 10^{-5} . When an entry of $ber_{n\times 1}$ is greater than the threshold, it means that the corresponding location is in the outage state. The BER outage probability of the room is calculated by:

$$outage = \frac{entries \ number \ of \ (ber_{n \times 1} > threshold)}{n}$$
(4-3)

4.3 Optimal Detection and BER Outage Analysis

In this section, an analytical method is used to predict VLC BER distribution and outage caused by inter-source interference under arbitrary source layout. To the best of the author's knowledge, there have been no research publications on this topic. A few investigations were about power distribution of VLC system, among which Wu and Kavian's work [29] was distinguished. Nevertheless, the analytical method to solve BER distribution is significantly more complex than power distribution because diffusion and lights overlapping increase BER in certain scenarios while decrease BER in other scenarios. Therefore, Barry's and Kavehrad's methods are generally applied to calculate impulse response from the sources to specific receiver location and test data is then used to compute the BER at this location. The process is repeated for all locations to obtain BER distribution in a room.

As mentioned in Chapter 2, the simulation method has a high computational complexity and requires a large volume of processing resources, especially for high resolution room model. That is because the method considers the Lambertian transmission between any pair of reflectors in each tier of diffusion; when room resolution increases by n times, the computational complexity increases by $n^2 \times n^2 = n^4$ times. The test data may also consume a long time to process when high accurate BER is needed. The analytical method proposed in this chapter is based on Lambertian transmission, geometrical computing and optimal detection theory.

4.3.1 Optimal Detection

As Lambertian transmission is described in Chapter 2 and the geometrical computing will be shown in the next section, we provide a brief introduction of optimal detection. On-Off Keying (OOK) is the most popular VLC modulation scheme, which is basically binary amplitude modulation; therefore, we take binary antipodal signaling as an example to explain optimal detection [30].

In a binary antipodal signaling scheme $s_1(t) = s(t)$ and $s_2(t) = -s(t)$. The probabilities of messages 1 and 2 are p and 1-p, respectively. The vector representations of the two signals are just scales with $s_1(t) = \sqrt{\varepsilon_s}$ and $s_2(t) = -\sqrt{\varepsilon_s}$, where ε_s is the energy in each signal. The decision region of signal 1 D_1 is given as

$$D_{1} = \{r : r\sqrt{\varepsilon_{s}} + \frac{N_{0}}{2} \ln p - \frac{1}{2}\varepsilon_{s} > -r\sqrt{\varepsilon_{s}} + \frac{N_{0}}{2} \ln(1-p) - \frac{1}{2}\varepsilon_{s}\}$$

$$= \{r : r > \frac{N_{0}}{4\sqrt{\varepsilon_{s}}} \ln \frac{1-p}{p}\}$$

$$= \{r : r > r_{tb}\}$$

$$(4-4)$$

where $\frac{N_0}{2}$ is the variance of Gaussian noise and r_{th} is the threshold defined as

$$r_{th} = \frac{N_0}{4\sqrt{\varepsilon_s}} \ln \frac{1-p}{p} \qquad (4-5)$$

To derive the error probability for this system, we have

$$P_{e} = \sum_{m=1}^{2} P_{m} \sum_{\substack{1 \le m' \le 2 \\ m' \ne m}} \int_{D_{m'}} p(r \mid s_{m}) dr$$

$$= p \int_{D_{2}} p(r \mid s = \sqrt{\varepsilon_{s}}) dr + (1 - p) \int_{D_{1}} p(r \mid s = -\sqrt{\varepsilon_{s}}) dr$$

$$= p \int_{-\infty}^{r_{th}} p(r \mid s = \sqrt{\varepsilon_{s}}) dr + (1 - p) \int_{r_{th}}^{\infty} p(r \mid s = -\sqrt{\varepsilon_{s}}) dr \qquad (4-6)$$

$$= p P[N(\sqrt{\varepsilon_{s}}, \frac{N_{0}}{2}) < r_{th}] + (1 - p) P[N(-\sqrt{\varepsilon_{s}}, \frac{N_{0}}{2}) > r_{th}]$$

$$= p Q(\frac{\sqrt{\varepsilon_{s}} - r_{th}}{\sqrt{\frac{N_{0}}{2}}}) + (1 - p) Q(\frac{r_{th} + \sqrt{\varepsilon_{s}}}{\sqrt{\frac{N_{0}}{2}}})$$

In the special case where $p = \frac{1}{2}$, we have $r_{th} = 0$ and the error probability simplifies to

$$P_e = Q(\sqrt{\frac{2\varepsilon_s}{N_0}}) \qquad (4-7)$$

Suppose there is an interference signal at next symbol period with energy ε_i , the error probability becomes

$$P_e = \frac{1}{2}Q(\sqrt{\frac{2}{N_0}}(\sqrt{\varepsilon_s} - \sqrt{\varepsilon_i})) \quad (4-8)$$

where the coefficient $\frac{1}{2}$ indicates that the interference symbol is different from the transmitted symbol by the probability of $\frac{1}{2}$.

4.3.2 BER Analysis

As referred in Chapter 2, there are two kinds of impulse distortions in VLC, which are multi-path effect and the inter-source interference. In most room areas, inter-source impulse distortion dominates. It can be modeled by multiple LOS Lambertian transmissions from different sources. The inter-source impulse distortion model can be further used to determine room BER distribution.

Suppose there are N sources $S_1, S_2, ..., S_N$ on the ceiling, streaming data to receiver R. The coordinate of source *i* is (s_{ix}, s_{iy}, s_{iz}) (i = 1, 2, ..., N) and the coordinate of the receiver is (r_x, r_y, r_z) . We define the distance matrix *HD* and the power matrix *HP* as

$$HD = [d_1, d_2, \dots d_N] \quad (4-9)$$
$$HP = [p_1, p_2, \dots p_N] \quad (4-10)$$

where d_i and p_i indicate the distance and the received power from source S_i (i=1,2,...,N) to the receiver R, respectively. Applying geometric method, it is straightforward to obtain d_i as

$$d_{i} = \left\| (s_{ix}, s_{iy}, s_{iz}) - (r_{x}, r_{y}, r_{z}) \right\| (i = 1, 2, ..., N)$$
(4-11)

Using LOS Lambertian model (2-1), we calculate p_i as

$$p_{i} = \frac{n+1}{2\pi} \cos^{n}(\phi_{i})(\cos(\theta_{i})A_{Ri}/d_{i}^{2})rect(\theta_{i}/FOV)$$

$$= \frac{n+1}{2\pi}(\cos^{n+1}(\phi_{i})A_{Ri}/d_{i}^{2})rect(\theta_{i}/FOV)$$

$$= \frac{n+1}{2\pi}((\frac{s_{zi}-r_{z}}{d_{i}})^{n+1}A_{Ri}/d_{i}^{2})rect(\theta_{i}/FOV)$$

$$= \frac{n+1}{2\pi}A_{Ri}(s_{iz}-r_{z})^{n+1}d_{i}^{-3-n}rect(\theta_{i}/FOV)$$
(4-12)

We consider LOS transmission from source S_i (i=1,2,...,N) to receiver R, since ceiling plane is parallel to floor plane, we have receiver incident angle θ_i equals to source transmission angle ϕ_i . Then, $\cos(\phi_i)$ can be calculated by trigonometric relation as

$$\cos(\phi_i) = \frac{(s_{iz} - r_z)}{d_i} \qquad (4-13)$$

In this chapter, we assume Lambertian pattern order n = 1, incident angle θ_i is always smaller than receiver *FOV* and all receivers are of the same area A_R . As a result, we have

$$p_i = \frac{1}{\pi} \frac{A_R (s_{iz} - r_z)^2}{d_i^4} \qquad (4-14)$$

From *HD* and *HP*, we know that the impulse response h(t) from the sources $S_1, S_2, ..., S_N$ to receiver *R* consists of *N* peaks, with different amplitudes arriving at different times as

$$h(t) = \sum_{i=1}^{N} \frac{1}{\pi} \frac{A_R (s_{iz} - r_z)^2}{d_i^4} \delta(t - \frac{d_i}{c}) \qquad (4-15)$$

As in (3-1), suppose transmission signal is s(t), combining with (4-12), the received signal is

$$r(t) = h(t) \otimes s(t) + n(t)$$

= $\frac{A_R}{\pi} \sum_{i=1}^{N} \frac{(s_{iz} - r_z)^2}{d_i^4} s(t - \frac{d_i}{c}) + n(t)$ (4-16)

We define the earliest arrived light impulse as the user impulse and suppose it comes from source S_0 . It satisfies $S_0 = \underset{1 \le i \le N}{\arg \min(d_i)}$.

When the peaks arrive after the user peak no later than a symbol period T, as s(t) remains unchanged, these peaks will enforce the user peak. Therefore, the received user energy is

$$\varepsilon_{s} = \int_{0}^{T} \left(\frac{A_{R}}{\pi} \sum_{\substack{i \\ 0 < d_{i} - d_{0} \le cT}} \frac{(s_{iz} - r_{z})^{2}}{d_{i}^{4}} s(t - \frac{d_{i}}{c})\right)^{2} dt \qquad (4-17)$$

We consider the peaks arriving after the user peak between T and 2T, and ignore the later peaks, since they are significantly weak. The interference optical energy is

$$\varepsilon_{i} = \int_{T}^{2T} \left(\frac{A_{R}}{\pi} \sum_{\substack{i \ cT < d_{i} - d_{0} \le 2cT}} \frac{(s_{iz} - r_{z})^{2}}{d_{i}^{4}} s(t - \frac{d_{i}}{c})\right)^{2} dt \quad (4-18)$$

Since by most optical detection approaches, optical energy will be converted to current for processing, the optical energy is proportional to square root of signal energy; therefore, we square each term of optical energy to calculate signal energy. Suppose binary antipodal modulation is used. By (4-8) the BER at that receiver location is

$$P_{e} = \frac{1}{2} Q(\sqrt{\frac{2}{N_{0}}} (\sqrt{\varepsilon_{s}} - \sqrt{\varepsilon_{i}}))$$

$$= \frac{1}{2} Q(\frac{A_{R}}{\pi} \sqrt{\frac{2}{N_{0}}} (\sqrt{\int_{0}^{T} (\frac{A_{R}}{\pi} \sum_{\substack{i \\ 0 < d_{i} - d_{0} \le cT}} \frac{(s_{iz} - r_{z})^{2}}{d_{i}^{4}} s(t - \frac{d_{i}}{c}))^{2} dt} - \sqrt{\int_{T}^{2T} (\frac{A_{R}}{\pi} \sum_{\substack{i \\ cT < d_{i} - d_{0} \le 2cT}} \frac{(s_{iz} - r_{z})^{2}}{d_{i}^{4}} s(t - \frac{d_{i}}{c}))^{2} dt}))$$

$$(4-19)$$

The coefficient $\frac{1}{2}$ means the interference symbol is different from user symbol by probability of 50%. Repeat the process for all locations and we can calculate the BER distribution and outage.

4.4 Simulation Results (Receiver FOV = 60°)

When the light pulse travels in indoor environment, it experiences reflections. When the number of these reflections is large, multi-path effect becomes significant and the pulse spreads in time domain. If multiple sources are applied, inter-source interference occurs. The interfering

sources produce additional delayed pulses to the ideal impulse response. We conduct our simulations through the following steps:

a. Locate *l* sources on the ceiling in specific layout for spatial diversity.

b. Generate indoor VLC spatial diversity matrix $H_{n\times 1}(t)$ by Kavehrad's method.

c. Receiver plane is uniformly divided into n elements as receiver locations. Each row of $H_{n \times 1}(t)$ refers to the impulse response from sources to certain receiver.

d. Generate test data d[m] and corresponding waveform s(t) by OOK modulation.

e. Calculate received waveform $r_{n \times 1}(t)$ by $r_{n \times 1}(t) = H_{n \times 1}(t) \otimes s(t) + n(t)$. \otimes means convolution and n(t) is Gaussian noise.

f. Recover data $\hat{d}_{n \times 1}[m]$ from $r_{n \times 1}(t)$ by sampling and decision making, where each element of $\hat{d}_{n \times 1}[m]$ is the received data at each receiver location.

g. Compare each element of $\hat{d}_{n\times 1}[m]$ with $d_{n\times 1}[m]$ to obtain error vector $e_{n\times 1}[m]$.

$$e_{n \times 1}[m] = XOR(\hat{d}_{n \times 1}[m], d_{n \times 1}[m])$$
 (4-17)

h. Each entry of $e_{n\times 1}[m]$ is a "1-0" sequence where a "1" represents an error. Counting the number of "1"s in each entry of $e_{n\times 1}[m]$, we estimate the BER vector $ber_{n\times 1}$. Each entry of $ber_{n\times 1}$ is the BER of corresponding location.

i. We set an acceptance threshold for BER for indoor wireless communication, for example, 10^{-5} . When an entry of $ber_{n\times 1}$ is greater than the threshold, it means that the corresponding location is in an outage state. Computing outage probability by:

$$outage = \frac{entries \ number \ of \ (ber_{n \times 1} > threshold)}{n} \quad (4-18)$$

The parameters used are in Table 4-3 below.

4.4.1 BER Distribution and Outage

Fig. 4-1 - Fig. 4-6 show the BER distribution and outage of 4-sources, 6-sources and 9sources, respectively. In these figures, X and Y values indicate the position of a receiver on the floor; Z value is the corresponding logarithmic BER. The higher BER in a location, the worse communication quality is expected for the user located there. X-Y views of the figures clearly demonstrate the BER distribution on the floor. Setting a BER threshold, for instance, 10^{-5} as we do, we can easily calculate the BER outage for the room in a specific source layout.

TABLE 4-3: BER DISTRIBUTION AND OUTAG	TABLE 4-3: BER DISTRIBUTION AND OUTAGE SIMULATION PARAMETERS				
Transmission power (lm)	680				
FOV(degree)	60				
Data length	1e6				
LED half power angle (degree)	60				
Room dimension (m*m*m)	6*6*3				
Reflection coefficients					
(ceiling, wall, floor)	0.9, 0.5, 0.1				
Bit rate (Mbps)	600				
Shot noise (dbmW)	-105				
4 sources locations (m,m,m)	(1.5, 1.5, 3), (1.5, 4.5, 3),				
	(4.5, 1.5, 3), (4.5, 4.5, 3)				
6 sources locations (m,m,m)	(1, 1.5, 3), (3, 1.5, 3),				
	(5, 1.5, 3), (1, 4.5, 3),				
	(3, 4.5, 3), (5, 4.5, 3)				
9 sources locations (m,m,m)	(1, 1, 3), (3, 1, 3), (5, 1, 3),				
	(1, 3, 3),(3, 3, 3), (5, 3, 3),				
	(1, 5, 3), (3, 5, 3),(5, 5, 3)				

Comparing Fig. 4-1 through Fig. 4-6, we find that a high BER exists in the sources overlapping areas and these areas enlarge as the number of sources increases. The results can be explained here. When the number of sources increases, the sources layout becomes denser and there are more illumination overlapping areas on the floor; in the overlapping areas, communication is subject to inter-source interference, and the impulse response from sources at these locations will be distorted more significantly. The distortion of impulse response will cause ISI, and thus a high BER. The simulation demonstrates that dense source layout will increase BER outage for a VLC system.

Another interesting observation is that there are less high-BER regions near room corners. This holds as layout density increases. The result shows that the interference from multi-path effect is insignificant on BER outage. Though multi-path impulse response degrades with increasing number of sources, the BER outage is not significantly impacted by multi-path impulse response spread.

4.4.2 Impulse Response Distortion

By computer simulations, we may work out the total impulse response from source to any possible receiver locations. In most locations, the impulse response has little impulse spread. However, we still find that there are two kinds of locations exhibiting considerable impulse distortion as shown in Fig. 4-7: the first location is the room corner (location A (0.2, 0.1)); the second location is the overlapping area of light footprints (location B (2.1, 2.4)).



Fig. 4-1: BER distribution and outage of 4 sources by simulation, FOV = 60°



600Mbps,4 sources, BER outage:0.03476

Fig. 4-2: BER distribution and outage of 4 sources by simulation, $FOV = 60^{\circ}$ (Top View)



Fig. 4-3: BER distribution and outage of 6 sources by simulation, FOV = 60°



600Mbps,6 sources, BER outage:0.40649

Fig. 4-4: BER distribution and outage of 6 sources by simulation, $FOV = 60^{\circ}$ (Top View)



Fig. 4-5: BER distribution and outage of 9 sources by simulation, FOV = 60°



600Mbps,9 sources, BER outage:0.63459

Fig. 4-6: BER distribution and outage of 9 sources by simulation, $FOV = 60^{\circ}$ (Top View)

At the room corner, the spread of impulse response comes from multi-path effect of light reflections. In the corner area, the receiver captures reflected lights, which experience different reflection paths, and cause the decreasing tail in impulse response. What is worth mentioning is that the tail has much lower power comparing with the peak. It can be shown later that this kind of distortion exhibits less significant influence on VLC.

In the overlapping areas, we can see two high sharp peaks in the impulse response. The reason for causing the multiple peaks is that lights from different sources enter receiver via different Line-of-Sight (LOS) paths, the time difference of the arrivals causes the multiple peaks. This is an important influencing factor for VLC.



Fig. 4-7: Impulse response distortion by (a) multi-path effect (b) inter-source interference (FOV = 60°)

4.5 Simulation Results (Receiver FOV = 30°)

To further explore the impact of inter-source interference to BER distribution, we repeated the simulation with receiver FOV = 30° . The results are given as Fig. 4-8 – Fig. 4-13. We find that the BER outage reduce from 0.4065 to 0.1709 in 6-sources layout and from 0.6346 to 0.3453 in 9-sources layout, respectively. That is because in many locations, small receiver FOV refuses much interference power and reduces BER; therefore, BER outage decreases. This effect can be clearly observed by comparing Fig. 4-7 with Fig. 4-14. They are impulse responses at same locations, with different FOVs. In Fig. 4-14, most inter-source interference peaks are filtered out. An interesting finding is that in 4 sources scenario, BER outage of FOV = 30° is higher than that of FOV = 60° . That is because small FOV also decreases total received power. When the amount of sources is small and small FOV is applied, the BER may rise due to insufficient Signal-to Noise Ratio (SNR), in certain locations. This point should be paid attention to, when we want to reduce inter-source interference by utilizing small FOV receivers.





Fig. 4-8: BER distribution and outage of 4 sources by simulation, $FOV = 30^{\circ}$



Fig. 4-9: BER distribution and outage of 4 sources by simulation, $FOV = 30^{\circ}$ (Top View)



600Mbps,6 sources, BER outage:0.17093

Fig. 4-10: BER distribution and outage of 6 sources by simulation, $FOV = 30^{\circ}$


Fig. 4-11: BER distribution and outage of 6 sources by simulation, $FOV = 30^{\circ}$ (Top View)



600Mbps,9 sources, BER outage:0.3453

Fig. 4-12: BER distribution and outage of 9 sources by simulation, $FOV = 30^{\circ}$



Fig. 4-13: BER distribution and outage of 9 sources by simulation, $FOV = 30^{\circ}$ (Top View)

4.6 Analytical Results and Comparisons

As shown in section 4.2.3, the BER caused by inter-source interference can be found by equation (4-17) to (4-19). By this method, we can analytically calculate the BER distribution and outage. The results of 4 sources and 6 sources of FOV = 60° are shown in Fig. 4-15 - Fig.4-18. Comparing with Fig. 4-1 - Fig. 4-4, we can see a good match in BER distribution and outage.



Fig. 4-14: Impulse response distortion by (a) multi-path effect (b) inter-source interference (FOV = 30°)



Fig. 4-15: BER distribution and outage 4 sources by analytical method of, $FOV = 60^{\circ}$



Fig. 4-16: BER distribution and outage of 4 sources by analytical method, $FOV = 60^{\circ}$ (Top View)



Fig. 4-17: BER distribution and outage of 6 sources by analytical method, $FOV = 60^{\circ}$



Fig. 4-18: BER distribution and outage of 6 sources by analytical method, $FOV = 60^{\circ}$ (Top View)

4.7 Conclusions

In this chapter, we investigate the possibility of VLC. This technology is a high speed, energy efficient and secure solution to RF band congestion. In general home environment, the major impact factors to VLC are multi-path effect and inter-source interference, which degrade communications performance by causing impulse response distortion. To further research these factors, at first, we establish the indoor optical wireless model by tracking light pulses experiencing reflections. After the channel model is obtained, we use data simulation to statistically calculate BER distribution and outage. By observing the results, we find that multipath effect exists at room corner locations, and inter-source interference exists at the overlapping area of light footprints. Moreover, the influence of inter-source interference is more significant than multi-path effect.

At the beginning of this chapter, we explore the methods to create indoor optical wireless channel. Lambertian emitting pattern, pulse tracking, SISO channel model and MIMO channel model are described. Next, we develop the simulation method to calculate BER distribution and outage on the base of channel model. After that, we propose an analytical method to predict BER distribution and outage, by Lambertian transmission theory, geometric method, and optimal receiver principle. In the end, we demonstrate the simulation results for BER distribution and outage. They are compared with the analytical results.

Chapter 5

Orthogonal Frequency-Division Multiplex (OFDM) for Indoor Optical Wireless Communications

5.1 Introduction

The Wireless Home Link (WHL) is being paid more and more attention nowadays. In next decades, wireless communication will play a significant role in electronic devices interconnections [31]. Users of WHL can access Internet anywhere.

Indoor Optical Wireless (IOW) is the technique which applies optical wireless in computer networks and/or office equipment communications. It will be a strong candidate of WHL, as it satisfies the high speed wireless communication proposed by WHL. Besides, it has other advantages: IOW is suitable for non-public networks as no authority is required; it is also resistant to spatial interference as light power is spatially constrained; In addition, it doesn't take restricted Radio Frequency (RF) band resources and can be applied in strict RF prohibited areas such as hospitals, aircrafts, power stations, mines and other areas. Therefore, IOW is a better solution than RF in WHL.

In recent years, the illumination technology based on Lighting Emitting Diode (LED) is developing fast because of its energy saving advantages. LED has features of high light intensity, high converting efficiency, and long lifetime. It is generally considered as the next generation light source and will in the end replace the universal incandescent bulb and fluorescent bulb in home and work places. As LED has a short response time, we can modulate it by user data to create illumination and communication dual functional "Base Station Light." It is a completely new access method of optical communications. Visible Light Communications (VLC) is an emerging wireless communication technology based on white LED. White LED has many advantages such as long lifetime, low power consumption, small size and environment friendliness. LED also has high response sensitivity so that it is competent for high speed communications. In VLC, LED can take both communication and illumination duties. That is because of its high light intensity as well as high modulation speed. Human can hardly sense the flickers.

The combination of Intensity Modulation/Direct Detection IM/DD system, Manchester Code and On-Off-Keying (OOK) modulation is the most popular VLC setup due to its simplicity. In IM/DD system, as multiple light sources are applied, every receiver can capture the light signal from multiple directions. As a result, blocking on one or several light paths will not break down the communication.

Comparing with RF, VLC has following advantages:

1). Visible light is safe to human while VLC transmits data via visible light.

2). VLC is everywhere. Since illumination infrastructures are widely installed. People can realize high speed wireless communication conveniently by these infrastructures.

3). VLC can have high transmitting power. The infrared (IR) has a strict transmitting power constrain due to eye safety consideration. The RF signal also has a power constrain considering health issues. In VLC, as visible light is utilized, the transmitting power can be high.

4). No band authorization is needed. As band resource is limited, people are facing band shortage problem.

5). No electromagnetic interference is generated. The VLC can be applied in hospitals, aircrafts, power stations, mines and other RF sensitive environments.

The OFDM, due to its effectiveness in solving Inter-symbol Interference (ISI) caused by dispersive channel, is widely applied in broadband wired and wireless communications [32][33]. There are two major advantages of OFDM. First, when conventional serial modulation methods are used, equalization complexity rises rapidly as data rate increases. However, the complexity of OFDM scales well as data rate and dispersion increase. Second, OFDM transfers the transmitters and receivers complexities from analog domain to digital domain. In OFDM, any phase variations with frequency can be corrected at little cost in digital parts of the receiver. Upon the advantages described above, OFDM can be a good modulation scheme for VLC. However, it is only recently that it has been considered.

5.2 OFDM Overview

5.2.1 Principle of OFDM

OFDM is a multi-carrier modulation technique. Suppose f_k (k = 1, 2, ..., N) are frequencies of N sub-carriers, the multi-carrier modulated signal in the i - th symbol interval can be expressed as [32][33]:

$$s_i(t) = \sum_{k=0}^{N-1} X_i(k,t) \exp(j2\pi f_k t)$$
 (5-1)

 $X_i(k,t)$ is the data carried in the i-th interval which determines the amplitude and phase of $s_i(t)$. When only one symbol interval is considered, the subscripts can be omitted. Also, when quadrature amplitude modulation (QAM) or M-phase-shift keying (MPSK) modulation is applied, $X_i(k,t)$ is independent from time t. Therefore, (5-1) can be simplified, without ambiguity, as:

$$s(t) = \sum_{k=0}^{N-1} X(k) \exp(j2\pi f_k t) \quad (5-2)$$

We want the multi-carrier modulation to be bandwidth efficient and low cost. Normally, it requires N oscillation sources and relevant band pass filters, which are expensive. However, the multi-carrier modulation and demodulation can be economically and efficiently realized by famous fast Fourier transform (FFT). Because of its simplicity and bandwidth efficiency, OFDM is paid more and more attention.

In order to find the frequency intervals of the sub-carriers, we consider the demodulation method. Suppose we sample the received signal by sampling frequency f_s and demodulate by N points discrete Fourier transform (DFT), given that no noise or ISI. The k-th frequency component is:

$$S(k\Delta f) = \sum_{n=0}^{N-1} s(n / f_s) \exp(-j2\pi nk / N)$$
 (5-3)

where $S(k\Delta f)$ is the k-th frequency component; $s(n/f_s)$ (n=0,1,2,...,N-1) are the sampled signals; $\Delta f = f_s/N$ is the DFT frequency resolution. To calculate the frequency spectrum correctly by DFT, the signal must be periodic with period of N. This can be satisfied when the signal contains only harmonics in the DFT. Substitute $t = n/f_s$ into (5-2):

$$s(n/f_s) = \sum_{k=0}^{N-1} X(k) \exp(j2\pi f_k n/f_s) \quad (5-4)$$

Substitute (5-4) into (5-3),

$$S(k\Delta f) = \sum_{n=0}^{N-1} \sum_{k=0}^{N-1} X(k) \exp(j2\pi f_k n / f_s) \exp(-j2\pi n k / N)$$

= $\sum_{k=0}^{N-1} X(k) \sum_{n=0}^{N-1} \exp(j2\pi f_k n / f_s) \exp(-j2\pi n k / N)$ (5-5)
= $\sum_{k=0}^{N-1} X(k) \delta(\frac{f_k}{f_s} - \frac{k}{N})$

where $\delta(m,n) = \begin{cases} 0, & m \neq n \\ 1, & m = n \end{cases}$

From (5-5), we discover that when the frequency of modulated signal is

$$f_k = \frac{kf_s}{N} \quad (5-6)$$

 $S(k\Delta f) = CX(k)$ holds where C is a constant. It shows that when the sub-carrier frequency

 f_k is multiple of DFT resolution $\frac{f_s}{N}$, we can use DFT for demodulation. In addition, to ensure a correct demodulation, X(k) is necessarily being a constant.

Specially, if the carrier frequency interval is $\frac{f_s}{N}$, by (5-4), we have:

$$s(n/f_s) = \sum_{k=0}^{N-1} X(k) \exp(j2\pi (kf_s / N)n/f_s)$$

=
$$\sum_{k=0}^{N-1} X(k) \exp(j2\pi n/N)$$
 (5-7)

(5-7) is exactly the inverse discrete Fourier transform (IDFT) of X(k) (k = 0,1,2,..., N-1). It means that when sub-carrier interval is $\frac{f_s}{N}$, the time series of multi-carrier modulated signal can be generated by IDFT.

Multi-carrier modulation can be realized by IDFT and demodulation can be realized by DFT. They both can be operated via high efficient FFT.

5.2.2 Realization of OFDM



The OFDM system block diagram is shown by Fig 5-1.

Fig. 5-1: OFMD system block diagram

The user bit series experiences serial-to-parallel (S/P) conversion. Following the modulation method used, the parallel bits are mapped by modulation constellation. Then, carriers are modulated by the mapped bits to generate modulated data series X(N). The OFDM time domain sampled series is obtained by performing IDFT over X(N). After that, cyclic prefix (CP) is added to it. Next, the time domain wave is produced from the time domain sampled series by digital-to-analog (D/A) conversion and sent to channel. When it is captured by the receiver, by applying analog-to-digital (A/D) conversion and removing CP, the OFDM time domain sampled series is obtained. DFT transfers it back to X(N). After signal mapping and parallel-to-serial (P/S) conversion, we get the user bit series as output.

5.3 Clipping-noise Resistant OFDM schemes for IOW

5.3.1 OFDM and IOW

OFDM converts serial data transmission to parallel data transmission. It is an effective mitigation to indoor optical wireless multi-path distortion. When the symbol period of each subcarrier is long enough comparing to the channel delay-spread, ISI is significantly reduced [34]. IOW, however, uses intensity modulation. The major deficiency of OFDM in IOW is the requirement of high direct current (DC) bias to keep signal waveform nonnegative. Consequently, researchers in this area are interested in reducing DC bias [35].

Recently, several approaches have been proposed. The most straightforward method is DCclipped OFDM (DC-OFDM). It applies hard-clipping on a portion of the negative part of the signal waveform to reduce required DC bias level [36][37]. By this method, system performance is degraded because of the clipping noise. The second approach is named asymmetrically clipped optical OFDM (ACO-OFDM). It clips off the entire negative part of the waveform. By modulating only the odd subcarriers, the clipping noise can be avoided [38]. The third method is pulse-amplitude-modulated discrete multitone (PAM-DMT). This method also clips off the entire negative part; however, the clipping noise is avoided by modulating only the imaginary parts of the subcarriers [39]. Neither ACO-OFDM nor PAM-DMT degrades system performance. As, nevertheless, only half of the spectrum is utilized, the spectrum efficiencies of them are reduced to 50% of normal OFDM schemes.

5.3.2 DC-Clipped OFDM

The primary drawback of IM OFDM is the high DC requirement to guarantee waveforms non-negative. In addition, OFDM signals inherently have high PAPR. The two factors result in a wide dynamic range of waveform and cause critical difficulties in system implementation. DC-Clipped OFDM is a straightforward technique to reduce DC bias γ by performing hard-clipping on the negative part of the waveform. It is demonstrated in Fig. 5-2 [34].



Fig. 5-2: Block diagram of a DC-OFDM transmitter

As we can see from the figure, input symbols to the IDFT are enforced to be Hermitian symmetry. This enforcement ensures that the time-domain waveform is real. The zero sub-carrier, or DC bias, is not modulated. After D/A conversion, the waveform becomes nonnegative and signal is intensity modulated onto an optical carrier.

When the number of subcarriers is large, the electrical OFDM signal γ can be modeled as a Gaussian random variable. Its mean equals to the DC bias γ and variance equals to the electrical

power $\sigma^2 = E[|x(t)|^2]$. When DC-OFDM is executed, only the positive part of the waveform remains. In Appendix A of [34], the authors demonstrated the formula representing the average optical power of DC-OFDM as

$$P_{DC-OFDM} = E\left[x_{clip}(t)\right] = \frac{\sigma}{\sqrt{2\pi}}e^{-\frac{\gamma^2}{2\sigma^2}} + \gamma(1 - Q(\frac{\gamma}{\sigma})) \quad (5-8)$$

where Q is the Q-function defined as in [30]

$$Q(x) = P[N(0,1) > x] = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} e^{-\frac{t^{2}}{2}} dt \quad (5-9)$$

5.3.3 Asymmetrically Clipped Optical OFDM

ACO-OFDM was first proposed by Armstrong and Lowery in [40]. They proved that modulating only odd subcarriers prevents clipping noise. This technique is able to reduce DC bias γ to zero. A typical ACO-OFDM transmitter is shown in Fig. 5-3 [34].



Fig. 5-3: Block diagram of an ACO-OFDM transmitter

To guarantee a real waveform, as in other IM OFDM schemes, ACO-OFDM subcarriers should have Hermitian symmetry. Since ACO-OFDM encodes only odd subcarriers, it has 50% spectrum efficiency of regular IM OFDM methods. After D/A conversion, electrical signal is clipped at zero and then modulated onto an optical carrier.

For large number subcarriers, we can also use a Gaussian random variable to represent the electrical signal x(t). As ACO-OFDM clips DC bias to zero, the Gaussian variable has zero mean. In [40], the authors showed that the mean and variance of the waveform are $\frac{\sigma}{\sqrt{2\pi}}$ and $\frac{\sigma^2}{2}$, respectively. As a result, the mean of optical power is proportional to the square-root of the electrical power as

$$P_{ACO-OFDM} = \frac{\sigma}{\sqrt{2\pi}} \quad (5-10)$$

5.3.4 PAM-Modulated Discrete Multitone

Lee and Koonen proposed PAM-DMT in [39]. In this paper, they demonstrated that impairment from clipping noise can be avoided by applying PAM modulation only on the imaginary components of subcarriers. The reason is that clipping noise is real, thus is orthogonal to the data. Similar to ACO-OFDM, PAM-DMT is able to reduce DC bias to zero. PAM-DMT block diagram is given by Fig. 5-4.



Fig. 5-4: Block diagram of an PAM-DMT transmitter

PAM-DMT also has Hermitian symmetric subcarriers to ensure real waveforms. Being different from ACO-OFDM, PAM-DMT uses all subcarriers for symbol encoding; however, only on the imaginary components. PAM-DMT, therefore, has the same spectrum efficiency of ACO-OFDM, which is only half of DC-OFDM. After D/A conversion, DC bias is reduced to zero and the electrical signal is modulated onto an optical carrier.

In [39], the authors gave the mean transmitted optical power as

$$P_{ACO-OFDM} = \frac{\sigma}{\sqrt{2\pi}} \quad (5-11)$$

where σ is the square-root of the OFDM electrical power.

5.3.5 OFDM Precoding

Though DC-OFDM, ACO-OFDM and PAM-DMT are able to reduce DC bias of transmitted signal, the high peak-to-average power ratio (PAPR) is still an unsolved challenge in OFDM IOW systems. Several precoding techniques have been explored to decrease signal

PAPR. These techniques include discrete Fourier transform (DFT), Zadoff-Chu matrix transform (ZCT) and discrete cosine transform (DCT) [41][42].

A. DFT-OFDM

IDFT and DFT are used for digital modulation and demodulation, respectively in OFDM systems [43]. Assuming there are N_s subcarriers in the modulator, input data is grouped into a group of N bits, where $N = N_s \times m_n$ with m_n as the number of bits for a symbol in one subcarrier. Subcarriers are spaced by an integer multiple of the subcarrier symbol rate R_s to ensure the orthogonality between subcarriers. Subcarrier symbol rate R_s can be calculated from the transmission bit rate R_c by $R_s = R_c/N$ [41]. Consequently, the overall output OFDM signal can be expressed as:

$$X(t) = \sum_{n=0}^{N_{S}-1} C_{k} e^{2\pi j \left(n - \frac{N_{S}}{2}\right) t/T_{S}}$$
(5-12)

where C_k are complex values representing subcarrier symbols and T_S is the symbol period [44].

B. Zadoff-Chu Sequence and ZCT-OFDM

Zadoff-Chu (ZC) sequences are class of poly phase sequences with optimal correlation properties. They have ideal periodic autocorrelation and constant magnitude. The ZC sequences of length N can be expressed as [42]:

$$a_n = \begin{cases} e^{\frac{2\pi r j}{N} \left(\frac{k^2}{2} + qk\right)} & N \text{ is even} \\ e^{\frac{2\pi r j}{N} \left(\frac{k(k+1)}{2} + qk\right)} & N \text{ is odd} \end{cases}$$
(5-13)

where k = 0, 1, 2, ..., N - 1, q is any integer, r is any integer relatively prime to N. The ZCT can be defined by reshaping the ZC sequence by k = m + lL as:

$$A = \begin{bmatrix} a_{00} & a_{01} & \cdots & a_{0(L-1)} \\ a_{10} & a_{11} & \cdots & a_{1(L-1)} \\ \vdots & \vdots & \ddots & \vdots \\ a_{(L-1)0} & a_{(L-1)1} & \cdots & a_{(L-1)(L-1)} \end{bmatrix}$$
(5-14)

where *m* indicates the row variable and *l* indicates the column variable. As we can see, the kernel of ZCT is filled with a $N = L^2$ points long ZC sequence column-wise.



Fig. 5-5: Block diagram of ZCT precoded OFDM systems

Fig. 5-5 demonstrates the block diagram of ZCT precoded OFDM systems [42]. Assuming the parallel input data is $X = [X_0, X_1, X_2, ..., X_{L-1}]^T$, each data block is multiplied by *V* different phase factor with length *L*, $B^{(v)} = [b_{v,0}, b_{v,1}, b_{v,2}, ..., b_{v,L-1}]^T$ (v = 1, 2, ..., V). As a result, the *v*-th phase sequence is $X^{(v)} = [X_0 b_{v,0}, X_1 b_{v,1}, X_2 b_{v,2}, ..., X_{L-1} b_{v,L-1}]^T$ (v = 1, 2, ..., V).

C. DCT-OFDM

In DCT-OFDM, a single set of cosinusoidal function is used as the orthogonal basis instead of the complex exponential functions set. The multi-carrier modulation can be implemented by DCT [45]. The output signal of DCT-OFDM is

$$X(t) = \sqrt{\frac{2}{N_S}} \sum_{n=0}^{N_S - 1} d_n \beta_n \cos(\frac{n\pi t}{T_S}) \quad (5-15)$$

where d_n ($n = 1, 2, ..., N_S - 1$) are independent data symbols in a modulation constellation, and β_n is defined as

$$\beta_n = \begin{cases} \frac{1}{\sqrt{2}} & n = 0\\ 1 & n = others \end{cases}$$
(5-16)

5.4 Simulation of General OFDM Schemes for IOW

5.4.1 Simulation Setup

We place the receiver at 3 different locations, representing different levels of multipath distortions. OFDM performance is simulated and compared with OOK performance. The simulation parameters are given in Table 5-1.

Suppose d(t) is the Binary Phase-Shift Keying (BPSK) modulated user data sent at the transmitters, the transmitted signal s(t) can be expressed as:

$$s(t) = \begin{cases} d(t) & OOK\\ IFFT(d(t)) & OFDM \end{cases}$$
(5-17)

The transmission model is:

$$r_i(t) = s(t) \otimes h_i(t) + n(t)$$
 (i = 1,2,3) (5-18)

where $r_i(t)$ is the received signal at location i; s(t) is the transmitted signal; $h_i(t)$ is the multipath impulse response from sources to receiver at location i; n(t) is the additive white Gaussian noise (AWGN).

TABLE 5-1: SIMULATION PARAMETERS	
Room length (m)	6
Room width (m)	6
Room height (m)	3
Light source locations (m, m, m)	(1, 1.5, 3), (3, 1.5, 3), (5, 1.5, 3),
	(1, 4.5, 3), (3, 4.5, 3), (5, 4.5, 3),
Receiver locations (m, m, m)	A(0.1, 0.1, 0.2), B(3, 0.1, 0.2),
	C(3,3,0.2)
Bit rate (Mbps)	200
Length of test bits	4.8*10 ⁶
AWGN SNR (db)	1~15
OFDM carrier number	200
OFDM frequency bin number	1000

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According to the method demonstrated in Chapter 2, the multi-path impulse responses $h_i(t)$ at location A, B, C are given in Fig. 5-6:



Fig. 5-6: Impulse response at location A, B and C

We can see that location A, room corner, has the most significant multi-path spread; therefore, experiences the most severe ISI. Location C, room center, has the least significant multi-path spread so that it experiences the least severe ISI. Location B, wall side, has the medium multi-path spread, thus experiences the medium severe ISI.

The demodulated received signal $\hat{r}_i(t)$ is:

$$\hat{r}_i(t) = \begin{cases} r_i(t) & OOK\\ FFT(r_i(t)) & OFDM \end{cases}$$
(5-19)

The received user data can be obtained by

$$\hat{d}_{i}(t) = \begin{cases} 1 & \text{when } \hat{r}_{i}(t) \ge \text{decision threshold} \\ 0 & \text{when } \hat{r}_{i}(t) < \text{decision threshold} \end{cases}$$
(5-20)

5.4.2 Simulation Results of General OFDM Schemes in IOW

Fig. 5-7 shows the performance of OFDM and OOK modulation at location A, B and C respectively on bit rate of 200Mbps.



Fig. 5-7: Comparison of OFDM and OOK BER performance

In each location, OFDM gives much lower BER than OOK. In the severe multi-path scenario as location A, OOK modulation cannot provide with BER less than 10^{-2} as far as E_b/N_0 reaches 15dB. However, OFDM can still provide with BER of 10^{-5} when E_b/N_0 is 15dB.

Comparing BER performance in three locations, both OFDM and OOK experience BER increase when multipath spread increases. However, OFDM is much less affected. Fig. 5-8 shows the spectrum of OFDM and OOK at location B. We can make following two observations from the figure. First, OFDM makes more efficient use of the spectrum by allowing overlap. Second, OFDM divides the channel into small narrowband flat fading sub-channels. Therefore, it is resistant to frequency selective distortion, which comes from multipath effect. In addition, OFDM can effectively mitigate ISI through using CP. The only substantial drawback is that OFDM signal has a noise-like amplitude with large dynamic range, thus has a high PAPR.



Fig. 5-8: Received signal spectrum of OFDM and OOK at location B

5.5 Simulation Results of Clipping-noise Resistant OFDM schemes in IOW

Section 5.4 provides the investigation of general OFDM schemes. In IOW, waveform is carried by optical power, which is non-negative. As a result, we must convert regular bipolar OFDM signal waves to special unipolar signal waves. This is usually achieved by clipping out the negative components. It, nevertheless, may generate clipping noise which degrades performance. In this section, we explore several clipping noise resistant OFDM schemes [46].

5.5.1 Simulation Setup

We explore the performances of ACO-OFDM and PAM-DMT in both additive white Gaussian noise (AWGN) and multipath indoor IOW channels. In addition, pre-coding approaches are applied and evaluated. In multipath channels, we use equalizer to mitigate impulse distortion.

OFDM modulation flow chart is demonstrated by Fig. 5-9. For ACO-OFDM, only odd subcarriers are modulated by the complex input symbols. Since clipping noise falls only on even subcarriers, user symbols are not interfered. It is very easy to recover transmitted symbols at the receiver. For PAM-DMT, symbols from a real valued constellation like M-PAM are used to modulate the complex part of each subcarrier. Clipping noise only exists in the real part of each subcarrier, therefore, user symbols are free from its impact.



Fig. 5-9: Flow Chart of OFDM modulation

The multipath indoor IOW channels are generated by the method described in Chapter 2. Parameters for the channel models are given by Table 5-2.

TABLE 5-2: MULTIPATH INDOOR OPTICAL WIRELESS CHANNEL PARAMETERS	
Room dimensions	6m × 5m × 3m
Surface reflectance (ceiling, walls, floor)	0.8, 0.8, 0.3
Source location for:	
low multipath distortion channel (x, y, z)	(1.0m, 2.0m, 3.0m)
medium multipath distortion channel (x, y, z)	(0.1m, 0.2m, 3.0m)
high multipath distortion channel (x, y, z)	(0.1m, 0.1m, 3.0m)
Receiver aperture area	1cm ²
Receiver field-of-view	85°
Receiver location (x, y, z)	(2.5m, 2.5m, 1.0m)

In this table, $h_1(t)$, $h_2(t)$ and $h_3(t)$ represent channel with high, medium and low delay-spread

(multipath distortion), respectively. Their impulse responses are given in Fig. 5-10.



Fig. 5-10: Impulse responses of low, medium and high delay-spread channels

5.5.2 Simulation Results of Clipping-noise Resistant OFDM in AWGN Channel

Fig. 5-11 compares the BER performance versus signal-to-noise ratio (SNR) between ACO-OFDM without pre-coding (a), with DCT precoding (b), with DFT precoding (c) and with ZC precoding (d) in AWGN channel, applying signal constellation from 4-QAM to 1024-QAM. Shown in the figures, there is no significant difference in simulation results of the four precoding approaches. It indicates that though DCT, DFT and ZC precoding are able to reduce signal PAPR, they do not degrade bit-error rate (BER) performance.

Similar observations can be made from the BER comparison between PAM-DMT without precoding and with DCT precoding. The results are demonstrated in Fig. 5-12.



Fig. 5-11: BER versus SNR of ACO-OFDM without precoding (a), with DCT precoding (b),DFT precoding (c) and ZC predocing (d)



Fig. 5-12: BER versus SNR of PAM-DMT without precoding (a), with DCT precoding (b)

5.5.3 Simulation Results of Clipping-noise Resistant OFDM in Multipath Indoor Optical Wireless Channel

We extend our research from AWGN channel to multipath IOW channel. Fig. 5-13 – Fig. 5-16 demonstrate BER performance of ACO-OFDM, DCT precoded ACO-OFDM, DFT precoded ACO-OFDM and PAM-DMT respectively. Take Fig. 5-13 as an example. We can make following observations from the results. Comparing Fig. 5-13 (a) and (c) with Fig. 5-11 (a), we find that BER are substantially increased by multipath distortion. Further comparison between Fig. 5-13 (a) and (c) indicates that BER increase is more significant in the high multipath distorted channel than in the low one. That is because high multipath distortion produces more severe ISI. The results of Fig. 5-13 (b) and (d), however, illustrate that Zero-forcing Decision Feedback Equalizer (ZF-DFE) can recover BER performance by reducing ISI. Similar observations can be made in Fig. 5-14 – Fig. 5-16. Comparison between Fig. 5-13 – Fig. 5-16 shows that precoding approaches, though decrease PAPR, do not impact BER performance. In multipath optical wireless channel, PAPR performance of clipping and filtering are given in Fig.

5-17. The probability of PAPR being greater than the threshold $PAPR_0$ without clipping, with 3dB, 6dB and 9dB clipping are shown. We observe a point where the probability of $PAPR \ge$ $PAPR_0$ significantly drops. This point shifts 2.5dB for each 3dB more clipping, with probability of $PAPR \ge PAPR_0 = 10^{-3}$.



Fig. 5-13: BER versus SNR of ACO-OFDM without precoding in low multipath distortion channel without equalization (a), low multipath distortion channel with ZF FDE (b), high multipath distortion channel with ZF FDE (c), high multipath distortion channel with ZF FDE (d)



Fig. 5-14: BER versus SNR of ACO-OFDM with DCT precoding in low multipath distortion channel without equalization (a), low multipath distortion channel with ZF FDE (b), high multipath distortion channel with ZF FDE (c), high multipath distortion channel with ZF FDE (d)



Fig. 5-15: BER versus SNR of ACO-OFDM with DFT precoding in low multipath distortion channel without equalization (a), low multipath distortion channel with ZF FDE (b), high multipath distortion channel with ZF FDE (c), high multipath distortion channel with ZF FDE (d)



Fig. 5-16: BER versus SNR of PAM-DMT in low multipath distortion channel without equalization (a), low multipath distortion channel with ZF FDE (b), high multipath distortion channel with ZF FDE (c), high multipath distortion channel with ZF FDE (d)



Fig. 5-17: PAPR with Clipping and Filtering

5.6 Conclusions

This chapter investigates OFDM for indoor optical wireless communications. It is divided into two parts.

The first part of this chapter compares regular OFDM and OOK performance in indoor optical wireless communications. The simulation is carried at three different locations where different levels of multipath effect exist. The results indicate that OFDM has significantly better performance than OOK in all locations. In the worst case, when the receiver is placed at the corner of a room, OOK can hardly work, but OFDM maintains acceptable performance. OFDM is a much better choice than prevalent OOK for indoor optical wireless communications due to its high resistance to multipath spread. The reasons are: first, OFDM makes more efficient use of the spectrum by allowing overlap; second, OFDM divides the channel into small narrowband flat fading sub-channels; third, OFDM can effectively mitigate ISI through using CP.

The second part explores specific OFDM schemes for IOW applications, which are resistant to clipping noise. As OFDM signals are carried by non-negative optical carriers in IOW systems, we have to clip off the negative components of regular OFDM signals. We successfully demonstrated that ACO-OFDM and PAM-DMT are immune to clipping noise. In addition, combining with precoding methods, such as DFT-OFDM, ZCT-OFDM and DCT-OFDM, they can effectively reduce PAPR.

Chapter 6

MIMO Technology for Optical Wireless Communications using LED Arrays and Fly-eye Receivers

6.1 Introduction

In radio, multiple-input and multiple-output (MIMO) is defined as applying multiple antennas at both the transmitter and receiver to improve communication performance [47]. MIMO is one form of the smart antenna technologies. The input and output here indicate communication channels rather than transmitting and receiving devices. Consequently, in indoor optical wireless communications (IOWC), MIMO systems refer to the systems with multiple optical channels between the source and the receiver.

People have paid considerable attention to MIMO technology, since it significantly improves data throughput and link range without increasing bandwidth or transmission power. It spreads the power over multiple antennas to improve bandwidth efficiency or/and achieve diversity gain. Due to the advantages of this technique, it has been extensively applied in modern wireless standards, such as 802.11n(Wi-FI), LTE, WiMAX and HSPA+.

6.2 MIMO Configurations

6.2.1 MIMO System Model

The difference between a MIMO and non-MIMO IOWC system is the MIMO system consists of multiple transmitters and receivers. Assuming there are N_t light sources on the ceiling and N_r photo detectors at the receiving device, the received signal vector is [48]

$$y(t) = H(t) \otimes s(t) + n(t) \quad (6-1)$$

where y(t) is the received vector; H(t) is the channel characteristics matrix, which is described in details in Chapter 2; \otimes stands for convolution; s(t) is the transmitted signal vector as $s(t) = [s_1(t), s_2(t), ..., s_{N_t}(t)]^T$, where $[\cdot]^T$ is the transpose operator; n(t) represents the total noise, including ambient shot light noise and thermal noise. We assume the noise n(t) is independent of the transmitted signal and is a real Gaussian random process. It has a zero mean with a variance $\sigma(t)^2 = \sigma_{shot}(t)^2 + \sigma_{thermal}(t)^2$, where $\sigma_{shot}(t)^2$ and $\sigma_{thermal}(t)^2$ are the variances of shot noise and thermal noise, respectively. In practice, the multiple sources are generally implemented by LED arrays. As the LED sources are in close proximity and can be driven by same driver, we suppose they are perfectly synchronized.

6.2.2 Spatial Diversity

Spatial diversity, also called repetition coding (RC), is the most basic application of MIMO technology [49]. Diversity is one very effective remedy. It exploits the principle of providing the receiver with multiple distorted replicas of the same information bearing signal. Spatial diversity has been extensively studied in conventional wireless systems. Fig. 6-1 demonstrates the receiver

diversity system. Based on the combining methods, it can be divided into four categories: Selective Combining (SC), Maximal Ratio Combining (MRC), Equal Gain Combining (EGC) and Switch Combining (SSC) [49].



Fig.6-1: A typical receiver diversity system

A. Selective Combining

In SC, the branch with highest signal-to-noise ratio (SNR) is always selected. This can be represented as

$$\tilde{r} = \max(\tilde{r}_k)$$
 (6-2)

In a continuous transmission system, SC is not practical because of its requirement on constant monitoring all the branches. With the availability of this monitoring, the MRC should be a better choice for its out-performing over SC.
B. Maximal Ratio Combining

In a typical MRC diversity system, branches are weighted by their respective complex fading gains and added together. MRC realizes a maximum-likelihood (ML) receiver.

C. Equal Gain Combining

EGC and MRC are similar in that their diversity branches are co-phased. The difference between the two is that EGC does not weight its diversity branches. For a practical consideration, EGC is useful for modulation schemes with equal energy symbols, such as M-phase-shift keying (M-PSK).

D. Switched Combining

The switched combining receivers scan through all the diversity branches until one, which reaches a specific SNR threshold, is found. This branch is chosen and used until its SNR drops below the threshold. After this, the receiver will rescan all the branches and choose the next one, whose SNR exceeds the threshold. The most conspicuous advantage of SSC is that only one detector is needed in this approach.

6.2.3 Spatial Multiplexing

Spatial multiplexing (SMP) is another well-known MIMO technique and it has been described in details in [48]. In this configuration, independent data streams are transmitted from

all transmitters at the same time. SMP provides substantial enhanced bandwidth efficiency as $N_t \log_2(M)$ bit/s/Hz.

The signal vector *s* has N_t elements, which are independently modulated signals. In order to keep the total transmission power, the emitting intensity of each source is divided by N_t . Suppose ML detection is performed at the SMP MIMO receiver, the pairwise error probability (PEP) is defined as the probability that the receiver mistakes one transmitted signal for another, while the channel characteristics matrix H(t) is known. As a result, the PEP for SMP can be calculated as

$$PEP_{SMP} = PEP(s_{m1}(t) \to s_{m2}(t)|H(t)) = Q(\sqrt{\frac{r^2 T_s}{4N_0}} \|s_{m1}(t) - s_{m2}(t)\|_F^2) \quad (6-3)$$

6.2.4 Spatial Modulation

Spatial modulation (SM) is the combination of MIMO and digital modulation. It was first proposed at [50] and further explored at [51][52][53]. This modulation scheme extends the signal constellation to spatial dimension, which carries additional bits. Each light source has an unique binary sequence defined as spatial symbol. A source is activated only when the corresponding spatial symbol is being sent. Only one source is therefore transmitting in a symbol period. Fig. 6-2 demonstrates the structure of SM system.



Fig. 6-2: SM system structure

As SM modulates data in the signal domain and spatial domain at the same time, the spectrum efficiency becomes $log_2(N_t) + log_2(M)$ bit/s/Hz. In addition, since only one light source is active in one symbol period, ISI is effectively eliminated. The PEP performance of SM can be estimated by

$$PEP_{SMP} = PEP(s_{m1}(t) \to s_{m2}(t)|H(t)) = Q\left(\sqrt{\frac{r^2 T_s}{4N_0}} \|s_{m1}(t) - s_{m2}(t)\|_F^2\right)$$
$$= Q\left(\sqrt{\frac{r^2 T_s}{4N_0}} \sum_{n_r=1}^{N_r} |I_{m^{(2)}}^{SM}(t) \otimes h_{n_r n_t^{(2)}}(t) - I_{m^{(1)}}^{SM}(t) \otimes h_{n_r n_t^{(1)}}(t)|^2\right)$$
(6-4)

6.3 Angle-diversity Receivers

6.3.1 Angle-diversity Receiver Overview

In IOWC systems, substantial performance promotion can be obtained through applying an angle-diversity receiver. It consists of multiple branches covering different directions [54][55][56][57]. Each branch is followed by its own photo detector and amplifier, which convert the received optical power of this branch into amplified electrical current for further signal processing. Angle-diversity receivers provide many advantages: first, they reduce ambient light by pointing directly to the desired light source; also, they reduce ISIs and multi-path distortions, as most unwanted sources interferences and diffusions are rejected from the branch; finally, the receivers provide wide total field-of-view (FOV).

As in [57], a typical angle-diversity receiver is implemented using multiple nonimaging elements oriented to different directions. The authors systematically studied the theoretical gain of this kind of receiver and demonstrated an optical wireless communication system achieving 70Mbps. Fig. 6-3 depicts the setup of an angle diversity receiver. Multiple narrow FOV branches collaboratively cover a wide FOV. Ambient light and diffusion light that do not coincide with the signal light are blocked out.



Fig. 6-3: A typical configuration of angle-diversity receiver

6.3.2 Fly-eye Receiver Design

Yun and Kavehrad proposed several designs of fly-eye receiver. Alignment and focusing are major challenges to fly-eye receiver design [58]. A fly-eye receiver includes multiple independent branches pointing at different directions. Each branch, as a result, is expected to be aligned independently to achieve optimal MIMO gain. In addition, when users roam around in a typical indoor environment, the distance from a light source to the corresponding receiving branch varies from a few meters to tens of meters.

The general solution to adjust focusing is changing the relative position between the receiver lens and photo detectors. This method, however, requires much adaptive mechanisms and will make the receiver costly and bulky. The authors of [59] explored the attempt of fixing the detectors at one focal length behind the lens and compared its performance with adaptive configurations. We call the design with fixed detectors as a far-sighted eye and that with adaptive detectors as a perfect eye. In Fig. 6-4, the curves of received power with respect to source distance are demonstrated. They indicate that the far-sighted eye performs very closely to the perfect eye when working distance is long, and it provides a flat response when working distance is short. From the results, we conclude that the far-sighted eye is a prudent replacement for the perfect eye, given that the photo detectors are sufficiently sensitive. Significant simplicity is obtained from the absence of adaptive focusing mechanisms. It results from the constant response at dynamic range and the performance similarity of two kinds of eyes at long distance.



Fig. 6-4: Comparison of perfect eye and far-sighted eye on received power

Fly-eye receivers are expected to view various directions simultaneously. The most straightforward design of fly-eye receivers is employing distributed eyes with their own lens and photo detectors. The large number of lenses in this method, however, makes the system complex and bulky. A few novel ideas are proposed to solve the problem.

One instance is applying a transparent ball, glass or plastic, as the lens for all branches. Because of the symmetry of the ball, all branches share the same ball as demonstrated in Fig. 6-5. This design effectively reduces the complexity of the system. A problem of this design is the ball can be very heavy when large aperture is required. This system is most suitable for the networks covering small rooms.



Fig. 6-5: Fly-eye design with ball lens

For the scenarios where large apertures are required, an alternative approach is utilizing the off-axis imaging ability of a lens. The fly-eye design is depicted in Fig. 6-6. This system provides optimistic performance for large apertures, especially when a Fresnel lens is applied. The major drawback comparing with the ball lens is different light paths cannot be located far from the optical axis of the lens; otherwise the aberration will be significant. Although increasing photo detector area may improve system tolerance to aberration, the flexibility of this design is limited.



Fig. 6-6: Fly-eye design using off-axis imaging

6.4 Simulation Results and Discussions

6.4.1 Simulation Parameters

The communication environment is assumed as an ordinary office room. Room dimensions and the reflectance of each surface are reasonably chosen by referring engineering conventions. The source array consists of 7 sources. They form a hexagon on the ceiling with one in the center and six around. The fly-eye receiver consists of 7 branches. One branch is placed in the center facing vertically upwards and the other six are evenly distributed around the center one with 30 degree tilt. The fly-eye receiver is located at 841 locations, covering the whole room, for performance test. More details of the source, the fly-eye receiver, environment, and communication parameters are given in Table 6-1 – Table 6-4, respectively

Number of sources	7
Source location (m)	Source 1: (1.2379 3.0000 3.0000)
	Source 2: (2.1190 1.4740 3.0000)
	Source 3: (3.8810 1.4740 3.0000)
	Source 4: (4.7621 3.0000 3.0000)
	Source 5: (3.8810 4.5260 3.0000)
	Source 6: (2.1190 4.5261 3.0000)
	Source 7: (3.0000 3.0000 3.0000)
Source direction	(0 0 -1) for each source
Transmitting power (W)	1 for each source
Source half-power angle (degree)	60 for each source

TABLE 6-1: SIMULATION PARAMETERS OF THE MIMO IOWC SYSTEM – SOURCE ARRAY

TABLE 6-2: SIMULATION PARAMETERS OF THE MIMO IOWC SYSTEM - RECEIVER	
THEE C 2. SINCE THE INTERIOR OF THE INTERIOR CONCENTRAL	

Number of receiver branches	7
Branch direction	Branch 1: (0.0000 0.0000 1.0000)
	Branch 2: (0.6428 0.0000 0.7660)
	Branch 3: (0.3214 0.5567 0.7660)
	Branch 4: (-0.3214 0.5567 0.7660)
	Branch 5: (-0.6428 0.0000 0.7660)
	Branch 6: (-0.3214 -0.5567 0.7660)
	Branch 7: (0.3214 -0.5567 0.7660)
Branch responsivity (A / W)	1 for each branch
Branch field-of-view (degree)	20 for each branch
Branch aperture area (cm ²)	1 for each branch

 TABLE 6-3: SIMULATION PARAMETERS OF THE MIMO IOWC SYSTEM - ENVIRONMENT

Room dimensions (m)	Length: 6
	Width: 6
	Height: 3
Room surface reflectance	Ceiling: 0.9
	Wall _{East} : 0.7
	Wall _{West} : 0.7
	Wall _{Nouth} : 0.7
	Wall _{South} : 0.7
	Floor: 0.1
Spatial resolution (m)	0.02
Number of reflections traced	3

TABLE 6-4: SIMULATION PARAMETERS OF THE MIMO IOWC SYSTEM - COMMUNICATION SYSTEM		
Test data length	1e6	
Test data rate range (Mbps)	3000 ~ 20	
Test data rate change	Data rate reduced by half for each simulation	
Test noise level range (dBmW)	-120 ~ -70	
Test noise increment (dBmW)	1	

A single branch non-MIMO receiver model is created for reference. In order to make a fair comparison between MIMO and non-MIMO systems, its aperture area equals the total area of all MIMO receiver branches, and its FOV equals the total FOV of the MIMO receiver. All simulations on the MIMO receivers will be repeated on the non-MIMO receiver.

6.4.2 BER Spatial Distributions for MIMO IOWC Systems

The BER distributions of the non-MIMO system, the MIMO EGC system, and the MIMO MRC system of 600Mbps data rate and -105dBmW noise are given in Fig 6-7 to Fig. 6-12. The noise level is a reasonable estimation for a regular office environment [84]. The results show that the MIMO systems provide very low outage probabilities. They indicate that the MIMO methods guarantee most areas in the room satisfactory communications.



600Mbps IOW BER performance, without MIMO (-105dBmW noise), BER outage:0.30559

Fig. 6-7: Non-MIMO system BER performance of 600Mbps data rate, -105dBmW noise



600Mbps IOW BER performance, without MIMO (-105dBmW noise), BER outage:0.30559

Fig. 6-8: Non-MIMO system BER performance of 600Mbps data rate, -105dBmW noise (top view)



600Mbps IOW BER performance,7X7 MIMO using EGC (-105dBmW noise), BER outage:0.04042

Fig. 6-9: MIMO EGC system BER performance of 600Mbps data rate, -105dBmW noise



Fig. 6-10: MIMO EGC system BER performance of 600Mbps data rate, -105dBmW noise (top view)



600Mbps IOW BER performance,7X7 MIMO using MRC (-105dBmW noise), BER outage:0.00832

Fig. 6-11: MIMO MRC system BER performance of 600Mbps data rate, -105dBmW noise



600Mbps IOW BER performance,7X7 MIMO using MRC (-105dBmW noise), BER outage:0.00832

Fig. 6-12: MIMO EGC system BER performance of 600Mbps data rate, -105dBmW noise (top view)

In order to make a clearer observation of the BER spatial distributions, we increase the noise level to -90dBmW, which represents a tough IOWC environment involving direct sunlight exposure. The results are shown in Fig. 6-13 to Fig. 6-18. In this section, we focus on the BER distributions; the impact of noise level to BER performance will be discussed in details in next section.



600Mbps IOW BER performance, without MIMO (-90dBmW noise), BER outage:0.56718

Fig. 6-13: Non-MIMO system BER performance of 600Mbps data rate, -90dBmW noise



600Mbps IOW BER performance, without MIMO (-90dBmW noise), BER outage:0.56718

Fig. 6-14: Non-MIMO system BER performance of 600Mbps data rate, -90dBmW noise (top view)



600Mbps IOW BER performance,7X7 MIMO using EGC (-90dBmW noise), BER outage:0.46611

Fig. 6-15: MIMO EGC system BER performance of 600Mbps data rate, -90dBmW noise



600Mbps IOW BER performance,7X7 MIMO using EGC (-90dBmW noise), BER outage:0.4661

Fig. 6-16: MIMO EGC system BER performance of 600Mbps data rate, -90dBmW noise (top view)



600Mbps IOW BER performance,7X7 MIMO using MRC (-90dBmW noise), BER outage:0.1724[,]

Fig. 6-17: MIMO MRC system BER performance of 600Mbps data rate, -90dBmW noise



600Mbps IOW BER performance,7X7 MIMO using MRC (-90dBmW noise), BER outage:0.1724

Fig. 6-18: MIMO MRC system BER performance of 600Mbps data rate, -90dBmW noise (top view)

The following observations can be made from these figures. Generally, the BER distributions are closely related to the sources array's lay-out. We name the source in the array center as the center source and the other six around as the peripheral sources. The areas in peripheral sources' footprints have low BER. That is because these areas have line-of-sight (LOS) transmission paths with small emitting angles and receiving angles. Those factors provide high SNR. Regarding the areas in the footprint of the center source, the situation is more complex. On one hand, in the non-MIMO system, though these areas receive strong signal power from LOS paths from the center source, they also receive LOS signal power from the peripheral sources. Since these LOS signals arrive at different times, they cause inter-source interference. The inter-source interference produces ISI in the communication system, which significantly increases BER. As a result, the areas in the center source footprint in the non-MIMO system have high BER. On the other hand, in MIMO EGC and MIMO MRC systems, the areas in the center source footprint also have LOS links from the center source and the peripheral sources. The MIMO angle diversity mechanism, however, distinguishes the signal power from different sources. By applying different combining algorithms, the ISI is significantly reduced. The areas in the center source footprint, therefore, have low BER in the MIMO systems. For the areas outside the sources footprints, they have high BER in both MIMO and non-MIMO systems due to low SNR.

6.4.3 Impact of Ambient Noise

In order to explore the impact of ambient noise to system performance, we observe the BER distributions from -75dBmW to -114dBmW with 600Mbps data rate. The results for the non-





Fig. 19 to Fig. 21 give the BER distributions of the non-MIMO system with noise decreasing from -70dBmW to -120dBmW. The low BER areas first appear in the footprints of the peripheral sources. With the noise decrease, these BER areas extend; therefore, the system outage probability reduces. It converges to 32% when the noise level reaches -108dBmW. An interesting observation from the figures is that the areas in the center source footprint, which are

located in the center of the room, never have low BER. That is because the inter-source interference from neighbor sources increases channel delay-spread, thus increases BER.



Fig. 6-20: BER distributions and outage probabilities of the non-MIMO system (noise from -87dBmW to -102dBmW)



Fig. 6-21: BER distributions and outage probabilities of the non-MIMO system (noise from -105dBmW to -120dBmW)

Fig. 6-22 to Fig. 6-24 give the BER distribution of the MIMO EGC system with noise from -70dBmW to -120dBmW. Being different from the non-MIMO system, the MIMO EGC system has low BER areas first appearing in the center source footprint. As the center low BER areas extend with noise decrease, other low BER areas emerge under the peripheral sources' footprints. They extend with the center low BER areas as noise decreases. The outage probability converges to 0 at -114dBmW noise level. An interesting observation is that there are several high BER "islands" surrounded by low BER areas. The reason is that in these locations, multiple sources send light into the most efficient receiver branch through LOS paths and produce intersource interference.



Fig. 6-22: BER distributions and outage probabilities of the MIMO EGC system (noise from -75dBmW to -84dBmW)



Fig. 6-23: BER distributions and outage probabilities of the MIMO EGC system (noise from -87dBmW to -102dBmW)



The BER performance of the MIMO MRC system is given in Fig. 6-25 to Fig. 6-27. The BER distribution changes in the same pattern as the MIMO EGC system; however, the outage

probability converges much faster. It decreases to 0 at the noise level of -102dBmW. That is because MIMO EGC receivers weight different branches according to their SNRs. High SNR paths are emphasized in the combining. The MIMO MRC, therefore, has a better BER performance than MIMO EGC at the same noise level.



Fig. 6-25: BER distributions and outage probabilities of the MIMO MRC system (noise from -75dBmW to -84dBmW)



Fig. 6-26: BER distributions and outage probabilities of the MIMO MRC system (noise from -87dBmW to -102dBmW)



Fig. 6-27: BER distributions and outage probabilities of the MIMO MRC system (noise from -105dBmW to -120dBmW)



Fig. 6-28: Outage probability versus noise level

Fig. 6-28 demonstrates the outage probability change with noise. An important observation from the results is that the gain of applying MIMO significantly varies with noise. With noise above -84dBmW, the outage probability of the non-MIMO system is the lowest. That is because in high noise environments, SNR is the dominant factor determining BER. The non-MIMO system has the highest SNR, for it has the largest aperture area and FOV. Though the non-MIMO system suffers more severe ISI than the MIMO systems, its overall performance exceeds the MIMO systems in high noise environments. With noise decrease, the ISI replaces SNR as the dominate factor of system performance. The MIMO EGC system and MIMO MRC system outperform the non-MIMO system at the noise level of -84dBmW and -89dBmW, respectively. After -89dBmW, the MIMO systems provide lower outage probability than the non-MIMO system, and the MRC method is always better than the EGC method. The outage probabilities of both MIMO systems converge to zero. The outage probability of the non-MIMO system, however, doesn't converge to zero. That is because severe ISI remains in the footprint of the center source.

6.5 Conclusions

This chapter investigates the application of MIMO technology for IOWC. MIMO approaches can improve IOWC performance through two mechanisms: one is sending same data on all MIMO channels to improve communication reliability and we call this method diversity; the other is sending different data on different MIMO channels to increase channel capacity and this method is called multiplexing.

In this chapter, we simulate the BER performance of the MIMO EGC and the MIMO MRC spatial diversity system, and compare them with the non-MIMO system. The performance is evaluated by BER outage probability. The results indicate that in regular indoor environments, MIMO systems outperform non-MIMO systems. MIMO MRC systems are more efficient than MIMO EGC systems. The reason is that multiple small FOV and aperture branches of MIMO receivers separate the light from different sources, thus mitigate ISI. Nevertheless, when the noise level is high, non-MIMO systems may outperform MIMO systems because non-MIMO systems have larger aperture size, which become the dominant factor to system performance. But in this environment, both MIMO and non-MIMO systems have limited mobilities and can only work in a small area.

Chapter 7

Wireless Solutions for Aircrafts based on Optical Wireless Communications and Power Line Communications

7.1 Introduction

To implement wireless communications in aircrafts, the most important concern is the interference of wireless devices to aircraft navigation and communication systems [60]. Personal electronic devices (PED) generate two kinds of microwave radiations: intentional and spurious. Intentional radiations usually come from the PEDs of wireless communication function. They are generated when the devices transmit data via radio frequency (RF) wireless links. Spurious radiations are unintentional, but increase the RF noise level. Though they widely exist in all PEDs, spurious radiations are much more significant in wireless PEDs. Intentional radiations are not commonly considered as an interference, because of the strict band limitations as given in Table 7-1 [61]. The accumulated spurious radiations, however, can be very high in a frequency for aircraft navigation and communication systems; thus cause considerable impacts to operations. Studies pointed out that laptop computers are most frequently suspected as sources of interference. In 40 PED related reports collected by the International Air Transport Association (IATA), laptop computers contribute 40% of them. The IATA reports also indicate that navigation system is the most frequently affected system. 68% of the cases in the reports happen in navigation system [62]. Another study on Aviation Safety Reporting System (ASRS) database confirmed that navigation system is most vulnerable to PED interference; cellphones and laptops are the most common causes [63].

Omega navigation	ADF	HF	Marker beacon	VOR, localizer	VHF COM	Glide slope
10 <mark>1</mark> 4 kHz	1901750 kHz	230 MHz	74.85, 75, 75.15 MHz	108118 MHz	118136 MHz	328335 MHz
GSM 400	GSM 850	GSM 900	DME	TCAS/ATC	GPS	SATCOM
450.4467.6 MHz 478.8496 MHz	824894 MHz	876960 MHz	9601220 MHz	1030,1090 MHZ	1575 MHz	1529, 1661 MHz
GSM 1800 17101880 MHz	European UMTS	GSM 1900	IMS band: WLAN 802.11b,g Bluetooth, Home RF	Low-range altimeter	Microwave landing system	WLAN 802.11a
	18802025 MHz 21102200 MHz	18501900 MHz	2446.52483.5 MHz	4.3 GHz	5.03, 5.09 GHz	51505350 MHz
Weather radar 5.4 GHz	WLAN 802.11a 57255825 MHz	Weather radar 9.3 GHz	Sky radio 11,700 MHz	DBS TV 12.212.7 GHz	Frequency separa International Tele Union, Geneva, S	ation regulated by ecommunication witzerland

TABLE 7-1: FREQUENCY SEPARATION (THE AIRCRAFT NAVIGATION AND COMMUNICATION ARE INDICATED IN BLUE)

Recent researchers have drawn considerable interest in using optical wireless communications (OWC) in airplanes and space vehicles [64]. Light-emitting diode (LED) is widely applied for illumination in new generation commercial aircrafts. OWC demonstrates significant advantages in size, weight, power, cost and electromagnetic interference (EMI) reduction. In addition, OWC combining with powerline communications (PLC) creates an efficient delivery mechanism for fulfilling broadband access onboard an aircraft, while providing efficient and economic lighting [65]. This chapter explores the potential capabilities of these two emerging techniques.

7.2 Powerline Communications Channel Model

In an aircraft powerline grid, signal propagation does not take place along a direct path from a transmitter to a receiver. Echoes exist because of the reflections at grid junctions and they cause multipath distortion. At the receiver, each transmission path is weighted by a coefficient, g, which is defined as the product of the reflection coefficient and the transmission coefficient of the nodes along transmission path. Since both reflection and transmission coefficients are no greater than one, the weighting coefficient g is always equal to or smaller than unity. By applying these weighting coefficients, the grid channel model can be formulated as the summation of multiple paths with different lengths and weighting factors. As a result, the channel model of a powerline network can be expressed as:

$$H(f) = \sum_{i=1}^{N} g_i e^{-\alpha(f)d_i} e^{-j\beta(f)d_i} \quad (7-1)$$

where *N* indicates the number of significant arrived paths at the receiver; d_i is the length of the *i*-th path; g_i is the weighting factor of the *i*-th path.

Based on this model, the theoretical Shannon capacity of an aircraft PLC network can be calculated. The simulation result is shown in Fig. 7-1 [65]. It indicates 326Mbps throughput can be achieved at transmission power of 20dBm. ($N_0 = -117$ dBM/Hz is assumed)



Fig. 7-1: Shannon capacity of aircraft's powerline

7.3 Optical Wireless Communications

7.3.1 Simulation Configurations

Using our OWC simulation methods described in previous chapters, we can analyze OWC performances in most environments by carefully customizing environment parameters. With the support from Boeing Company and Airbus SAS through National Science Foundation (NSF) Industry/University Cooperative Research Center on Optical Wireless Applications (COWA), we explored the visible light propagation features and OWC performance in airplane cabin. Based on our findings, we validated the possibility of applying LED Visible Light Communications (VLC) in airplane cabins for high speed wireless transmissions. Our work focused on the received power and delay-spread distribution in the cabin when using existing overhead LED reading lights as VLC transmitters. We simulated the environment of one passenger seating row in a typical Boeing 737-900 airplane. The interior dimensions we referred are shown in Fig. 7-2.

SURFACE	SAMPLE	REFLECTANCE
Seatback	sample 4 – rough plastic	0.5
Upholsteries	sample 44 – linen	0.3
Floor	sample 18 – rug	0.1
Ceiling	sample 4 – rough plastic	0.5
Cabin interior wall	sample 4 – rough plastic	0.5

TABLE 7-2: MATERIAL SAMPLES AND REFLECTANCE FOR SURFACES IN THE CABIN



Fig. 7-2: Cabin cross-sections

Different surfaces in the cabin are made from different materials, thus have different reflectances. We chose material samples for cabin surfaces from Fig. 7-3 [66] and looked up their corresponding reflectances. The results are given in Table 7-2. Sources and receivers profiles are given in Table 7-3 and Table 7-4, respectively.

TABLE 7-3: SOURCES PROFILES		
MODEL	2LA455953	
Light Color	Warm White	
Beam Angle	15^{0}	
Operating Current	700mA max	
Power Consumption	3.2W max	
	(10, 7, 45)	
Source Locations	(29, 7, 45)	
	(48, 7, 45)	

TABLE 7-4: RECEIVERS PROFILES

FIELD-OF-VIEW	60^{0}
Responsivity	1A/W
Aperture Area	1 cm^2
Photo current from background noise	1e-5 uA

By tracing the bounces of lights, we simulated the impulse response from the light sources to all locations in the seating area. Received power and delay-spread for each location are further computed based on individual impulse response. They are demonstrated as their spatial distributions in Fig. 7-4 and Fig. 7-5.

7.3.2 Illuminance Distribution Results

In photometry, illuminance, formly brightness, is the total luminous flux incident on a surface, per unit area. The unit of illuminance is lux (lx), or lumen per square meter. Photometric and radiometric parameters are listed in Table 7-5 and Table 7-6, respectively.
	Symbol	Quantity	SI Unit	Abbr.			
	F	luminous flux	lumen	lm			
I_{v}		luminous intensity	candela (=lm/sr)	cd			
	E_{ν}	illuminance	$lux (=lm/m^2)$	lx			
_	M	luminous efficiency	lumen per watt	lm/W			
Symbol	ા હા	JANTITY	SI Unit	Abbr.			
P	rad	iant power	watt				

watt per solid angle

watt per square meter

watt per steradian per square meter

The	pho	otom	etric	para	met	ter	lun	ninous	flux	(unit:	lm)	and	the	radiometric	parameter	r radiant
(•,	II	1	1		,	11	(7 , 0)	1	(7,2)		<i>.</i> .	1			

power (unit: W) can be calculated by (7-2) and (7-3), respectively.

radiant intensity

irradiance

radiance

Ι

Ε

L

$$F = 683 \int_{380nm}^{780nm} S(\lambda) V(\lambda) d\lambda \quad (7-2)$$
$$P = 683 \int_{\lambda_I}^{\lambda_H} S(\lambda) d\lambda \quad (7-3)$$

where $S(\lambda)$ is the spectral power and $V(\lambda)$ is the luminous efficiency function, or "photopic" function [67]. The "photopic" function correlates with human brightness perception and is given in Fig. 7-3.



Fig. 7-3: Luminous efficiency function

W/sr

W/m²

 $W/(sr \cdot m^2)$

From illumination considerations, specific illuminance level should be maintained for different applications [68]. From communication considerations, sufficient illuminace should be guaranteed for high signal-to-noise ratio (SNR) [69].

Fig. 7-4 is the illuminace distribution of OWC in airplane cabins. The environment model is established by using interior dimensions in Fig. 7-2 and material samples in Table 7-2. The color of each element in the figure indicates illuminance level of the corresponding area. To unify our results, in all following figures of this chapter, we use blue for the areas where relatively better communication is performed and red for the areas where relatively worse communication is performed.



Fig. 7-4: Power Distribution of LED Reading Lights in a Typical Cabin Environment on Boeing 737-900

We can make following observations from the figure. First, center area has highest illuminance and it decreases in peripheral area. Second, optical wireless is able to provide sufficient illuminance for illumination. Illuminance requirement for using computer monitor is 30lx, for reading is 300lx [68]. 100.00% of seating area satisfies monitor operation requirement, with the minimum illuminance of 44.08lx; 100.00% of reading area satisfies reading requirement, with the minimum of 470.27lx. Third, optical wireless is able to provide sufficient SNR for communications. 10dB SNR is required to provide 10⁻⁵ bit-error-rate (BER) for basic on-off keying (OOK) modulation in additive white Gaussian noise (AWGN) channel [69]. 100.00% of seating area satisfies the illuminance requirement, with the minimum SNR of 15.11dB.

7.3.3 Delay-spread Distribution Results

Delay-spread is the difference between the time of arrival of the earliest significant multipath component and that of the latest. It is a measure of the multipath richness of a communication channel and indicates the channel bandwidth.

Delay-spread is most commonly quantified through rooted-mean-square (RMS) delayspread, as [25]

$$s = \sqrt{\int_{-\infty}^{\infty} (t - \mu)^2 h^2(t) dt} / \int_{-\infty}^{\infty} h^2(t) dt$$
 (7-4)

where h(t) is the impulse response and μ is average delay defined by

$$\mu = \sqrt{\int_{-\infty}^{\infty} th^2(t)dt / \int_{-\infty}^{\infty} h^2(t)dt} \quad (7-5)$$

Previous research shows the maximum transmission rate over a wireless channel is one to several times of the inverse of its delay-spread, given that no diversity or equalization applied [26][27].

Fig. 7-5 is the delay-spread distribution in aircraft VLC system. Like illuminance distribution, center area has the lowest delay-spread and it increases in peripheral area. Defining bandwidth as the inverse of delay-spread, 100% of the seating area is able to provide hundreds MHz bandwidth.



Fig. 7-5: Delay-spread Distribution of LED VLC System on Boeing 737-900

Commercial LEDs provide bandwidth from a few MHz to tens of MHz; advanced LEDs is able to provide a few hundreds MHz to GHz [70][71]. Commercial photodiodes have reached hundreds of MHz [72]. The overall bandwidth of a communication system is jointly determined by source, channel and receiver. Our research shows that channel is not the bandwidth bottleneck in cabin environment.

7.3.4 Bit-error Rate Distribution and Outage Probability

Bit-error Rate (BER) is the number of bit errors divided by the total number of received bits of a data stream over a communication channel. It is the most direct measure of communication performance. "A bit error rate of better than 10e-5 is considered acceptable in wireless local area network (LAN) applications [73]." As BER varies at different locations in the cabin, we apply BER outage probability to evaluate network coverage with satisfactory BER.

BER outage [74] is defined as "the fraction of the useful area over that the BER criterion is not satisfied." It is a statistic index of indoor wireless network performance. It indicates the portion of the room area, where the BER is above criterion. In other words, lower outage, larger reliable communication area. The flowchart for BER outage calculation is given by Fig. 7-6.



Fig. 7-6: BER outage probability calculation flowchart

BER distribution and outage probability for bitrates from 3Gbps to 375Mbps are given in Fig. 7-7 – Fig. 7-10.



Fig. 7-7: BER distribution and probability for 3Gbps transmission



BER distribution of in-cabin LED VLC system,1500Mbps, BER outage =0.10922

Fig. 7-8: BER distribution and probability for 1.5Gbps transmission



Fig. 7-9: BER distribution and probability for 750Mbps transmission



Fig. 7-10: BER distribution and probability for 375Mbps transmission

Top views provide us better understanding of BER distribution in the cabin. Fig. 7-11 is a group of BER top views from 3Gbps to 23Mbps.



Fig. 7-11: Top views of BER distributions, 3Gbps - 23Mbps

Fig. 7-12 gives the statistic BER outage probabilities for different bitrates. The bitrate requirements for popular multimedia services are provided for references in Table 7-7.



Fig. 7-12: BER outage probabilities with bitrates

SERVICE	BITRATE REQUIREMENT
Blu-ray Disc	40 Mbps
HDV 720p MPEG2	19 Mbps
DVD MPEG2	9.8 Mbps
Mp3 music services	Up to 320 kbps
Phone services	8 kbps
Teleconference services	384 kbps

From these results, we observe that BER outage decreases with bitrate decrease. It reduces to less than 4% at 200Mbps. The little high BER area only remains in peripheral area, due to low SNR. We therefore confidently believe that optical wireless is able to support high speed multimedia service in aircraft cabin environment.

7.4 Wireless Applications for Commercial Airplanes

Our core research partner, Boeing Company discussed several emerging wireless applications in commercial airplanes in [76]. The PLC-OWC solution is a strong candidate to implement these ideas for its merits in size, weight, power consumption, EMI safety and labor cost.

A. Reading light passenger service units (PSUs)

The existing reading lights above passenger sitting area provide two services. One is general illumination and the other is data, voice and video communications.

B. Passenger infotainment

Passenger infotainment system is interconnected by visible light between seats to form a high speed mesh network. It projects to offer in-flight entertainment (IFE). Our research shows PLC-OWC can provide sufficient bandwidth for this application in a commercial airplane cabin which may have 400 or more passengers.

C. Cabin interphones

The lighting LED can transmit low bandwidth voice signal to flight attendants' head phones.

D. Interconnection of line-replaceable-units (LRUs) over environmental barrier

It is not possible to interconnect the LRUs through enclosed areas with fiber. OWC provides optical transmission through these barriers without routing long distance.

7.5 Conclusions

This chapter investigates the PLC-OWC solution for wireless communications in airplane cabin. At first, we discussed the PLC channel model and the estimated capacity. Next, we investigate the OWC performance in airplane cabin, when LED is used as transmitter. The cabin environment model is established by the interior dimensions of a typical Boeing 737-900 commercial plane. Using this model, we simulate the visible light propagation characteristics, and calculate power and delay-spread distributions. In addition, based on the impulse response matrix obtained, we simulate BER distribution and explore BER outage probabilities versus bitrates.

PLC-OWC provides an economical, lightweight, high-speed, energy saving and EMI free solution. The results consolidate the possibility of applying this technology for high-speed in-cabin wireless services.

Chapter 8

Indoor Positioning Methods based on Light-Emitting Diodes Visible Light Communications

8.1 Introduction

This chapter proposes a novel indoor positioning algorithm using Visible Light Communications (VLC). The algorithm is implemented by pre-installed Light-Emitting Diode (LED) illumination systems. It recovers the VLC channel features from illuminating visible light and estimates receiver locations by analytically solving Lambertian transmission equation group. According to our research, the algorithm is able to provide positioning resolution higher than 0.5 millimeters, in practical indoor environment. The performance significantly exceeds conventional indoor positioning approaches using microwaves.

8.2 LED and Indoor Navigation Methods

Due to concerns over energy efficiency, it is being realized for some time now that LED will be the future desired lighting source. Compared to traditional light sources, LED offers many advantages such as long life expectancy, brightness efficiency and environmental friendliness. Moreover, as a semiconductor light emitting device, LED is easy to modulate at relatively high rates for other than lighting applications [2][77]. Global Positioning System (GPS) is the most popular positioning system in the world; however, its performance is substantially limited in indoor environments, because of the significant power attenuation when Electromagnetic (EM) waves pass through ceilings and walls. Consequently, most existing indoor positioning approaches are based on short-range indoor wireless communication techniques.

There are two categories of indoor positioning methods [78][79][80][81]. The first kind is based on microwaves, including Wireless Local Area Network (WLAN), Blue-Tooth (BT), Radio Frequency Identification (RFID) and Ultra-Wide Band (UWB). Because of the impacts of EM interference, noise, stability and other factors, these methods could only provide positioning accuracy of tens of centimeters [82]. The other kind is based on optical tracking and imaging. This method can provide good positioning accuracy of a few millimeters; however, it costs more in infrastructure and involves complex image processing.

This chapter proposes a method of indoor positioning leveraging pre-installed LED illumination systems. The method recovers channel characteristics from the incident light and estimates receiver location by analytically solving Lambertian equation group. On one hand, as visible light is free from EM interferences, this method provides much higher resolution than microwave systems, to 0.5 millimeters. On the other hand, it works on existing illumination infrastructures and requires no image processing technique; thus the cost is lower compared with the imaging method. To sum up, this method will be a strong candidate for indoor localization for its performance and cost advantages.

The rest of the paper is organized in the following way: we introduce the algorithm and the corresponding system in section 8.3; demonstrate and discuss our research results in section 8.4; draw conclusions in section 8.5.

8.3 Design, Operation and Algorithm

8.3.1 System Design

The illumination lights are always installed in a specific layout, for instance the rectangular grid shape as shown in Fig. 8-1 (a). We pick up an elementary unit, the cuboid as shown in Fig. 8-1 (b) spanned by four LEDs to apply our algorithm. When the positioning algorithm is researched within one unit, the results can be straightforwardly extended to all other units, thus covers the entire area.



Fig. 8-1: LED illumination layouts (a) and the elementary unit (b)

Suppose the user, or Mobile Terminal (MT) is placed at an arbitrary location in the unit, facing vertically upward. The light sources LEDi (i=1, 2, 3, 4) transmit signals to the user via visible light. As the radiation follows Lambertian pattern, the impulse response from LEDi to the MT is [3][9]

$$h_{i}(t) = \begin{cases} \frac{n+1}{2\pi} \frac{\cos^{n}(\phi_{i})\cos(\theta_{i})A_{R}}{R_{i}^{2}} \delta(t - \frac{R_{i}}{c}) & \theta_{i} \le FoV \\ 0 & \theta_{i} > FoV \end{cases}$$
(8-1)

when the four lights are modulated by $s_i(t)$ (i=1,2,3,4), respectively, the signal received by the MT is

$$r_{MT}(t) = P_0 \sum_{i=1}^{4} s_i(t) \otimes h_i(t) + n(t) \quad (8-2)$$

where n is the mode number of the radiation lobe that specifies the source directionality; ϕ_i is the angle between source orientation vector and the vector pointing from source to receiver; θ_i is the angle between receiver orientation vector and the vector pointing from receiver to source; A_R is receiver area; FoV is the Field-of-View of the receiver; R_i is the distance between the source and receiver; c is the speed of light; P_0 is the power emitted by each LED.

In this chapter, we consider that the LEDs are of first order Lambertian pattern, which is the most common case of general illumination LEDs; the receiving angle is always smaller than FoV; ignore the difference of impulse delay, as it is of nanosecond scale while the system is designed to work at tens of kbps. The received signal becomes

$$r_{MT}(t) = P_0 \sum_{i=1}^{4} s_i(t) h_i = \frac{P_0 A_R}{\pi} \sum_{i=1}^{4} s_i(t) \frac{\cos(\phi_i) \cos(\theta_i)}{R_i^2} \quad (8-3)$$

8.3.2 Channel Features Recovery

The lights are individually modulated by their location information, as positioning references to the MT. The signals, however, sent from different sources are mixed in the air interface. We need to retrieve individual signal and the corresponding channel features. The Time Division Multiplexing (TDM) method is developed to achieve this goal. All sources transmit synchronized frames periodically. In one frame period, the i-th LED is assigned a

specific time slot between T_{i-1} and T_i , in which it sends its encoded location information. We encode X, Y, Z coordinates of the LED into three 16 bits RZ codes, respectively as s(t) and OOK modulation is applied. In this slot, since high and low light intensity is emitted with equal probability, the average power is $\frac{P_0}{2}$. In other slots, the source emits constant high light intensity for only illumination purpose. The average powers of these slots are P_0 . The frame structure is given by Fig. 8-2.



Fig. 8-2: Frame Structure of the Positioning System (One Period)

Based on the mixed signal $r_{MT}(t)$, $s_i(t)$ can be straightforwardly obtained by sampling in the specific slot. To retrieve h_i is more challenging, due to power mix up. For this reason, we intentionally design the frame structure to keep transmission power stable, regardless of the content of the location codes. By this method, we develop an algorithm to calculate the channel features as:

$$h_{i} = \frac{2}{P_{0}} \left(\frac{4}{3.5} \frac{\int_{0}^{T_{4}} r_{MT} dt}{T_{4} - 0} - \frac{\int_{T_{i-1}}^{T_{i}} r_{MT} dt}{T_{i} - T_{i-1}}\right) \quad (8-4)$$

8.3.3 Positioning Algorithm

Based on the h_i and $s_i(t)$ obtained from section 8.3.2, we can derive the MT location. We decode the X, Y, Z coordinates of LEDi from $s_i(t)$ and define them as (L_{ix}, L_{iy}, L_{iz}) . The coordinates of the MT are unknown and are notated by (M_x, M_y, M_z) . Placing these into (8-1), we obtain:

$$h_{i} = \frac{1}{\pi} \frac{\cos(\phi_{i})\cos(\theta_{i})A_{R}}{R_{i}^{2}}$$

$$= \frac{A_{R}}{\pi} \frac{(L_{iz} - M_{z})^{2}}{((L_{ix} - M_{x})^{2} + (L_{iy} - M_{y})^{2} + (L_{iz} - M_{z})^{2})^{2}} \quad (8-5)$$

$$= K \frac{(L_{iz} - M_{z})^{2}}{((L_{ix} - M_{x})^{2} + (L_{iy} - M_{y})^{2} + (L_{iz} - M_{z})^{2})^{2}}$$

For computation convenience, we rewrite (5) as

$$\sqrt{\frac{h_i}{K}} = \frac{|L_{iz} - M_z|}{(L_{ix} - M_x)^2 + (L_{iy} - M_y)^2 + (L_{iz} - M_z)^2} = \frac{L_{iz} - M_z}{(L_{ix} - M_x)^2 + (L_{iy} - M_y)^2 + (L_{iz} - M_z)^2}$$
(8-6)

where K is a constant defined as $K = \frac{A_R}{\pi}$. Since L_{iz} is always greater than M_z , we can remove

the absolute symbol of $\left|L_{iz} - M_z\right|$.

For values i = 1, 2, 3, 4, we obtain an equation group about user locations. Solving the equation group by matrix operation, we have:

$$\begin{pmatrix}
(L_{1x} - L_{2x}), (L_{1y} - L_{2y}), (L_{1z} - L_{2z} - \frac{1}{2}\sqrt{\frac{K}{h_{1}}} + \frac{1}{2}\sqrt{\frac{K}{h_{2}}}) \\
(L_{2x} - L_{3x}), (L_{2y} - L_{3y}), (L_{2z} - L_{3z} - \frac{1}{2}\sqrt{\frac{K}{h_{2}}} + \frac{1}{2}\sqrt{\frac{K}{h_{3}}}) \\
(L_{3x} - L_{4x}), (L_{3y} - L_{4y}), (L_{3z} - L_{4z} - \frac{1}{2}\sqrt{\frac{K}{h_{3}}} + \frac{1}{2}\sqrt{\frac{K}{h_{4}}})
\end{pmatrix} \begin{pmatrix}
M_{x} \\
M_{y} \\
M_{z}
\end{pmatrix} = \\
\begin{pmatrix}
L_{1x}^{2} + L_{1y}^{2} + L_{1z}^{2} - L_{2z}^{2} - L_{2y}^{2} - L_{2z}^{2} + L_{2z}\sqrt{\frac{K}{h_{2}}} - L_{1z}\sqrt{\frac{K}{h_{1}}} \\
L_{2z}^{2} + L_{2y}^{2} + L_{2z}^{2} - L_{2z}^{2} - L_{3y}^{2} - L_{3z}^{2} + L_{3z}\sqrt{\frac{K}{h_{3}}} - L_{1z}\sqrt{\frac{K}{h_{2}}} \\
L_{3x}^{2} + L_{3y}^{2} + L_{3z}^{2} - L_{3z}^{2} - L_{4y}^{2} - L_{4z}^{2} + L_{4z}\sqrt{\frac{K}{h_{3}}} - L_{1z}\sqrt{\frac{K}{h_{3}}}
\end{pmatrix}$$
(8-7)

The solution (M_x, M_y, M_z) is the estimated 3D user location coordinates. Error is defined by the Eucleadian distance between the estimated position and the real, as: $\sqrt{(M_{x-est} - M_{x-real})^2 + (M_{y-est} - M_{y-real})^2 + (M_{z-est} - M_{z-real})^2}$. Since three coordinates are considered, this error is called 3D positioning error. In most cases, users are more concerned about their 2D locations (M_x, M_y) and the 2D estimation error $\sqrt{(M_{x-est} - M_{x-real})^2 + (M_{y-est} - M_{y-real})^2}$. We, therefore, refer estimation to 2D estimation by default. 3D estimation will be analyzed later in the chapter.

8.4 Results and Discussion

8.4.1 Test Parameters

The environment is assumed as a 3m tall large indoor plaza. LED lights are installed in a rectangular grid shape layout, with neighboring distance of identically 3m. The elementary positioning unit becomes a $3m \times 3m \times 3m$ cuboid, with four LEDs on the ceiling corners.

All LEDs have identical half-power angl of 60° and emitting light intensity of 680 lm. The floor is divided into a 30×30 grid, while each element of the grid indicates a test spot. We assume the MT is at 1m height on each test spot. Camera area is 1 cm^2 and its sensitivity is 1 A/W. The receiver is facing vertically upward and FoV is 70° .

The frame structure is described in section 8.3.2. To sum up, there are $16 \times 3 \times 4 = 192$ bits in a frame. We want to average the received waveforms to mitigate the noise impact. Suppose frames repeat 50 times in a second for one test, the bit rate of the system shall be $192 \times 50 = 9600$ (bps), which is applicable to most commercial LEDs [83].

We evaluate the system performance under different noise levels. For large-area optical receivers, thermal noise is often negligible comparing to shot noise. The shot noise is associated with the background light induced current I_{bg} . It can be modeled as Gaussian with Power Spectrum Density (PSD) of $N_0 = 2qI_{bg}$, where q is an electron charge. With Wolf's research [84], in a 10kbps system as we use, the shot noise ranges from -137.96dbmW when the receiver is exposed to direct sunlight, to -180.97dbmW when only artificial light is applied. For convenience, we assume that the noise power in practical environment is between -140dbmW to -180dbmW.

8.4.2 Positioning Error and Distribution

Fig. 8-3, Fig. 8-4 and Fig. 8-5 show the 2D positioning error distribution of this algorithm under the noise level of -140dbmW (strong noise, direct sunlight exposure), -160dbmW (moderate noise) and -180dbmW (weak noise, artificial light exposure only), respectively. As we can see, in all cases, the large estimation errors exist at the near-corner areas of the unit and the

error decreases as it approaches to the room center. That is because of the directionality of Lambertian transmission. When the user is near the corner, the received intensity from the diagonal source can be significantly degraded and it is vulnerable to noise interferences. This will cause substantial positioning error when it is placed into matrix computation. We also notice that the positioning error at the same location decreases 10 times as the noise decreases by 20dbmW.



Fig. 8-3: Spatial Distribution of Positioning Error at Noise Power of -140dbmW (Strong Noise, Direct Sunlight)

Fig. 8-6, Fig. 8-7, and Fig. 8-8 exhibit the positioning error distribution by histograms, at the three noise levels. When noise is strong, the positioning errors are widely spread from 0mm to 30mm. This distribution tends to concentrate to zero, as noise reduces. The histograms are good explanations to the error spatial distributions. Based on the results, we believe that this algorithm is able to provide 0.5 millimeters positioning accuracy in weak noise environment; even in strong noise, the accuracy can be guaranteed to tens of millimeters.



Fig. 8-4: Spatial Distribution of Positioning Error at Noise Power of -160dbmW (Moderate Noise)



Spatial Distribution of Positioning Error at Noise Power of -180dbmW

Fig. 8-5: Spatial Distribution of Positioning Error at Noise Power of -180dbmW (Weak Noise, Artificial Light Only)



Fig. 8-6: Histogram of Positioning Error at Noise Power of -140dbmW



Positioning Error Statistics at Noise Power of -160dbmW

Fig. 8-7: Histogram of Positioning Error at Noise Power of -160dbmW



Fig. 8-8: Histogram of Positioning Error at Noise Power of -180dbmW



Fig. 8-9: Mean of Positioning Error

Mean of positioning error is another measure of the approach. Fig. 8-9 depicts the average 2D positioning error in the unit at different noise levels. From the figure, we observed a dramatic drop of average positioning error. In a practical environment (noise power starting from - 140dbmW to -180dbmW), the average error is below 7.3mm and it decreases to below 0.5 mm at -163dbmW.



Fig. 8-10: Outage probability of positioning error

8.4.3 Positioning Error Outage

Instead of the positioning error for specific locations, we are also interested in what fraction of the room area is able to provide positioning error below certain criteria. We use outage probability to indicate the partition of the whole area, where positioning error is greater than the threshold. Fig. 8-10 is the simulation results of positioning error outage, with respect to decreasing noise power. Thresholds are from 0.5mm to 500mm. As we can see, the outage decreases with noise power and converges to zero. In the worst practical environment, where the receiver is directly exposed to sunlight, the probability of positioning error greater than 50mm is nearly 0 percent. That means all of the room area provides positioning resolution of 50mm. If we track other curves in the figure, we will find that the outage probability of 25mm, 10mm, 5mm and 0.5mm reaches zero at -148dbmW, -155dbmW, -163dbmW and -180dbmW, respectively.



Fig. 8-11 Impact of receiver height on positioning error

8.4.4 Impact of Receiver Height

Since in our simulations, users are placed at the same height. We are also interested if the receiver height impacts the positioning error. The simulation is repeated at different receiver heights (0.6m, 0.8m) and the error outage result is shown in Fig. 8-11. For the same outage

threshold, we observe excellent match of the curves representing different receiver heights. The result indicates that positioning error is unsubstantially impacted by receiver height.



Fig. 8-12 Comparison of mean of 2D and 3D positioning error

8.4.5 Comparison of 2D and 3D Positioning

In some cases, users also need the height information for certain purposes, for instance, automatically positioning items in storage stacks. 3D positioning errors are similar to 2D in spatial error distribution pattern and the noise vulnerability features. We focus only on the difference between the two. Fig. 8-12 shows the comparison of mean of 2D and 3D errors. The result shows that 3D errors are always roughly 1.8 times of 2D errors, as they both converge to zero with noise decreasing. It implies that positioning error in vertical direction is more

significant than in horizontal. Fig. 8-13 is the comparison of 2D and 3D positioning outages. We find offset between each pair of curves. For each pair, the maximum shift can reach 7db. It means that, 3D positioning requires 7db more Signal-to-Noise Ratio SNR to reach the same resolution as 2D.



Fig. 8-13: Comparison of outage probability of 2D and 3D positioning error

8.5 Conclusions

This chapter proposes a novel visible light indoor positioning technique, by solving the Lambertian transmission equation group. According to our results, this method reduces positioning error to less than 0.5 millimeters, as the state-of-the-art microwave indoor positioning can only reduce the error to tens of centimeters. Our research also shows that the performance of this approach is not impacted by receiver height. Moreover, this method provides

not only conventional 2D position information, but also 3D information. There is 80% increase in the mean of positioning error and 7dB less noise tolerance, when 3D is applied instead of 2D positioning.

Chapter 9 Conclusions and Future Work

9.1 Conclusions

In this dissertation, several fundamental topics in IOWC have been investigated. This technology is a high speed, energy efficient and secure solution to RF band congestion.

The research starts with developing an efficient method to establish indoor optical wireless models by tracking light pulses experiencing reflections and diffusions. When a general channel model is obtained, data simulation is applied to statistically calculate BER distribution and outage. By observing the results, the author finds that multi-path effect significantly exists at room corner locations, and inter-source interference significantly exists at the overlapping area of light footprints. Moreover, the influence of inter-source interference is more considerable than multipath effect to communication performance. Based on the observations, the author concludes that the major impact factors to VLC are multi-path effect and inter-source interference in general indoor environment. They degrade communications performance by causing impulse response distortion.

Based on the investigation of IOWC channel model and the impact factors, the author explores OFDM and MIMO techniques to increase IOWC capacity. For the OFDM approaches, ACO-OFDM and PAM-DMT are successfully demonstrated to be immune to clipping noise. In addition, precoding methods, such as DFT-OFDM, ZCT-OFDM, and DCT-OFDM, can effectively reduce PAPR. For the MIMO approaches, BER performance of the MIMO EGC and the MIMO MRC spatial diversity systems are simulated and compared with the non-MIMO system. The results indicate that in regular indoor environments, MIMO systems outperform non-MIMO systems because the multiple-branch receivers separate the light from different sources, thus mitigate ISI. Nevertheless, when the noise level is high, non-MIMO systems may outperform MIMO systems because non-MIMO systems have larger aperture size, which become the dominant factor to system performance in high noise environment.

The dissertation also covers specific industrial applications of IOWC. For the investigation of IOWC in airplane cabin, the light propagation features, power, delay-spread, BER, and outage probability analyses are performed according to the interior dimensions of a typical Boeing 737-900 commercial plane. The results confirm that PLC-IOWC provides an economical, lightweight, high-speed, energy saving and EMI free solution to airplane onboard wireless access. For using IOWC for indoor navigations, a novel positioning algorithm is proposed by solving the Lambertian transmission equation group. According to the results, this method theoretically reduces positioning error to less than 0.5 millimeters, as the state-of-the-art microwave indoor positioning can only reduce the error to tens of centimeters.

9.2 Future Work

IOWC is a promising technology, which is developing with an incredible speed. Nowadays, complex algorithms and modulations can be incorporated into efficient and small embedded systems.

One interesting topic for future research is spatial multiplexing for IOWC. Spatial multiplexing provides wireless access to multiple users simultaneously. It will significantly increase the bandwidth efficiency. With the widely installation of LED arrays and reduced cost of manufacturing multiple branch receivers, increasingly more attention will be paid to this technique.

There is also a clear roadmap to implement spatial coding. In spatial coding, the uncorrelated MIMO channels carry different signals to represent the transmitted symbol. Combining with error-correct coding theories, this technique will increase bandwidth efficiency, and provide additional protection to blocking.

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