A Study on the Impact of Nonlinear Characteristics of LEDs on Optical OFDM

by

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Abstract:

“The truly learned man is he who understands that what he knows is but little in comparison with what he does not know.” (Ali Bin Aby-Talib (r.a))

In the 21th century, the demand for wireless bandwidth will continue to increase to serve bandwidth-intensive applications such as high definition (HD) video streaming via Internet. The existing radio frequency (RF) technology standards, i.e. the IEEE (institute of electrical and electronics engineers) 802.11a/b/g/n standards, suffer from the bandwidth bottleneck. However, optical wireless (OW) technology has the potential for being an attractive complementary to realize broadband wireless access indoors.

Recently, indoor OW communication systems based on light emitting diode (LEDs) as light sources and orthogonal frequency division multiplexing (OFDM) as a modulation technique have gained significant attention. For instance, in visible light communication (VLC), in addition to lightening capabilities of white LEDs, they can be modulated to achieve high-speed wireless data transmission. Optical OFDM (O-OFDM) systems are able to support high data rates through parallel transmission of high order multi-level quadrature amplitude modulation ($M$-QAM) symbols on orthogonal subcarriers. Complex channel equalizers are not required and the time-varying channel can be easily estimated using frequency-domain channel estimation. Adaptive modulation per subcarrier can be easily realized based on the uplink/downlink requested data rates and quality of service (QoS). Also, the inherent robustness of OFDM against multipath effects, and the possibility to combine it with any multiple access scheme, e.g. time division multiple access (TDMA), frequency division multiple access (FDMA), and code division multiple access (CDMA), makes it an excellent choice for indoor OW links. However, only little research has been done in this area, and many open questions exist such as the performance of O-OFDM systems in the presence of LED nonlinearities.
The time domain OFDM signal exhibits a random-like variations which is characterized by a high peak-to-average power ratio (PAPR). Therefore, special attention must be paid to its sensitivity to nonlinearities. This thesis analyzes the performance of indoor OW systems based on OFDM in the presence LED nonlinearities, i.e. amplitude and clipping distortions. The key parameters investigated are the LED DC bias point, the average OFDM signal power modulating the LED, and the constellation orders. The distortion analysis shows that LED nonlinearities can drastically degrade system performance. As a consequence, a linearization technique is proposed. A digital pre-distorter is applied to compensate for LED nonlinearities. It is found that the degradation can greatly be mitigated by using the proposed predistortion technique.

In a practical system, the OFDM signal is forcibly clipped at the LED turn-on voltage (TOV) and purposely clipped prior to the LED modulation for voltage levels above the saturation voltage which corresponds to the maximum permissible AC/pulsed current recommended by the LED manufacturer. The OFDM pre-conditioning prior to the LED modulation is to ensure that the LED chip does not overheat, in order to avoid degradation in output light or, in the worst case, total failure. Therefore, there is a need to describe the nonlinear behavior of an LED with a model that fits the LED measured transfer function and to level off at both extremes to represent the TOV at one end and the saturation voltage at the other end. A model offering the desired ”S-shaped” curve is proposed.

Analytical methodologies to evaluate error probability of uncompensated O-OFDM system and compensated O-OFDM system are proposed. A comparison with Monte Carlo simulation results are carried out. The simulation curves and the analytical curves show a close match in terms of symbol error ratio (SER).

Finally, hardware prototypes to demonstrate online short-range optical transmission are developed. Visible light transmission is considered. Apart from providing proof of concept, experimental measurements for practical scenarios are obtained to analyze the performance in real environments. Realizing the analog optical transmitter and receiver front ends, involves schematic design, layout design, printed circuit board (PCB) fabrication, component assembly and soldering. The O-OFDM model is based on Matlab/Simulink/C programming. Digital signal processing (DSP) development boards are configured to run the O-OFDM model.

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Author: Hany Elgala
Supervisor: Prof. Dr. Harald Haas
I certify that I have prepared this PhD Thesis by my own without any inadmissible outside help.

Jacobs University, 29.03.2010

(Hany Elgala)
I would like to take this opportunity to extend my deepest gratitude to my supervisor, Professor Harald Haas, of the Faculty of Engineering and Science, Jacobs University Bremen. He has a sharp eye for details and superb skills in developing novel ideas, these have been instrumental in the success of my PhD. His huge enthusiasm on research and fathomless knowledge in many areas has deeply inspired me. I fully appreciate his continuous support and encouragements, especially during my difficult times in work and life. His innumerable and useful advises on my career development is treasured.

A special word of thank goes to Dr. Raed Mesleh for the fruitful discussions and the valuable comments he provided me through my PhD study.

Finally, this thesis is dedicated to my family. My unutterable appreciation is to their unfailing love, continuous support and countless encouragements over all these years. Especially, I would like to thank my parents, Abdelmoniem and Raya, for their devoted nourishments and considerations. Their assistance has been invaluable to me during all these years. In addition, a special word of love goes to my wife, Riham, for her support and for being patient on me during the past three years.

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Bremen
April 20, 2010
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List of Abbreviations

HD: High definition
RF: Radio frequency
OW: Optical wireless
LED: Light emitting diode
OFDM: Orthogonal frequency division multiplexing
O-OFDM: Optical orthogonal frequency division multiplexing
VLC: Visible light communication
M-QAM: Multi-level quadrature amplitude modulation
QoS: Quality of service
TDMA: Time division multiple access
FDMA: Frequency division multiple access
CDMA: Code division multiple access
PAPR: Peak-to-average power ratio
SER: Symbol error ratio
PCB: Printed circuit board
DSP: Digital signal processing
OC: Optical communication
FSO: Free-space optics
LASER: Light amplification by stimulated emission of radiation
OF: Optical fiber
LD: Laser diode
PIN: p-intrinsic-n
PD: Photodiode
APD: Avalanche photodiode
IR: Infrared
SSL: Solid-state lighting
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<th>Acronym</th>
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<tr>
<td>VLCC</td>
<td>Visible light communications consortium</td>
</tr>
<tr>
<td>IGvlc</td>
<td>Visual Light Communication Interest Group</td>
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<tr>
<td>HANs</td>
<td>Home access networks</td>
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<tr>
<td>IM</td>
<td>Intensity modulation</td>
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<tr>
<td>DD</td>
<td>Direct detection</td>
</tr>
<tr>
<td>LOS</td>
<td>Line of sight</td>
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<tr>
<td>FOV</td>
<td>Field-of-view</td>
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<tr>
<td>O-WLAN</td>
<td>Optical wireless local area network</td>
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<tr>
<td>ISI</td>
<td>Inter-symbol interference</td>
</tr>
<tr>
<td>MSD</td>
<td>Multi-spot diffusing</td>
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<tr>
<td>ADR</td>
<td>Angle diversity receiver</td>
</tr>
<tr>
<td>4G</td>
<td>Next generation wireless communication systems</td>
</tr>
<tr>
<td>EMI</td>
<td>Electro-magnetic interference</td>
</tr>
<tr>
<td>ICI</td>
<td>Inter-carrier interference</td>
</tr>
<tr>
<td>OAP</td>
<td>Optical access point</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
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<tr>
<td>BS</td>
<td>Base station</td>
</tr>
<tr>
<td>PoE</td>
<td>Power over Ethernet</td>
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<tr>
<td>WHN</td>
<td>Wireless home networking</td>
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<tr>
<td>EVM</td>
<td>Error vector magnitude</td>
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<td>AWGN</td>
<td>Additive white Gaussian noise</td>
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<tr>
<td>FEC</td>
<td>Forward error correction</td>
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<tr>
<td>BER</td>
<td>Bit error ratio</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital to analog converter</td>
</tr>
<tr>
<td>ADC</td>
<td>Analog to digital converter</td>
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<tr>
<td>CP</td>
<td>Cyclic prefix</td>
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<tr>
<td>TIA</td>
<td>Transimpedance amplifier</td>
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<tr>
<td>LPF</td>
<td>Low pass filter</td>
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<tr>
<td>SC</td>
<td>Single carrier</td>
</tr>
<tr>
<td>OOK</td>
<td>On-off keying</td>
</tr>
<tr>
<td>PPM</td>
<td>Pulse-position modulation</td>
</tr>
<tr>
<td>RZ</td>
<td>Return-to-zero</td>
</tr>
<tr>
<td>NRZ</td>
<td>Non-return-to-zero</td>
</tr>
<tr>
<td>DFE</td>
<td>Decision-feedback equalizer</td>
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<tr>
<td>SSM</td>
<td>Single-subcarrier modulation</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
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<tr>
<td>MSM</td>
<td>Multiple-subcarrier modulation</td>
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<tr>
<td>BPSK</td>
<td>Binary phase shift keying</td>
</tr>
<tr>
<td>TOV</td>
<td>Turn-on voltage</td>
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<tr>
<td>IFFT</td>
<td>Inverse fast Fourier transform</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier transform</td>
</tr>
<tr>
<td>RMS</td>
<td>Root-mean-square</td>
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<tr>
<td>SSPA</td>
<td>Solid state power amplifier</td>
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<tr>
<td>CIE</td>
<td>International commission on illumination</td>
</tr>
<tr>
<td>ZF</td>
<td>Zero-forcing</td>
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<tr>
<td>CCI</td>
<td>Co-channel interference</td>
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<tr>
<td>PA</td>
<td>Power amplifier</td>
</tr>
<tr>
<td>pdf</td>
<td>Probability density function</td>
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<tr>
<td>RV</td>
<td>Random variable</td>
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List of Symbols

$D$ Duty cycle  
$S$ Slots per symbol time  
$M$ Constellation size of a QAM modulation  
$N_g$ Number of guard subcarriers  
$N$ Number of time-domain samples or data symbols or subcarriers per OFDM symbol  
$j$ The imaginary unit  
$x_k$ OFDM time-domain samples  
$X_n$ Data symbols  
$(\cdot)^*$ Complex conjugate of a symbol  
$R^{\{DCO\}}$ Data rate of DCO-OFDM system  
$R^{\{ACO\}}$ Data rate of ACO-OFDM system  
$T_s$ Sampling clock period  
$f_s$ Frequency separation between subcarriers  
$v_{\text{out}}(v_{\text{in}})$ PA output voltage  
$v_{\text{in}}$ PA input voltage  
$v_{\text{max}}$ PA maximum output voltage  
$k$ knee factor of the PA or the S-shaped LED model  
$|\cdot|$ Absolute value of element  
$v_{\text{LED}}$ Voltage across the LED  
$i_{\text{LED}}$ Current through the LED  
$i_{\text{max}}$ Maximum permissible AC/pulsed current through the LED  
$f(v_{\text{LED}})$ Function describing the relation between the voltage across the LED and the current through the LED  
$\mu$ OFDM signal mean
$\sigma_o^2$ \hspace{1cm} OFDM signal variance  
$X$ \hspace{1cm} Continuous random variable  
$x$ \hspace{1cm} Value taken by the random variable $X$  
$P_X(x)$ \hspace{1cm} Probability density function (pdf)  
$\rho$ \hspace{1cm} Effective signal-to-noise ratio  
$\sigma_{AWGN}^2$ \hspace{1cm} Gaussian noise  
$p_n$ \hspace{1cm} Nonlinearity induced noise power  
$p_{ad}$ \hspace{1cm} Noise component due to the amplitude distortion of the OFDM signal  
$p_{uc}$ \hspace{1cm} Noise component due to the clipping of the upper peaks of the OFDM signal  
$p_{lc}$ \hspace{1cm} Noise component due to the clipping of the lower peaks of the OFDM signal  
$m$ \hspace{1cm} bias point  
$i_{LED}(v_{LED})$ \hspace{1cm} S-shaped LED model equation  
g($v_{LED}$) \hspace{1cm} Unity slope linear curve which intersects with S-shaped LED curve at the bias point  
$[v_l, v_u]$ \hspace{1cm} Interval of the integral to estimate the amplitude distortion contribution to the nonlinear induced noise power  
P_{V_{LED}}(v_{LED}) \hspace{1cm} pdf of the OFDM signal  
v_{up}$ \hspace{1cm} Voltage corresponding to the bias point voltage subtracted from the voltage at the intersection of the linearized LED curve with the maximum permissible AC/pulsed current level  
v_{lp}$ \hspace{1cm} Voltage corresponding to the intersection of the linearized LED curve with the $x$-axis (i.e. no current conduction through the LED) subtracted from the bias point voltage  
$Q(x)$ \hspace{1cm} Q-function  
P_t$ \hspace{1cm} Transmitted optical power  
P_r$ \hspace{1cm} Received optical power  
$H(0)$ \hspace{1cm} Optical path-loss  
$A_{PD}$ \hspace{1cm} Photodiode active area  
d \hspace{1cm} Distance between the transmitter and the receiver  
$\phi$ \hspace{1cm} Angle with respect to the transmitter  
$\psi$ \hspace{1cm} Angle with respect to the receiver  
$T_s(\psi)$ \hspace{1cm} Filter gain  
g($\psi$) \hspace{1cm} Concentrator gain  
$R_0(\phi)$ \hspace{1cm} Transmitter radiant intensity  
$\alpha$ \hspace{1cm} Transmitter beam angle
$F$  luminous flux in lumens

$I_v$  Luminous intensity measured in candela

$P(\lambda)$ Spectral power distribution

$\Delta\lambda$  Step of changing $\lambda$ which is equal to 10nm

$V(\lambda)$  CIE 1931 (international commission on illumination) eye sensitivity function in the photonic vision regime

$\gamma$  Peak luminous efficacy based upon the sensitivity of the eye at 555nm

$\xi$  Conversion factor relating the photometric power to the radiometric power

$\Omega$  LED solid angle in steradian

$d$  Separation distance between the transmitter and the receiver

$b$  Vertical distance between the transmitter and the receiver

$a$  Horizontal distance between the transmitter and the initial position directly located under the transmitter
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1 Introduction

1.1 Brief History of Optical Communication

The use of light to send messages is a very old idea. Fire and smoke signaling were used in ancient civilizations. For example, the ancient Greeks used polished shields to reflect sunlight to signal in the battle and Roman records indicate that polished metal plates were used as mirrors to reflect sunlight for long distance signaling. Chinese started using fire beacons followed by the Romans and American Indians using smoke signals [1].

Optical communication (OC) systems based on semaphore lines technique were the earliest form of technological OC and date back to the 1790s. In 1792, the French inventor Claude Chappe built the first visual telegraphy system between cities in France [1]. The line between two cities was composed of a series of towers. As shown in Figure 1.1 each tower is equipped with a semaphore. The design had two arms connected by a cross-arm. Information is encoded by the position of the mechanical arms. Each arm had seven positions, and the cross-arm had four more permitting a 196-combination code. These positions were assigned to numeric symbols in connection with a code book.

In the early 1800s, the US military uses a wireless solar telegraph called ”Heliograph” that signals using Morse code flashes of sunlight reflected by a mirror. The flashes are produced by momentarily pivoting the mirror, or by interrupting the beam with a shutter. The navy often uses blinking lights, i.e. Aldis lamps, to send messages also using Morse code from one ship to another. In 1880, the first example of free space optics (FSO) technology was demonstrated by Alexander Graham Bell with his ”Photophone” that used sunlight reflected off a vibrating mirror and a selenium photo cell to send voice on a light beam [2].
Figure 1.1: Optical telegraph tower on the Litermont (Liter Mountain) near Nalbacht, Saarland, Germany, employing the semaphore system invented by French engineer Claude Chappe in 1791, towers are 5 km to 10 km apart.

Until the late 1960s, radio and radar communications were more successful than OC and took the spotlight [3]. OC started to get real attention with the invention of the light amplification by stimulated emission of radiation (LASER) and the laser diode (LD) in the 1960s, followed in the 1970s by the development of low-loss optical fibres (OFs) as a medium for transmitting information using light, the invention of the OF amplifier in the 1980s, and the invention of the in-fibre Bragg grating in the 1990s. These inventions formed the basis for the telecommunications revolution of the late 20th century and provided the infrastructure for the Internet. For example, the Nobel prize in physics 2009 has honored three scientists, who have played important roles in shaping the modern information technology, with one half to Charles K. Kao. He initiated the search and the development of the low-loss OF. These inventions also influenced the mainstream of the research to be focused on OF transmission more than on free space transmission. However, FSO is expansively used today in a variety of applications [4]. Short-range communication via IrDA (infrared data association) interfaces is an example of free space optical communication [5–7]. The advancement of inexpensive opto-electronic devices, such as LEDs and LDs, p-intrinsic-n (PIN) photodiodes (PDs) and avalanche photo-diodes (APDs) and various optical components, has resulted in the improvement of free space optical systems. There has been a great deal of research activity that has examined the extent to which FSO can support future requirements for high bit rate services and methods for mitigating the inherent
problems associated with the transmission medium.

Since the late 1970s and to date, significant research has been done on the applications of optical wireless (OW) data communications indoors. The research has been focused on the infrared (IR) region and the first indoor OW system was developed over 25 years ago. In 1979, an indoor OW communication system has been presented by F. R. Gfeller and U. Bapst [8]. In their system, diffuse optical radiation in the near-IR region was utilized as signal carrier to interconnect a cluster of terminals located in a room to a common cluster controller. Some of the significant published system demonstrations in the 1970s and in the 1980s are summarized in Table 1.1. Link configuration and directionality are covered in section 1.2.1.

Table 1.1: Chronology of indoor optical wireless communication research.

<table>
<thead>
<tr>
<th>Date</th>
<th>Organization</th>
<th>Directionality</th>
<th>Bit rate</th>
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<tr>
<td>1978/81 [8,9]</td>
<td>IBM</td>
<td>Diffuse</td>
<td>64-125 kbps</td>
</tr>
<tr>
<td>1985 [12]</td>
<td>Fujitsu</td>
<td>Hybrid</td>
<td>48 kbps</td>
</tr>
<tr>
<td>1985 [13]</td>
<td>HP Labs</td>
<td>Directed LOS</td>
<td>1 Mbps</td>
</tr>
<tr>
<td>1987 [15]</td>
<td>Bell Labs</td>
<td>Directed LOS</td>
<td>45 Mbps</td>
</tr>
</tbody>
</table>

Over the last five years, the focus of research has changed drastically with the emerging of visible light communication (VLC) [16]. Solid-state lighting (SSL) is a rapidly developing area, both in terms of commercial exploitation, and academic and industrial research. LEDs with a wide range of colors are available, including white emission, and the output power available and device efficiencies are increasing rapidly. The field of applications is also expanding. White LEDs are commonly used as replacements for incandescent lamps and as their performance constantly improves. These devices will become candidates for replacement of general illumination. These solid-state sources can be modulated at rates many times that of incandescent or fluorescent alternatives, thus offering the possibility of broadcasting information at the same time as providing illumination. VLC originated in Japan and the visible light communications consortium (VLCC) established in November 2003 [17], with major companies in Japan is aiming to publicize and standardize the VLC technology. There is now growing interest in Europe, and
work to develop standards within the IEEE is also underway (IEEE 802.15 WPAN Visual Light Communication Interest Group (IGvlc)) [18]. The OMEGA project aims to realize the EUs vision of the future Internet and home access networks (HANs) [19]. OW communications techniques will be combined in order to provide a range of communications channels. Boston University’s College of Engineering is launching a major program, under a National Science Foundation grant, to develop the next generation of wireless communications technology based on visible light [20].

1.2 Review on Indoor Optical Wireless Systems

1.2.1 Optical Carrier Modulation and Link Configurations

For most OW indoor applications, LEDs are the favored light sources due to relaxed safety regulations, low cost, and high reliability compared to LDs. LEDs emit incoherent light. Therefore, it is very difficult to collect appreciable signal power in a single electromagnetic mode. This incoherent reception does not provide a stable carrier, which makes it impossible to construct an efficient coherent receiver. Simple and low cost optical carrier modulation and demodulation are usually achieved through intensity modulation (IM) with direct detection (DD). The desired waveform is modulated onto the instantaneous power of the optical carrier and the detector generates a current proportional to the received instantaneous power, i.e. only the intensity of the optical wave is detected and there is no frequency or phase information.

Conventionally, six different indoor link configurations are defined by Kahn and Barry in [21] and are classified according to the existence of a line of sight (LOS) between the transmitter and the receiver as well as the degree of directionality, i.e. source beam-angle and detector field-of-view (FOV). The most common configurations are the directed-LOS, non-directed LOS, and non-directed non-LOS configurations. A directed LOS link is established using a narrow-beam source and a narrow FOV detector as shown in Figure 1.2(a). This configuration encounters minimal signal degradation and maximal transmission data rates. However, user mobility is limited, only point-to-point communication is supported, and a LOS path must be always maintained. Such link can be created, for instance, between
two portable devices or by aiming the user device at a hotspot [22]. Alternatively, a non-directed LOS link with a wide-beam source, as depicted in Figure 1.2(b), offers a point-to-multipoint communication and supports user mobility at most positions inside the room. Although a dominant LOS component exists, multipath components must be carefully controlled.

![Diagram of different indoor optical link configurations. (a) Directed-LOS link, (b) Non-directed-LOS link, (c) Diffuse link, and (d) Quasi-diffuse link.](image)

Figure 1.2: Different indoor optical link configurations. (a) Directed-LOS link, (b) Non-directed-LOS link, (c) Diffuse link, and (d) Quasi-diffuse link.

The utmost robustness against shadowing and the transmitter-receiver alignment is achieved by the non-directed non-LOS link design, which is often referred to as a diffuse link (see Figure 1.2(c)). In a diffuse link, the communication between a wide-beam source and a wide FOV receiver relies on signal reflections from the surfaces in the room such as the ceiling or walls, i.e. the LOS path is not required. Such link is easy to use from a users prospective, more robust to shadowing, and offers better mobility which make it well suited for optical wireless local area network (O-WLAN) application. Nevertheless, the diffuse link significantly suffers from high path-losses and multipaths create inter-symbol interference (ISI). Hence, the achieved data throughput is sternly limited.

A quasi-diffuse link employing multispot diffusion and angular diversity aims
to combine the robustness of the diffuse link with high data rates. This technique is based on multi-spot diffusing (MSD), first proposed by Yun and Kavehrad in 1993 [23]. As shown in Figure 1.2(d), the link utilizes multiple narrow-beam transmitter and an angle diversity receiver (ADR) [24–26]. The function of the transmitter is to generate several diffusing spots to cover a room ceiling with a uniform distributed optical signal. The receiver consists of several narrow FOV detectors aimed at different directions. Such link offers lower path-loss and less multipaths, at a lower transmission power compared to a conventional wide-beam diffuse link, while providing a high level of user mobility compared to LOS links. However, performance enhancement is achieved at the expense of increasing complexity.

For IR applications, peak wavelengths between 780 nm and 950 nm are available and coincide with the peak sensitivity of silicon PDs [27]. The majority of IR systems use the near infrared IR band mainly due to the availability of effective, low-cost sources and detectors. With directed LOS links, high data rates in the order of Gbps can be achieved [28]. The first IR system was based on a diffuse link operating at 950 nm and 125 kbps proposed by Gfeller and Bapst in 1979. A faster system proposed by March and Khan in 1996 achieves a data rate of 50 Mbps [29]. In quasi-diffuse systems, a data rate of 70 Mbps is demonstrated by Carruther and Kahn in 2000 [30].

Recently, VLC technology using white LEDs (wavelengths from 380 nm to 700 nm) is gaining attention in academia and industry, driven by progress of white LED technology for SSL and the potential for simultaneously using such LEDs for wireless data transmission. Generally, white LEDs used are classified into two types, namely trichromatic and blue-chip LEDs [31–33]. Trichromatic LEDs are fabricated by mixing light of the three primary colors (red, green, and blue) obtained using three different LED chips (see the spectrum in Figure 1.3(a)). This type is mainly used for architectural design purposes and is more expensive compared to blue-chip LEDs. Blue-chip LEDs have a phosphor layer on top of a single blue chip and they are typically found in most white LED bulbs available in the market (spectrum shown in Figure 1.3(b)). In general, LOS links are the most commonly used links for VLC (see Figure 1.2(a) and Figure 1.2(b)). Trichromatic LEDs are investigated by Tanaka et al. and simulation results for data rates up to 400 Mbps are reported [34]. The maximum measured data rate for a VLC system using blue-chip LEDs is reported in [33,35] where a modified version of the classi-
1.2 Review on Indoor Optical Wireless Systems

![Diagram showing wavelength for LED chips and phosphor layers.](image)

Figure 1.3: (a) Trichromatic white LEDs, (b) Single-chip white LEDs. The trichromatic LED has peaks for red, green, and blue. The blue-chip LED has a dominant peak and a lower wider peak from 500 nm to 600 nm.

Cal orthogonal frequency division multiplexing (OFDM) modulation technique is considered to achieve data rates higher than 100 Mbps.

1.2.2 Technology Potentials and Possible Applications

Next generation wireless communication systems (4G and beyond) will be based on several complementary access technologies and OW is expected to form an essential part in the 4G and beyond visions [36]. For example, bandwidth-intensive applications such as Internet multimedia streaming perform more effectively over networks with high bandwidth and data throughput. These requirements cannot be met with the capabilities of the radio frequency (RF) technology because of the restricted spectrum availability and interference [37]. Instead, OW technology can be considered motivated by several benefits as compared to the RF technology. These benefits include:

- Unregulated huge (THz) bandwidth.
- License-free operation.
• Low-cost analog front ends.

• No interference with RF based systems which makes it a preferred solution for electro-magnetic interference (EMI) sensitive operating environments as in airplanes and hospitals.

• It is free of any health concerns as long as eye and skin safety regulations are fulfilled [38, 39].

• No inter-cell interference transpires as the optical signals do not penetrate through walls and, thus, high degree of privacy and security against eavesdropping is inherently offered.

White LED bulbs pose clear advantages over conventional incandescent and fluorescent lighting sources in power efficiency and long life-time expectancy, which makes them a strong candidate for future illumination equipments. For example, the European Commission has decided recently to prohibit the sale of particularly energy-intensive lamps for household use in a series of stages up to 2016 [40]. As soon as high efficient white LED bulbs are manufactured economically enough to overtake the currently favored compact fluorescent bulbs, the same bulb could simultaneously be used as an optical access point (OAP) to provide portable and fixed devices with wireless access. These LED bulbs can be modulated at rates many times that of incandescent or fluorescent alternatives, thus offering the possibility of broadband transmission at the same time as providing illumination [32, 41]. The function of illumination is not affected by the envisaged piggy-backed communication as the blinking rate of the intensity modulated light is sufficiently rapid and cannot be detected by the human eyes.

It is predictable that indoor OW applications will most likely be based on white LEDs rather than on IR LEDs [42]. VLC offers many advantages over IR transmission:

• Data transmission along with the illumination of rooms and different interior spaces.

• The installation of a wireless network based on an existing interior lighting infrastructure would probably be easier and more cost effective than setting up a separate IR network.
The signal is less likely to be obstructed and the LOS component is dominate in most positions in the room (the effect of multipath is small) because of the distributed ceiling installations. The estimated bandwidth is at least 88 MHz [43].

High signal-to-noise ratio (SNR) is obtainable, which is an indirect consequence of the illumination requirements (>60 dB through the entire room) [33].

Especially for optical OFDM (O-OFDM), a DC bias that carries no information is necessary for both IR and VLC. However, in VLC, it is required for illumination and may not cause any severe power efficiency loss.

A possible scenario of future wireless communication is depicted in Figure 1.4. Service is established through a combination of wired and wireless technologies. The wired base stations (BSs) are merged with the LED based illumination equipment to provide wireless network access. Data broadcasting through a ceiling bulb realizes a point-to-multipoint connection and a focused spotlight realizes a point-to-point connection. User requests are sent through an uplink channel which offers mobility compared to a fixed terminal scenario as shown in Figure 1.4. The power over Ethernet (PoE) technology can be used to transport data traffic and supply the BSs as well as the lamps with the required power.

VLC is becoming well-liked in various operating environments due to the aforementioned advantages. Local information points in public areas, e.g. airports, train

Figure 1.4: Future wireless networking scenario. A mixture of VLC and PoE technologies provides the combination of back-haul and short-range wireless access.
stations, underground railway stations, commercial centers, medical facilities, are considered as potential areas where this technology can be used. VLC is considered as a candidate for broadband and electrosmog-free wireless home networking (WHN) [43]. In addition, VLC can grant wireless access to variety of services using the LED reading lamps in airplane cabinets, passenger trains, and coach buses [44–47]. The technology can serve an assortment of applications ranging from inter-system communication, information provision for maintenance, advertising and augmented reality, entertainment systems in the transport sector, real-time travel updates, web browsing, wireless connection of sensor networks, intelligent home control systems, etc.

1.2.3 Design Challenges

Although the IR or visible light band promises significant advantages as a medium for indoor communication, they also have drawbacks. The key design challenges to achieve high-speed OW transmission indoors are in achieving a sufficiently high SNR at useful data rates and due to the multipath channel dispersion [48–51]. Both challenges are greatly influenced by the link configuration.

As for the SNR, the difficulty evolves because of limited average power level mostly due to eye and skin safety requirements, high free space propagation losses, shadowing, the square-law reception, and noise/interference from background ambient light [52]. Optical path-loss of a diffuse IR links employing a Lambertian transmitter is measured between 50 dB to 80 dB. Based on the link configuration, shadowing can contribute up to 10 dB to the path-loss [53]. Due to the square-law relation between optical power and electrical power, the electrical path loss is twice the optical path-loss in dB. This means that a 50 dB optical path-loss corresponds to 100 dB electrical path-loss, i.e. the receiver electrical SNR is proportional to the square of the average received optical power where a 1 dB change in the average power corresponds to a 2 dB change in the electrical SNR. This implies that the system should be designed to minimize path-loss and should employ a large area PD to collect more photons. The problems resulting from high-capacitance large-area PDs, such as noise and restricted bandwidth, can be limited by the receiver circuit topology [54].

The main components of noise in OW systems are shot noise, receiver noise,
1.3 Optical Modulation

and periodic noise. Shot noise is due to background light sources, such as sunlight, fluorescent lamp light, and incandescent lamp light [55, 56]. Typical intensity levels of the ambient light are usually much higher than data signal intensity levels. Receiver noise is due to thermal effects in the receiver circuitry, and is particularly dependent on the type of amplifier used. When the ambient light is weak, the receiver noise becomes the dominant factor. Periodic noise is the result of the periodic variations of fluorescent light intensity due to the method of driving the lamp using the ballast [57, 58]. This generates a periodic signal with a fundamental frequency together with significant harmonics to several MHz [21]. The usual approach to diminish noise is optical filtering [59]. Electrical filtering can be used to reduce the effect of illumination harmonics, but at the cost of inducing baseline restoration [60]. In VLC, a blue filter removes the yellow phosphor spectrum component creating shot noise. Also by careful selection of the modulation technique, noise or interference can be reduced.

Signal degradation also arises from multipath dispersion induced ISI. The ISI problem is significant at high data rates and is influenced by the room size and the reflection coefficients of the surfaces inside the room. Although, optical links using IM with DD are immune to multipath fading, multipath dispersion is still present since the light can arrive at the receiver from multiple paths of different lengths. Measurement and simulation results demonstrate a channel delay spread in the order of 30 ns to 50 ns. Relatively moderate ISI leads to an optical power penalty, but severe ISI may lead to a BER floor [21]. Diffuse systems are more prone to multipath effects than directed beam systems. This is because of their larger beam widths leading to more potential reflectors, and the larger FOV of their detectors resulting in more reflected light being detected.

1.3 Optical Modulation

1.3.1 Pulsed Modulation Techniques

Evaluation metrics for optical modulation techniques may include power efficiency, i.e. the electrical receiver SNR required to achieve a desired bit-error ratio (BER), bandwidth efficiency, robustness to multipath distortion, implementation complexity, immunity against interference sources, and, in networks, the suitability to
Introduction

Multiple user access. All these metrics are greatly influenced by the optical link configuration.

There are different modulation schemes which are suitable for OW communication systems each with its own advantages and disadvantages. A vital criterion for evaluating optical modulation techniques is power efficiency. High average power efficiency can be achieved by employing single carrier (SC) pulsed modulation techniques in which the time dependent characteristics of the optical pulse is used to convey information. Among several techniques, two schemes are widely used, namely on-off keying (OOK), and pulse-position modulation (PPM) [61]. OOK in its various flavors (such as the return-to-zero (RZ) and the non-return-to-zero (NRZ)) is one of the oldest formats and is the simplest in terms of hardware implementation and integration. It also exhibits a good compromise between complexity and performance. Within a symbol time, a positive intensity pulse represents a bit ”1” and the absence of that pulse represents a bit ”0”. The duty cycle \( D \) determines the bandwidth and average power requirements. The OOK-NRZ has \( D = 1 \) and the normalized bandwidth and the normalized average optical power requirements are 1 and 0, respectively. Reducing the duty cycle \( D \) leads to a bandwidth increase by a factor of \( 1/D \) and required power decreases by \( 5 \log_{10}(D) \) [21, 62]. OOK is not efficient at very small duty cycles, thus it is more appropriate to code the information into the position of the pulse such as in PPM.

In PPM, an optical pulse is transmitted in one out of \( S \) slots per symbol time. The occupied slot position denotes the bit combination conveyed by the symbol. PPM expands the signal bandwidth compared to OOK, but provides higher power efficiency. For example, the normalized bandwidth requirement for PPM (normalized to OOK-NRZ) is \( S/\log_2(S) \), but the normalized average optical power requirement for PPM is \(-5 \log_{10}(S \log_2(S)/2)\) for soft decoding [21, 62]. Besides, the use of PPM imposes more system complexity compared to OOK since both slot- and symbol-level synchronizations, critical to system performance, are required at the receiver.

Multipath propagation induces ISI and PPM is particularly sensitive to these dispersive effects of the optical channel due to the required bandwidth [63]. On ISI channels, unequalized SC pulsed modulation suffers severe performance penalties. Several equalization techniques, for example, using linear and decision-feedback equalizers (DFEs) are considered to mitigate the effects of ISI at high
1.3 Optical Modulation

Figure 1.5: Demonstrating SC pulsed modulation schemes, OOK and PPM.

data rates [21]. Various methods are tested to combat channel impairments. Each method has its advantages and disadvantages and their effectiveness depends on the used modulation technique. For detailed explanation and evaluation of these and other techniques, the reader is kindly referred to [21, 64–66]. Noteworthy is the fact that the near IR-region and visible light are close together in wavelength, and they exhibit qualitatively similar behavior. Therefore, most experimental and simulation results obtained based on a specific technology are significant for both technologies.

As for the suitability to multiple user access, if a single high-speed wireless link is used to convey multiplexed lower-bit-rate data to several users, SC pulsed modulation presents a drawback. The reason is that each receiver is required to detect the aggregate high-speed bit stream and perform digital demultiplexing to obtain the desired data. The drawbacks of SC pulsed modulation are overcome by using an alternative modulation technique; that is, multiple-subcarrier modulation (MSM).

1.3.2 Optical OFDM

MSM technique based on single-subcarrier modulation (SSM) is proposed in [67]. In SSM, a bit stream is used to modulate, typically, a sinusoid signal (see the example for the binary phase shift keying (BPSK) in Figure 1.6). A DC bias is required (above the LED turn-on voltage (TOV)) to ensure that the instantaneous optical power is positive (to avoid signal clipping at the TOV). Following this principle, MSM generates multiple single subcarriers and multiplexes a certain
number of subcarriers in the electrical domain during each symbol interval. After adding an appropriate DC bias which carries no information, the signal is used to modulate the optical carrier (see Figure 1.6). MSM can be more bandwidth-efficient than OOK and PPM, and can provide immunity to interference near DC, for example, from fluorescent lamps. MSM also reduces the symbol rate of each subcarrier relative to that of a SC, and therefore promises to minimize ISI on multipath channels [68]. Another advantage of MSM over SC modulation is evident when it comes to multiple user access since each user can process and recover specific bit streams using bandpass demodulators.

OFDM is a practical realization of MSM. OFDM is a parallel data transmission on orthogonal subcarriers. O-OFDM systems are able to support high data rates by using high order multi-level quadrature amplitude modulation (M-QAM) without the need of complex channel equalizers. Attaching a cyclic prefix (CP) to the transmitted OFDM symbols converts the linear convolution of the channel with the OFDM signal to a circular convolution. As a result, the time-varying channel can be easily estimated using frequency-domain channel estimation and simple frequency domain equalizer can be employed. The CP also acts as a guard interval to avoid ISI due to multipath dispersion effects, i.e. resulting in inherent robustness of OFDM against multipath effects. Adaptive modulation per sub-
carrier can be applied based on the uplink/downlink requested data rates and quality of service (QoS) [69]. Also, the possibility to combine it with any multiple access scheme, e.g. time division multiple access (TDMA), frequency division multiple access (FDMA), and code division multiple access (CDMA) [25], makes it an excellent choice for indoor OW links.

In an OFDM system, the serial stream of data and redundancy bits (coding bits, pilots for channel estimation, etc.) is demultiplexed into parallel streams. After being modulated, each symbols stream is transmitted on a separate subcarrier as shown in Figure 1.7. The inverse fast Fourier transform (IFFT) operation modulates, multiplexes the subcarriers, and generates the time domain OFDM symbol,

\[ x_k = \frac{1}{N} \sum_{n=0}^{N-1} X_n \exp \left( j \frac{2\pi}{N} nk \right) \]  

(1.1)

where, \( j \) is the imaginary unit, \( x_k \) with \( k = 0, \ldots, N - 1 \), are the \( N \) time-domain output samples, and \( X_n \) with \( n = 0, \ldots, N - 1 \), are the input symbols.

![Figure 1.7: An optical OFDM transmitter. The analog OFDM signal at the digital-to-analog (DAC) output modulates the intensity of the transmitter LED. Number of samples forming one OFDM symbol corresponds to the IFFT/FFT length or number of subcarriers.](image)

In general, the output signal of the OFDM modulator is complex. In IM optical
systems, quadrature modulation is not possible; this means that this signal must be real. Therefore, the OFDM chain commonly used in RF communications must be modified. To fulfil the requirement of having a real-valued signal from the OFDM modulator, a complex conjugate input data symbols (e.g. \( X_n = X_{N-n}^* \)) are used at the IFFT input to produce a real time-domain output signal as illustrated in Figure 1.8.

![Figure 1.8: IFFT bin assignment and time domain O-OFDM signal. \( T_s \) is the sampling clock period and \( 1/(N \times T_s) \) is the subcarriers separation.](image)

The symbols \( X_0 \) and \( X_{N/2} \) are set to zero to ensure that the output consists of only real values. Two forms of optical OFDM, namely, DC biased optical OFDM (DCO-OFDM) and asymmetrically clipped optical OFDM (ACO-OFDM) are reported in literature.

### 1.3.3 DCO-OFDM

DCO-OFDM [34, 70, 71] assigns data to subcarriers as follows,

\[
\begin{bmatrix}
0 & X_n & 0 & X_{N-n}^*
\end{bmatrix}
\quad n = 1 : 1 : \frac{N}{2} - 1
\quad (1.2)
\]

where, \( N \) is the number of available subcarriers.
Data symbols are assigned to all odd and even subcarriers. The data rate of DCO-OFDM system is given by,

\[
R^{\{DCO\}} = \left( \frac{N/2 - 1}{N + N_g} \right) B \log_2 M \text{ bits/s.} \tag{1.3}
\]

where, \(B\) is the channel bandwidth, \(N_g\) is the number of guard subcarriers, and \(M\) is the size of the considered constellation diagram.

This means that a maximum of \((N/2) - 1\) subcarriers out of \(N\) subcarriers can be used to carry useful data. As shown in Figure 1.8, the DCO-OFDM has \((N/2) - 1\) independent complex inputs and the bipolar time domain signal is used to modulate a DC biased LED. This DC bias which is carrying no information results in optical power efficiency penalty, mainly in IR transmission. However, in VLC, a DC bias point is required for illumination and may not cause severe power efficiency loss. The DC bias value depends on the LED characteristics and can significantly affects system performance [72, 73].

### 1.3.4 ACO-OFDM

In ACO-OFDM, only odd subcarriers are modulated as follows,

\[
\begin{bmatrix}
0 & X_n & 0 & X_{N-n}^* \\
\end{bmatrix}
\quad n = 1 : 2 : \frac{N}{2} - 1
\tag{1.4}
\]

Therefore, the achieved data rate for ACO-OFDM systems is given by,

\[
R^{\{ACO\}} = \left( \frac{N/4 - 1}{N + N_g} \right) B \log_2 M \text{ bits/s.} \tag{1.5}
\]

ACO-OFDM uses only half of the sub-carriers used by the DCO-OFDM to carry data, i.e. there are only \((N/4) - 2\) independent complex inputs. In contrast to DCO-OFDM, the generated bipolar signal is converted to unipolar through clipping of all negative values at zero before modulating the LED. It has been shown in [74] that the negative clipping noise in ACO-OFDM system falls on the even subcarriers only. Hence, it is orthogonal to the transmitted data on the odd subcarriers and has no significant impact on the performance.

ACO-OFDM systems have several advantages as follows,
• A DC bias is avoided which can be considered as an extra power consumption.

• A larger amplitude of the signal can be considered which covers the full dynamic range of the LED.

Conversely, ACO-OFDM systems suffer from several drawbacks as follows,

• ACO-OFDM systems sacrifice a large portion of the data rate to achieve the asymmetrical property.

• For VLC systems, a DC bias is needed for lighting purposes which makes the clipping at the zero level ineffectual.

1.4 Thesis Contribution

The bipolar time domain O-OFDM signal exhibits a random-like variations. The signal has a Gaussian distribution for a sufficiently large number of subcarriers and is characterized by a high peak-to-average power ratio (PAPR). This signal modulates the LED intensity as shown in Figure 1.9. Therefore, special attention must be paid to its sensitivity to LED nonlinearities. As explained in section 2.2.1, the LED nonlinear behavior degrades the OFDM signal through amplitude distortion, clipping of the lower peaks, and clipping of the upper peaks. The lower peaks are clipped at the TOV of the LED and the upper peaks are clipped at the saturation voltage which corresponds to the maximum permissible AC/pulsed saturation current of the LED recommended by the manufacturer. The clipping of the upper peaks ensures that the LED chip does not overheat, in order to avoid degradation in output light or, in the worst case, total failure. For example, Figure 1.10 demonstrates the clipping effects on the OFDM spectrum and the estimated symbols. Figure 1.10(a) shows the spectrum of the OFDM signal through ideal transmission chain and the distorted spectrum due to the clipping of the upper and lower peaks of the same OFDM signal. The corresponding time domain signals are shown in Figure 1.10(b). Finally, in Figure 1.10(c), the transmitted symbols per subcarrier and the corresponding estimated symbols are plotted. The effect of the signal clipping on the estimated symbols is obviously noticed.
Figure 1.9: Demonstrating the O-OFDM chain using IM with DD.

Figure 1.10: Demonstrating the clipping effect on the O-OFDM signal in time and frequency domains and on the symbol error. The IFFT length is 64 and uncoded 4-QAM constellations are considered.
In contrast to former studies which only address the clipping of the negative signal peaks at the TOV, the key objective of this thesis is to study the in-band distortion effects of LEDs on the overall performance of OFDM optical systems as a result of amplitude distortion, clipping of the lower peaks, and clipping of the upper peaks. Based on a measured static transfer function of the LED, a polynomial is used as the model of the transfer function. This approach has been used to model the nonlinearity of LEDs or LDs. In order to quantify the LED generated distortion, the deviation of the estimated constellations from the ideal constellations is represented by the error vector magnitude (EVM) as a distortion metric. To identify the quality of transmission in the presence of the LED nonlinearity in an additive white Gaussian noise (AWGN) environment (modeling the shot noise and thermal noise at the analog optical receiver), the bit-error performance is obtained as a function of several parameters such as the LED bias point, the average OFDM signal power modulating the LED, and the constellation orders.

Backing off the average power of the OFDM signal modulating the LED and the reduction of the PAPR are widely used approaches to control the generated distortion. However, this thesis is focused on applying a digital predistorter as an LED linearization technique. Key features of the predistorter reside in the use of the LED inverse characteristics as distortion compensator. The approach to obtain the predistorter polynomial is described. The bit-error performance after compensation is obtained as a function of the LED bias point, the average OFDM signal power modulating the LED, and the constellation orders.

As part of the distortion analysis, a model to describe the LED transfer function while considering the OFDM signal clipping is proposed. Also the concept of soft-clipping (rather than limiting the signal peaks) the upper peaks using the proposed model is introduced. The model is suitable for conducting numerical analyzes and Monte Carlo simulations. In practical systems, the model can be used to clip/condition the OFDM signal prior to LED modulation. In this context, methodologies that allow the analytical evaluation of the bit/symbol error probability of uncompensated/compensated optical communication systems are presented. A comparison with Monte Carlo simulation results is carried out to verify the accuracy of these methods.

In addition, proof-of-concept hardware demonstrators for visible light data broadcasting with focus on indoor environments are realized. Throughout the
simulations and the experimental measurements, simple and low cost optical carrier modulation and demodulation through IM and DD are considered. Analog front ends required for the hardware demonstrators are developed (schematics design, layouts design, and PCB fabrication). A visible light transmitter and receiver are designed and assembled to cover 100 kHz signal bandwidth.

The link chain includes forward error correction (FEC) coding, frame synchronization routines, and pilot symbols for channel estimation and symbol equalization. The complete chain including the transmitter and the receiver is running on digital signal processing (DSP) boards. The transmission is based on the assumptions of direct LOS and simplex channel conditions. The measured BER results under moderate indoor ambient light conditions are reported. The target is not to showcase high data rates; rather to study via a simple proof-of-concept hardware demonstrator achievable rates for phase-incoherent optical OFDM and to investigate performance for different electrical SNRs. The current bandwidth is limited by the onboard audio codec but is sufficient for messaging or information services, several audio channels, low quality video streaming applications.

1.5 Thesis Structure

This chapter has provided a brief introduction to the history of OC. The development of indoor OW communication has been reviewed. The potential of indoor OW communication has been highlighted. The implementation of the O-OFDM system has been discussed. The reason for choosing the LED nonlinearity study as the subject to be investigated in this PhD project has been motivated.

Chapter 2 presents an indepth analysis of the effect of LED nonlinearity on system performance. The proposed linearization technique is explained.

Chapter 3 presents the analytical approach and the Monte Carlo simulation results to evaluate the bit/symbol error probability.

Chapter 4 is dedicated to the development of hardware demonstrators and the conducted experimental measurements.

The objective of the last chapter is to conclude the work presented, to point out potential future work, and to discuss limitations of this work.
1 Introduction
2 Distortion Characterization and Linearization Technique

2.1 Introduction

Communication systems, in general, suffer from the nonlinearity of devices in electronic circuits, which is a design challenge specially for the orthogonal frequency division multiplexing (OFDM) technique due to its particular sensitivity to nonlinear distortion. When the OFDM signal with its envelope variations passes through a nonlinear device, such as a power amplifier (PA) of a radio frequency (RF) transmitter or a light emitting diode (LED) of an optical transmitter, it undergoes in-band distortion, which creates inter-carrier interference (ICI) and degrades the bit-error performance.

The general aim of this work is to gain insight into the effects of LED nonlinearities on bit-error performance degradation. With this knowledge, it is possible to control the induced nonlinear distortion. This is done through a proper setting of the characteristics of the time domain electrical OFDM signal modulating the LED intensity (in this work, varying the average OFDM signal power calculated over a single OFDM symbol is considered) and a proper driving of the transmitter LEDs, i.e. proper setting of the bias point. Therefore, it is very important to first characterize the induced nonlinear distortion, which is carried out in this chapter.

In this context, the behavior of a commercially available IR LED is modeled using a polynomial equation. This polynomial equation is integrated within a Matlab/Simulink based O-OFDM model, which is introduced in section 2.3. Simulations are conducted to quantify the induced nonlinearity through error vector magnitude (EVM) as nonlinearity indicator. The EVM metric is briefly intro-
duced in section 2.3.2. Throughout the simulations, the OFDM signal is clipped before the LED polynomial at the turn-on voltage (TOV) and at the saturation voltage which corresponds to the maximum permissible AC/pulsed current.

Additionally, the bit-error ratio (BER) is obtained via simulations in an additive white Gaussian noise (AWGN) environment for several bias points and as a function of the average electrical OFDM signal power modulating the LED. It is found that the induced nonlinear distortion can drastically limit the system performance. This has led the author to propose a linearization technique; that is, a digital predistorter to improve the linearity and to reduce the LED induced distortion.

In section 2.2.3, the procedure to obtain the polynomial equation describing the predistorter for the same LED is explained. The predistorter is applied to the system and via simulations, the BER results after compensation are presented. It is shown that the performance of the compensated system is tremendously enhanced.

### 2.2 Polynomial Modeling Approach

#### 2.2.1 LED Nonlinear Behavior

Each LED has a minimum threshold value known as the TOV which is the onset of current flow and light emission (below the TOV, the LED is considered in a cut-off region and is not conducting current) \[75\]. Above the TOV, the current flow and light output increase exponentially with voltage (current conduction region) (see Figure 2.1). The LED outputs light that is linear with the drive current. However, thermal aspects causing a drop in the electrical-to-optical conversion efficiency (light output of the LED decreases and slowly approaches a steady state value; LEDs self heating characteristic) must be considered \[75\]. The DC and AC/pulsed currents must be adjusted according to the manufacturer data sheet \[76\] to ensure that the LED chip does not overheat, in order to avoid degradation in output light or, in the worst case, total failure. These current values depend very much on the diode type and varies depending on the semiconductor materials, packaging, and ambient temperature.

In OFDM optical systems, the real-valued time domain OFDM signal is used
2.2 Polynomial Modeling Approach

Figure 2.1: Nonlinear and linearized LED transfer characteristic. The non-linear transfer characteristic distorts the OFDM signal in optical applications. Linearization through predistortion helps to improve system performance.

In particular, for visible light communication (VLC), the bias point should be set to achieve the minimal brightness required for illumination according to lighting standards [77]. Another form of O-OFDM converts the bipolar signal to unipolar through clipping of all negative values at zero [74]. However, only half of the subcarriers are used to carry data.

Beside a proper setting of the DC bias point, several techniques can be applied to reduce the possibility of clipping of the bipolar OFDM signal peaks [78]. Conventionally, backing-off the average power of the OFDM signal ensures that the LED operates in a quasi-linear region of operation [72]. Peak-to-average power ration (PAPR) reduction techniques can be considered to reduce power back-offs
levels in optical systems [79]. However, neither power back-off nor PAPR reduction techniques necessarily result in an improvement in system performance and trade-offs must be considered. Power back-off might result in a significant power efficiency penalty and can significantly compromise signal coverage [80]. PAPR reduction techniques increase the system complexity and/or sacrifice bandwidth efficiency [81, 82]. Instead, a linearization through a predistorter as proposed in this thesis [83] can compensate the LED nonlinear curve as shown in Figure 2.1.

2.2.2 LED Polynomial Model

A high power IR LED (SFH 4230 from OSRAM) is considered [76]. The relation between the forward voltage across the LED and the current through the LED is modeled through a polynomial using the least-square curve fitting technique. A polynomial of the sixth degree is used to model the LED transfer characteristic shown in Figure 2.3. The curve in Figure 2.4 shows the nonlinear behavior of the LED using the developed polynomial for forward voltage amplitudes in the range from 1.3 V up to 2.1 V. The LED TOV is considered to be at 1.3 V. At 2.1 V (saturation voltage), the forward current is considered to be the maximum permissible AC/pulsed current. Throughout the simulations, the OFDM signal amplitudes below 1.3 V and above 2.1 V are clipped before the LED polynomial.
2.2 Polynomial Modeling Approach

Figure 2.3: The V-I data sheet curve for the high power IR LED (OSRAM, SFH 4230).

Figure 2.4: The V-I curve using the developed LED polynomial function for the high power IR LED (OSRAM, SFH 4230).
2.2.3 Predistorter Polynomial Model

The predistorter uses the LED inverse characteristics as nonlinear compensator to condition the OFDM signal prior to the LED modulation. The proposed procedure to model the LED and its predistorter is valid for any LED. For the considered IR LED, predistortion linearizes the LED response over the range from 1.3 V up to 2.1 V. The solid curve in Figure 2.5 illustrates the linearized V-I relation. It is important to note that a linear region of operation without a predistorter is between 1.6 V and 1.8 V. Hence, a 1.7 V bias point is recommended for the considered LED.

The idea of the predistortion is illustrated on the same figure. Assuming $v_{in}$ is the input signal amplitude and $i_{out-pd}$ is the desired output current known from the linear response. Then, the original input amplitude, $v_{in}$, is adjusted to produce $v_{out-pd}$ which produces the correct output current amplitude, $i_{out-pd}$, that gives the overall predistorted-LED chain a linear response.

Through predistortion, a linear response curve is achieved over a large range of the input signal amplitudes. However, the region which can be linearized is limited. The maximum input amplitude that will be modulated linearly depends upon the maximum permissible AC/pulsed current through the LED. Therefore
and for the considered LED, the OFDM signal amplitudes below 1.442 V and above 1.961 V are clipped.

The polynomial for the predistorter is obtained by the following procedure:

- Obtain the polynomial equation, \( f(v) \) (current through the LED as a function of the voltage across the LED), using the measured data of the LED forward voltage and forward current relation. See Figure 2.6(a).

- Obtain the polynomial equation, \( f(i) \) (voltage across the LED as a function of current through the LED), using the same electrical measurements. See Figure 2.6(b).

- Determine the polynomial equation for the required linearized voltage to current relation. See Figure 2.6(c) where the curve illustrates the linearized behavior of the LED.

- Substitute in \( f(i) \) the forward current values in the linearized range (1.442 V-1.961 V) to obtain the corresponding values of the forward voltage. See Figure 2.6(d).

- Obtain the predistorter polynomial equation using the values of the forward voltage obtained in the previous step. See Figure 2.6(e).

- Figure 2.6(f) shows the input signal before the distorter (x-axis) and the current through the LED (y-axis). Figure 2.6(f) demonstrates exact match with Figure 2.6(c).

2.3 Simulation Model

2.3.1 Optical OFDM Building Blocks

The OFDM simulation model is shown in Figure 2.7. The predistorter and the LED are modeled through the V-V and the V-I blocks, respectively. Shot noise and thermal noise at the reciever are modeled as AWGN [48].

The model continuously generates a random stream of bits. Data protection is realized through the use of forward error correction (FEC) coding (convolutional
Figure 2.6: (a) V-I curve of the LED polynomial function, \( f(v) \). (b) I-V curve of the LED polynomial function, \( f(i) \). (c) V-I curve of the linearized LED polynomial function. (d) I-V curve of the linearized LED polynomial function. (e) The predistorter polynomial function. (f) The linearized response of the cascade predistorter and the LED under investigation.
encoder) and interleaving. Different constellation orders are considered. The generated serial stream of symbols at the modulator output is split into parallel streams, each is transmitted on a separate subcarrier. The inverse fast Fourier transform (IFFT) operation is used to modulate the available subcarriers and to generate the time domain OFDM signal. At the input of the IFFT, complex conjugate data symbols are used to produce a real time domain output. For the purpose of channel estimation, a single OFDM symbol as block-type pilots is used [84–86]. The complex conjugate requirement is fulfilled while generating the pilots symbol. The OFDM frame is formed by one OFDM symbol for channel estimation and 20 OFDM symbol with data subcarriers. Attaching a cyclic prefix (CP) to the transmitted OFDM symbols converts the linear convolution of the channel with the OFDM signal to a circular convolution [87]. As a result, simple frequency domain equalizer can be employed. Frequency domain equalization is realized using conventional OFDM zero-forcing detection. The equalized symbols are demodulated and deinterleaved. Finally, the encoded symbols are decoded by the Viterbi hard-decision algorithm to obtain the estimated stream of data bits. In addition, the OFDM demodulator calculates the EVM used as a distortion indicator.

2.3.2 EVM Calculation and Model parameters

The error vector is a common figure of merit for system linearity in digital wireless communication standards. It is a measure of the fidelity of a digital communication system and is related to in-band distortion and signal-to-noise ratio (SNR) [88]. On a constellation diagram, the error vector is a measure of the departure of signal constellation points from its ideal reference as shown in Figure 2.8. The error vector is the scalar distance between the ideal constellation vector and the measured vector of the displaced constellation point after it has been compensated in timing, amplitude, frequency, phase, and DC offset. The EVM is the root mean square value of the error vector over time. To calculate the EVM, the model uses the recovered constellations to regenerate the ideal constellations. The EVM is calculated by subtracting the recovered constellations from the corresponding ideal references, taking the absolute values and calculating the root-mean-square (RMS) value over one OFDM symbol.

Important simulation parameters are listed in Table 2.1.
Figure 2.7: The building blocks of the optical OFDM simulation model implemented in Matlab. The LED characteristics and the sub-carriers separation is equal to \( \frac{f_s}{N} \)where \( f_s \) is the sampling frequency, \( N \) are the number of the IFFT bins (IFFT length, or the OFDM sub-carriers), \( f_{Nyquist} \) is the Nyquist frequency, and the sub-carriers separation is equal to \( \frac{f_s}{N} \).
2.4 Results

Simulations are conducted to investigate the influence of the bias point on the generated distortion due to the LED nonlinearity. In these simulations, the AWGN channel is not considered and a BPSK (binary phase shift keying) modulation scheme with 3/4 channel coding rate is used. The average power of the OFDM signal modulating the LED is calculated over one OFDM symbol. In all figures, a 20 mW electrical average power is considered the 0 dB power back-off. The power back-off indicates relative decrease in the signal power to the initial signal at 20 mW. The distortion is characterized by the EVM in percentage which is also computed over one OFDM symbol.

At 0 dB power back-off, the obtained EVM values for 100 OFDM symbols without the predistorter and after applying the predistorter are plotted in Figure 2.8(a)

<table>
<thead>
<tr>
<th>Table 2.1: Simulation model parameters</th>
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<tbody>
<tr>
<td>OFDM</td>
</tr>
<tr>
<td>IFFT length</td>
</tr>
<tr>
<td>Data subcarriers</td>
</tr>
<tr>
<td>CP samples</td>
</tr>
<tr>
<td>Pilot symbols</td>
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<tr>
<td>Data symbols</td>
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</table>

Figure 2.8: Nonlinearity effects on the constellations. A constellation diagram for a QPSK modulation with 3/4 channel coding rate and 15 dB SNR.
and Figure 2.9(b), respectively. Before applying the predistorter (Figure 2.9(a)), the 1.7 V bias point achieves the lowest EVM floor and EVM peak values among the bias points under investigation. The EVM floor is defined as the lowest EVM value whereas the EVM peak is defined as the highest EVM value in a burst of 100 OFDM symbols. Although the 1.725 V bias point achieves fair EVM floor and EVM peak values, it will not be considered later for system performance evaluation since the forward current at 1.725 V is 1.07 A which is above the maximum permissible DC current (1 A according to the data sheet). Although a low bias point improves the power efficiency, it is noticed that the lowest bias point, 1.6 V, has the highest EVM floor and EVM peak values compared to the other bias points under investigation and the system is expected to show the worst bit-error performance at this bias point. In Figure 2.9(b), the predistortion indeed achieves better EVM floor values for the 1.7 V and the 1.675 V bias points. However, degradation is noticed at the 1.6 V, 1.625 V, and 1.65 V bias points. This can be related to the fact that the input signal amplitudes above 1.961 V and below 1.442 V are clipped in the presence of the predistorter while amplitudes above 2.1 V and below 1.3 V are clipped in the absence of the predistorter. Therefore, at high inputs signal powers (20 mW), signal clipping distortion is expected to dominate the bit-error performance rather than amplitude distortion.

In addition to linearization with the applied predistorter, different power backoff values are applied to the 20 mW OFDM signal to investigate the influence of the power reduction on the generated distortion. For example, Figure 2.10(a) and Figure 2.10(b) show the obtained EVM values at 2 dB power back-off. Both EVM floor and EVM peak values with and without the predistorter are significantly improved for all bias points. However, EVM floor values still exist without the predistorter while an almost 0% EVM floor is noticed for all biased points with the predistorter, except for the 1.6 V bias point which shows an EVM floor value greater than 3%.

Figure 2.11(a) and Figure 2.11(b) show the average EVM values without predistortion and with predistortion, respectively, over 1000 simulated values of all bias points under investigation for power back-offs in steps of 1 dB up to 8 dB. As expected, the EVM values at 0 dB are better without the predistorter for the 1.6 V, 1.625 V, and 1.65 V bias points. With the predistorter, however, slight improvement in the EVM values of the 1.7 V and 1.675 V bias points is noticed.
Figure 2.9: The EVM scatter plot of 100 OFDM symbols versus bias points at 0 dB power back-off. (a) without using a predistorter. (b) with predistorter.

Figure 2.10: The EVM scatter plot of 100 OFDM symbols versus bias points at 2 dB power back-off. (a) without using a predistorter. (b) with predistorter.
At 2 dB power back-off, less than 10% EVM is achieved with the 1.7 V, 1.675 V, and 1.65 V bias points without the predistorter, while the 1.625 V and the 1.6 V bias points achieve around 13% EVM and 18% EVM, respectively. Correspondingly, with the predistorter all bias points achieve EVM values less than 10%. A 0% EVM is noticed for almost all bias points with the predistorter at 3 dB power back-off, while 8 dB power back-off value is needed for the 1.7 V bias point to achieve a similar EVM value without using the predistorter.

In order to study the effect of LED nonlinearity on bit error performance, first, simulations are conducted without the LED model (only the AWGN channel model is considered) to determine the required SNR to achieve a target BER for two modulation schemes under investigation, BPSK and 64-QAM (quadrature amplitude modulation). The curves are depicted in Figure 2.12 and the required
SNR values to achieve approximately $2.5 \times 10^{-5}$ BER are shown on the figure. For example, MPEG-4 video transmission has slight visible degradation at $10^{-5}$ BER [89].

![Figure 2.12: BER versus SNR using AWGN channel model and 3/4 convolutional channel coding rate for BPSK, and 64-QAM. The approximate SNR values required to achieve a target BER of $2.5 \times 10^{-5}$ are 6 dB, and 22 dB for BPSK, and 64-QAM, respectively.](image)

Using the SNR values from Figure 2.12, the BER and EVM for 1000 OFDM symbols (more than $10^6$ bits) are simulated in the presence of the LED non-linearity and in AWGN environment. The BER and EVM simulation results for BPSK without predistortion and BPSK with predistortion are shown in Figure 2.13. Figure 2.14 shows the BER and EVM simulation results for the 64-QAM without predistortion and 64-QAM with predistortion. The effect of LED non-linearity is obvious in all figures and the degradation in BER performance is consistent with the obtained EVM values.

In Figure 2.13 and at 0dB power back-off, the BER values are higher than $10^{-4}$ for all bias points. The target BER of $2.5 \times 10^{-5}$ is achieved for 1.7 V bias points at 4 dB power back-off value. With further increase in the applied power back-off, the obtained BER values are improved towards the target BER. However, for the other bias points, the target BER can not be achieved even at 8 dB power back-off. When using the proposed predistorter, significant enhancements are noticed and the target BER is achieved for 1.7 V, 1.675 V, and 1.65 V bias points with only
2 dB power back-off. For the other bias points, 4 dB power back-off is sufficient to achieve the target BER.

In contrast to low order modulation schemes, signal distortion is shown to have a great impact on the achieved bit-error performance of higher modulation orders, namely 64-QAM. A slight increase of the EVM leads to a significant degradation in the BER performance. Therefore, and as expected, the 64-QAM modulation is very sensitive to signal distortion. As shown in Figure 2.14, even at 6 dB power back-off, the BER values are higher than $10^{-4}$ for all bias points. The target BER cannot be achieved with any bias point even at 8 dB power back-off. However, with the predistorter and 5 dB power back-off, the target BER is achieved for all bias points.
2.4 Results

Figure 2.13: (left) The BER for BPSK in the presence of LED non-linearity and AWGN channel without a predistorter. (right) The BER for BPSK in the presence of LED non-linearity and AWGN channel with a predistorter. (a) The corresponding EVM values. (b) The corresponding EVM values.
Figure 2.14: (a) The BER for 64-QAM in the presence of LED nonlinearity and AWGN channel without a predistorter. (b) The corresponding EVM values. 

(a) The BER for 64-QAM in the presence of LED nonlinearity and AWGN channel with a predistorter. (b) The corresponding EVM values.
2.5 Summary

In this chapter, a polynomial modeling of the LED nonlinear behavior was proposed. This model can be easily integrated within a simulation model to gain an insight on the generated nonlinear distortion and to determine bit/symbol error performance degradation. It was shown that the generated distortion is greatly influenced by the chosen DC operation point of the LED and the electrical OFDM signal power modulating the LED intensity. It was noticed that LED nonlinearity can significantly degrade the BER performance. Therefore, an optimum bias point must be identified for each LED to control the distortion levels, \textit{i.e.} to reduce magnitude distortion and to control signal clipping at the TOV and saturation voltage. In addition, a trade-off between signal coverage or SNR (determined by the applied power back-off on the electrical OFDM signal modulating the LED) and BER performance has to be made based on the target application. Finally, it was also demonstrated that this degradation can greatly be mitigated by using the proposed predistortion technique. The procedure to obtain the polynomial equation to describe the pre-distorter is simple and valid for any LED. The obtained results confirmed the BER enhancement after applying the LED pre-distorter.
2 Distortion Characterization and Linearization Technique
3 The S-shaped LED Model

3.1 Introduction

As explained previously in section 2.2.1, the bipolar time domain optical orthogonal frequency division multiplexing (O-OFDM) signal is clipped at the light emitting diode (LED) turn-on voltage (TOV) since the optical power cannot be negative. Also the O-OFDM signal is purposely clipped before modulating the LED for voltage levels above the saturation voltage to ensure that the LED chip does not overheat. Therefore, in a practical system, the O-OFDM signal should be conditioned prior to LED modulation. This can be achieved through an LED model equation which needs to describe the relation between the voltage across the LED and the current through the LED, i.e. V-I transfer function. The model equation needs also to level off at both extremes to represent the TOV at one end and the saturation voltage at the other end, i.e. lower peaks OFDM signal clipping and upper peaks OFDM signal clipping. The S-shaped LED model can also be used to evaluate the OFDM system performance (analytically or through Monte Carlo simulation) in the presence of LED nonlinearities. The behavior of the proposed model looks a bit like the letter ”S” and therefore the name ”S-shaped LED model” is given to the proposed model. This chapter aims to present the proposed model. In section 3.2, the procedure to obtain the model equation is introduced.

An S-shaped model describing a linear V-I transfer function is used to study the effect of the slope of this linear rise, the bias point, and signal clipping at the saturation voltage on the error performance of O-OFDM systems. At the saturation voltage, the O-OFDM signal can either be limited (hard-clipping) or smoothly clipped (soft-clipping) as explained in 3.2.2. Monte Carlo simulation results for the symbol error ratio (SER) versus the average power (over one OFDM symbol)
of the electrical O-OFDM signal modulating the LED are discussed in 3.2.4.

Furthermore, an analytical approach to evaluate the SER using the proposed S-shaped model fitted to a practical V-I transfer function of a commercially available white LED is illustrated. In section 3.3.3 a comparison with Monte Carlo simulation results is carried out to verify the accuracy of this approach.

Additionally, the SER using the predistorter technique introduced in section 2.2.3 is evaluated analytically and illustrated for the same white LED. Again a comparison with Monte Carlo simulation results is carried out to verify the accuracy of the analytical approach.

Finally, another predistortion technique based on a unity slope linear curve is proposed. This linear curve intersects with the LED transfer function at the bias point. For the same white LED, the SER using this predistortion technique is evaluated analytically and through Monte Carlo simulations. The obtained results are compared with the results obtained for the predistortion technique introduced in section 2.2.3.

3.2 The S-shaped LED Model

3.2.1 The solid state power amplifier (SSPA) Model

In radio frequency (RF) OFDM systems, the main source of nonlinearity is the power amplifier (PA) as shown in Figure 3.1. The PA operates near the saturation level in order to achieve the maximum power efficiency. In this operation region, undesirable nonlinear effects due to amplitude and phase distortions are introduced. Additionally, signal clipping at the saturation level is a critical source of distortion [90].

Several models have been developed to describe the nonlinear behavior of PAs. The most commonly used ones to relate the output voltage amplitudes to the input voltage amplitudes, i.e. amplitude conversion, are the limiter and the SSPA models. The limiter model is the simplest model which considers only the clipping effect at the saturation level. Therefore, the OFDM time domain samples $v_{in}$ are clipped as follows:

$$v_{out}(v_{in}) = \begin{cases} 
    v_{in} & \text{if } v_{in} < v_{\text{max}} \\
    v_{\text{max}} & \text{if } v_{in} \geq v_{\text{max}}
\end{cases}$$
3.2 The S-shaped LED Model

where, $v_{out}(v_{in})$ is the PA output voltage, $v_{in}$ is the PA input voltage, $v_{\text{max}}$ is the maximum output voltage (saturation voltage/level).

The SSPA model (usually called Rapps model) considers the transition between the two operation regions of the amplifier which are respectively linear and saturation. The Rapps model is given by [91],

$$v_{out}(v_{in}) = \frac{v_{in}}{\left(1 + \left(\frac{v_{in}}{v_{\text{max}}}\right)^{2k}\right)^{1/2k}} \quad (3.1)$$

where, $k$ is called the knee factor and controls the smoothness of the transition from the linear region to the saturation region. The knee factor, $k$, is a characteristic parameter of the amplifier. The transfer curve has always a unity slope linear rise followed by a knee and then leveling off as shown in Figure 3.1. Therefore, the characteristic of the amplifier will distort the signal amplitude due to the knee shape transition and affects the peak-to-average power ration (PAPR) of the OFDM signal through clipping.
Figure 3.2: The PA curve described by \((3.1)\) for \(k = 3\), the absolute value of the PA curve and the S-shaped curve described by \((3.3)\).

Figure 3.3: The effect of \(k\) on the proposed S-shaped LED model.
3.2 The S-shaped LED Model

3.2.2 The S-shaped Model with Linear Rise

Based on (3.1), the required LED model behavior can be described as follows:

\[ i_{\text{LED}}(v_{\text{LED}}) = \begin{cases} h(v_{\text{LED}}) & \text{if } v_{\text{LED}} \geq 0 \\ 0 & \text{if } v_{\text{LED}} < 0 \end{cases} \]

where, \( i_{\text{LED}}(v_{\text{LED}}) \) is current through the LED with leveling-off, \( v_{\text{LED}} \) is the voltage across the LED, and

\[ h(v_{\text{LED}}) = \frac{f(v_{\text{LED}})}{\left(1 + \left(\frac{f(v_{\text{LED}})}{i_{\text{max}}}\right)^{2k}\right)^{1/2k}} \]  

(3.2)

where, \( i_{\text{max}} \) is the maximum permissible AC/pulsed current through the LED, and \( f(v_{\text{LED}}) \) is the function describing the measured/data sheet I-V characteristics.

By observing the behavior of the PA model in (3.1) for negative input amplitudes, the curve saturates at \( -v_{\text{max}} \) as shown in Fig. 3.2. Thus, based on (3.2), the LED model equation with leveling-off can be described by the following equation,

\[ i_{\text{LED}}(v_{\text{LED}}) = \frac{1}{2} \left[ h(v_{\text{LED}}) + |h(v_{\text{LED}})| \right] \]  

(3.3)

where, \(|x|\) is the absolute value of \( x \).

Indeed the proposed equation in (3.3) levels off as shown in Fig. 3.3. In a Monte Carlo simulation, the equation clips the time-domain samples of the OFDM signal which is superimposed an a DC level according to the bias point under investigation. In (3.1), the knee factor \( k \) is set to fit the amplifier characteristic. However, the knee factor \( k \) in (3.2) is set either to limit (hard-clipping) or to smoothly clip (soft-clipping) the upper peaks of the OFDM signal. In a practical system, the model is used to condition the O-OFDM signal prior to LED modulation. For demonstration purposes, the LED curves for different values of \( k \) using \( f(v_{\text{LED}}) = v_{\text{LED}} \) at \( i_{\text{max}} = 0.5 \) A are depicted in Fig. 3.3.

3.2.3 The S-shaped Model Fitted to a Specific LED

The curves in Figure 3.3 have a linear rise with unity slope. However, to integrate the behavior of a specific LED, \( f(v_{\text{LED}}) \) should describe the V-I transfer function
For demonstration purpose, the transfer function of a white LED (OSRAM, Golden DRAGON, LA W57B, LY W57B) shown in Figure 3.4 is considered [92]. The currents at 1.52 V and 2 V are 1 mA and 240 mA, respectively. From the data sheet, the maximum permissible AC/pulsed forward current $i_{\text{max}}$ is 0.5 A. The TOV is considered to be at 1.52 V. The S-shaped curve considering the transfer curve of a specific LED is obtained through the following procedure (see Figure 3.5):

- The I-V characteristics is measured or obtained from the data sheet.
- The curve is shifted along the $x$-axis by the value of the TOV and the current which corresponds to the TOV is set to be constant for $v_{\text{LED}} \leq 0$.
- The polynomial equation to describe this curve is used to replace $f(v_{\text{LED}})$ in (3.2).

The S-shaped curves for this white LED for different values of $k$ are depicted on Figure 3.5(b) and on Figure 3.5(c).
3.2 The S-shaped LED Model

Figure 3.5: (a) semi-log measured V-I curve and leveling off at the TOV. (b) linear I-V curve with and without the TOV. (c) the effect of the knee factor $k$ on the fitted S-shaped curve. (d) zoomed: the effect of the knee factor $k$ on the fitted S-shaped curve.
3.2.4 Results

For Monte Carlo simulations, an O-OFDM system is implemented using 64 subcarriers with 31 subcarriers carrying data. Each subcarrier is modulated using uncoded 4-QAM (quadrature amplitude modulation) constellations. The average power of the constellations is normalized to unity. The average power of the electrical O-OFDM signal modulating the LED is varied from 50 mW up to 300 mW in steps of 10 mW. This power is calculated over one OFDM symbol. The O-OFDM signal varies $v_{\text{LED}}$ which is superimposed on a constant value based on the LED bias point under investigation. The O-OFDM demodulator implements the necessary blocks to estimate the transmitted data bits from $i_{\text{LED}}(v_{\text{LED}})$.

For $f(v_{\text{LED}}) = v_{\text{LED}}$ in (3.2), the effect of the clipping level, the slope of the rising curve, and the knee factor $k$ on the SER is obtained through Monte Carlo simulations. The obtained SER curves for $i_{\text{max}} = 0.5$ A, $i_{\text{max}} = 0.45$ A and $i_{\text{max}} = 0.4$ A, are shown in Figure 3.6. The bias point is set to 0.3 V and $k = 2$. As expected, the clipping of the upper peaks tremendously affects the SER. At 50 mW, the obtained SER is $1 \times 10^{-4}$, $7 \times 10^{-4}$, and $4 \times 10^{-3}$, for $i_{\text{max}} = 0.5$ A, $i_{\text{max}} = 0.45$ A, and $i_{\text{max}} = 0.4$ A, respectively. Above 100 mW, all three curves have SER value higher than $3 \times 10^{-3}$.

The SER curves for $k = 2$, $k = 25$, and four bias points, 0.1 V, 0.2 V, 0.3 V, and 0.4 V, are shown in Figure 3.7. The maximum permissible AC/pulsed LED current, $i_{\text{max}}$, is set to 0.5 A. It is noticed that for all bias points, the SER curves for $k = 2$ show better performance. The degradation of the SER performance of the curves for $k = 25$ compared to the curves for $k = 2$ increases with the increase of the bias point value. This confirms that the ”soft-clipping” is a valid and effective approach to control nonlinearity induced distortion specially at high bias points (where the clipping of the upper peaks tremendously affects the SER). It is also noticed that for $k = 2$, the 0.3 V bias point is the optimum among the four bias points under investigation.

The SER curves for different slopes of the rising curve, $f(v_{\text{LED}}) = 1/2v_{\text{LED}}$, $f(v_{\text{LED}}) = v_{\text{LED}}$, and $f(v_{\text{LED}}) = 2v_{\text{LED}}$, are shown in Figure 3.8. The bias point is set to 0.3 V and $i_{\text{max}} = 0.5$ A. It is noticed that at a lower slope, the SER performance is tremendously improved and $1 \times 10^{-6}$ SER is achieved at 50 mW. At the same power level, $1 \times 10^{-4}$ SER is obtained for the unity slope curve. For the higher slope, a SER floor around $8 \times 10^{-2}$ is observed.
3.2 The S-shaped LED Model

Figure 3.6: The effect of the clipping level \( i_{\text{max}} \) on the SER. The obtained SER curves for \( i_{\text{max}} = 0.5 \) A, \( i_{\text{max}} = 0.45 \) A, and \( i_{\text{max}} = 0.4 \) A. The bias point is set to 0.3 V and \( k = 2 \).

Figure 3.7: The effect of the knee factor \( k \) on the SER. The obtained SER curves for \( k = 2 \), \( k = 25 \), and four bias points, 0.1 V, 0.2 V, 0.3 V, and 0.4 V. The clipping level \( i_{\text{max}} \) is set to 0.5 A.
Figure 3.8: The effect of the slope of the rising curve on the SER. The obtained SER curves for different slopes of the rising curve, $f(v_{\text{LED}}) = 1/2v_{\text{LED}}$, $f(v_{\text{LED}}) = v_{\text{LED}}$, and $f(v_{\text{LED}}) = 2v_{\text{LED}}$. The bias point is set to 0.3 V and $i_{\text{max}} = 0.5$ A.

### 3.3 The Analytical Approach

A time domain O-OFDM signal is basically the sum of $N$ independent QAM modulated symbols, each symbol is transmitted over a different subcarrier. For large number of subcarriers, the OFDM signal can be accurately modeled by a Gaussian random process with a zero mean value and a variance $\sigma^2$. The probability density function (pdf) is given by:

$$P_X(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left(-\frac{x^2 - \mu}{2\sigma^2}\right) \quad (3.4)$$

where, $X$ is the continuous random variable (RV) which corresponds to the amplitude of the OFDM time domain samples, $x$ denotes a value taken by the RV $X$, and $\mu$ is the OFDM signal mean.

Simulation results (see Figure 3.9 and [93]) show that the OFDM signal constellations, in the presence of an LED (using the OSRAM Golden DRAGON white LED S-shape curve), experience amplitude and phase distortions in a random fashion. Hence, by modeling this nonlinearity induced noise as Gaussian noise,
3.3 The Analytical Approach

![Figure 3.9](image)

Figure 3.9: The constellations of a 4-QAM optical OFDM using the OSRAM white LED S-shaped curve with $k = 2$, biased at 2 V and the OFDM signal power is 200 mW. The DC subcarrier is set to zero.

A simple evaluation of error probability could be obtained as a function of the nonlinearity induced noise power $p_n$. The $p_n$ can be considered as an additional noise term that reduces the effective signal-to-noise ratio (SNR), $\rho$, to,

$$
\rho = \frac{\text{OFDM signal power}}{\text{Effective noise power}} = \frac{\sigma^2_o}{\sigma_{AWGN}^2 + p_n} 
$$

where, $\sigma_{AWGN}^2$ represents the shot noise and thermal noise at the optical receiver which can be modeled as AWGN [48], and $p_n$ is given by:

$$
p_n = p_{ad} + p_{uc} + p_{lc} 
$$

where, $p_{ad}$ is the noise component due to the amplitude distortion of the OFDM signal, $p_{uc}$ is the noise component due to the clipping of the upper peaks of the OFDM signal, and $p_{lc}$ is the noise component due to the clipping of the lower peaks of the OFDM signal.
3.3.1 Analytical Evaluation without a Predistorter

The noise component due to $p_{ad}$ is given by:

$$
p_{ad} = \int_{v_l}^{v_u} \left( g(v_{LED}) - i_{LED}(v_{LED}) \right)^2 P_{V_{LED}}(v_{LED}) dv_{LED} \tag{3.7}
$$

where, $i_{LED}(v_{LED})$ is the S-shaped curve of the current through the LED as a function of $v_{LED}$, $g(v_{LED})$ is the unity slope linear curve which intersects with $i_{LED}(v_{LED})$ at the chosen bias point $m$, the interval of the integral $[v_l \ v_u]$ is the interval of the amplitude distortion contribution to $p_n$ in (3.10), and $P_{V_{LED}}(v_{LED})$ is the pdf of the OFDM signal given by (3.4) with a mean value equal to the bias point ($\mu = m$).

For demonstration purpose, the same white LED from OSRAM is considered. The bias point is set to 2 V. This corresponds to $m = 0.48$ V after subtracting the 1.52 V TOV. Figure 3.10 shows the S-shaped curve for the considered LED using an arbitrary knee factor $k = 50$ and the interval $[0.17 \ 0.81 \text{ V}]$ of the integral in (3.7). Figure 3.11 shows the same S-shaped curve in addition to the linear curve $g(v_{LED})$ and the pdf $P_{V_{LED}}(v_{LED})$ over the interval $[0.17 \ 0.81 \text{ V}]$ with $\mu = m = 0.48$ V.

The noise component due to $p_{uc}$ is given by:

$$
p_{uc} = \int_{0}^{\infty} (v_{LED} - (v_u - m))^2 P_{V_{LED}}(v_{LED}) dv_{LED} \tag{3.8}
$$

where, $P_{V_{LED}}(v_{LED})$ is the pdf of the OFDM signal given by (3.4) with zero mean ($\mu = 0$). The time domain OFDM signal values above $v_u - m$ contribute to the noise component due to the clipping of the upper peaks of the OFDM signal, $p_{uc}$. The upper subplot of Figure 3.12 shows the value of $v_u - m$ in (3.8) which is equal to 0.33 V and the pdf $P_{V_{LED}}(v_{LED})$ over the interval $[0.33 \ \infty \text{ V}]$ in (3.8) with $\mu = 0 \text{ V}$.

The noise component due to $p_{lc}$ is given by:

$$
p_{lc} = \int_{\infty}^{\infty} (v_{LED} - (m - v_l))^2 P_{V_{LED}}(v_{LED}) dv_{LED} \tag{3.9}
$$
3.3 The Analytical Approach

Figure 3.10: The S-shaped curve for the considered LED using an arbitrary knee factor $k = 50$ and the interval $[0.17 \, \text{V} \, 0.81 \, \text{V}]$ of the integral in (3.7).

Figure 3.11: The S-shaped curve for the considered LED using an arbitrary knee factor $k = 50$ in addition to the linear curve $g(v_{\text{LED}})$ and the pdf $P_{v_{\text{LED}}}(v_{\text{LED}})$ over the interval $[0.17 \, \text{V} \, 0.81 \, \text{V}]$. 
Figure 3.12: [upper subplot] the value of \( v_u - m \) in (3.8) which is equal to 0.33 V and the pdf \( P_{V_{LED}}(v_{LED}) \) over the interval \([0.33 \text{ V} \infty] \) in (3.8). [lower subplot] the value of \( m - v_l \) in (3.9) which is equal to 0.31 V and the pdf \( P_{V_{LED}}(v_{LED}) \) over the interval \([0.31 \text{ V} \infty] \) in (3.9).

where, \( P_{V_{LED}}(v_{LED}) \) is the pdf of the OFDM signal given by (3.4) with zero mean (\( \mu = 0 \)). The time domain OFDM signal values above \( m - v_l \) contribute to the noise component due to the clipping of the lower peaks of the OFDM signal, \( p_{lc} \). The lower subplot of Figure 3.12 shows the value of \( m - v_l \) in (3.9) which is equal to 0.31 V and the pdf \( P_{V_{LED}}(v_{LED}) \) over the interval \([0.31 \text{ V} \infty] \) in (3.9) with \( \mu = 0 \) V.

### 3.3.2 Analytical Evaluation with a Predistorter

Using a predistorter to linearize the LED transfer function, the nonlinear induced noise power \( p_n \) in (3.5) is given by:

\[
p_n = p_{uc} + p_{lc}
\]  

(3.10)

The noise component due to \( p_{uc} \) is given by:

\[
p_{uc} = \int_{v_{up}}^{\infty} (v_{LED} - v_{up})^2 P_{V_{LED}}(v_{LED})dv_{LED}
\]  

(3.11)
3.3 The Analytical Approach

where, $v_{up}$ is the voltage corresponding to the bias point voltage subtracted from the voltage at the intersection of the linearized LED curve with the maximum permissible AC/pulsed current level, and $P_{V_{LED}}(v_{LED})$ is the pdf of the OFDM signal given by (3.4) with zero mean ($\mu = 0$).

The noise component due to $p_{lc}$ is given by:

$$p_{lc} = \int_{v_{lp}}^{\infty} (v_{LED} - v_{lp})^2 P_{V_{LED}}(v_{LED}) dv_{LED} \quad (3.12)$$

where, $v_{lp}$ is the voltage corresponding to the intersection of the linearized LED curve with the $x$-axis (i.e. no current conduction through the LED) subtracted from the bias point voltage, and $P_{V_{LED}}(v_{LED})$ is the pdf of the OFDM signal given by (3.4) with zero mean ($\mu = 0$).

For the same white LED from OSRAM, Figure 3.13 shows the LED curve after and before the linearization. If the bias point is set to 2 V, the $v_{up}$ is equal to 0.3 V and the $v_{lp}$ is equal to 0.3 V.
3.3.3 Results

The same OSRAM white LED is used to demonstrate the validity of the analytical approaches introduced in section 3.3.1 and in section 3.3.2 to estimate $p_n$. The SER is evaluated analytically (as a function of the $\rho$) and compared with Monte Carlo simulation results. The equation used to calculate the analytical SER for the digital modulation technique under investigation, the 4-QAM, is given by [94],

$$\text{SER}_{4\text{-QAM}} = Q\left(\sqrt{\frac{1}{2} \times 10^{(\rho_{dB}/10)}}\right)$$

(3.13)

where, $Q(x)$ is the Q-function defined as,

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} \exp\left(-\frac{u^2}{2}\right) \, du$$

(3.14)

For Monte Carlo simulations, an O-OFDM system is implemented using 64 subcarriers with 31 subcarriers carrying data. Each subcarrier is modulated using uncoded 4-QAM constellations. The average power of the constellations is normalized to unity. Shot noise and thermal noise at the receiver are modeled as AWGN. The OFDM signal power modulating the LED is varied to explore the influence of the OFDM signal amplitude variations on the nonlinearity induced noise. Additionally, different bias points, are considered to investigate the influence of the bias point on the generated distortion.

The effective noise power and $p_n$ are evaluated numerically and Figure 3.14 shows the obtained results for different AWGN powers and OFDM signal powers ranging from 10 mW up to 300 mW. The bias point is set to 2 V. It is observed that for 0.01 mW and 0.1 mW AWGN powers, the contribution of the LED induced distortion noise power to the effective noise power is notable even at low OFDM signal power (e.g. 10 mW). Considering the 10 mW AWGN, the effective noise power starts increasing around 50 mW OFDM signal power.

The corresponding analytical SER results are shown in Figure 3.15. The SER calculated in the presence of the LED without additional AWGN is plotted as a reference. Starting 1 mW OFDM signal power, the SER is improved with the increase of the OFDM signal power, i.e. the signal power is getting larger than
3.3 The Analytical Approach

The effective noise power. With further increase of the OFDM signal power, the effective SNR, \( \rho \), starts to decrease at certain signal powers based on the AWGN power and consequently the SER starts to increase, \( i.e. \) the effective noise power is getting larger than the OFDM signal power. It is also noticed that with lower AWGN powers, the OFDM signal powers required to get minimum SER values decrease.

The SER results for 0.1 mW AWGN power by means of the analytical approach and Monte Carlo simulations are shown in Figure 3.16. Three bias points, 1.9 V, 2 V, and 2.1 V, are considered. The analytical curves clearly follow the behavior of the simulation curves. A close estimation for the simulated curve is noticed, \( e.g. \) a different of 0.5 dB is calculated for \( 1 \times 10^{-3} \) SER. The 2 V bias point offers the best SER performance. At 50 mW, a \( 9 \times 10^{-6} \) SER is obtained, however the SER increases to \( 1 \times 10^{-3} \) at 100 mW. The 2.1 V bias point has better performance compared to the 1.9 V bias point. The corresponding nonlinearity induced noise power calculated analytically is shown in Figure 3.17. At 50 mW, the highest noise power component is 3 mW which is the noise power due to lower clipping of the OFDM signal peaks of the 1.9 V bias point. The upper and lower clipping noise power curves at the 2 V bias point are almost overlapping to indicate the
symmetry of the LED curve around this bias point. For the 2 V bias point, the maximum contribution of the clipping noise power at 300 mW is around 35 mW. For 300 mW OFDM signal power, the upper clipping noise power at 2.1 V and the lower clipping noise power at 1.9 V is around 50 mW. For the same signal power, the upper clipping noise power at 1.9 V and the lower clipping noise power at 2.1 V is around 25 mW. The variation of the amplitude distortion noise power is very small and the values range from 0.45 mW up to 0.7 mW. For OFDM signal power above 100 mW, the behavior of the curves (the noise power decreases with further increase of the OFDM signal power) confirms the dominant contribution of the clipping noise powers.

To highlight the dependance of the SER on $p_{ad}$, the left subplot of Figure 3.18 shows the analytical SER with/without $p_{ad}$. It is noticed that the deviation of the SER curves without $p_{ad}$ from the SER curves with $p_{ad}$ increases for lower OFDM signal power. This behavior is expected since the contribution of $p_{ad}$ is major for low OFDM signal power. For high OFDM signal power, $p_{dc}$ and $p_{ac}$ dominate $p_n$. This deviation shows also bias point dependance. For the 2 V bias point, the deviation is larger compared to the other bias points. The right subplot of Figure 3.18 shows the analytical upper and lower clipping noise powers. The 2 V
Figure 3.16: Simulation and analytical SER for the OSRAM Golden DRAGON white LED and 0.1 mW AWGN power.

Figure 3.17: Analytical noise power for the OSRAM Golden DRAGON white LED.
Figure 3.18: [left] Analytical SER without amplitude distortion [right] Upper and lower clipping noise powers.

Figure 3.19: Analytical SER for different amplitude distortion intervals.
Figure 3.20: Simulation and analytical SER for the OSRAM Golden DRAGON white LED using a predistorter and 0.1 mW AWGN power.

bias point offer less clipping noise, therefore it is more influenced by $p_{ad}$, specially for low OFDM signal power.

To highlight the dependence of the SER on the interval in (3.7) to estimate $p_{ad}$, Figure 3.19 shows the obtained analytical SER for the intervals $[0, 0.81]$ and $[0.17, 0.81]$. It is clearly shown that the behavior of the SER for the $[0, 0.81]$ interval doesn’t follow the behavior for the $[0.17, 0.81]$ interval. This confirms that the OFDM time domain samples below 0.17 V contribute to $p_{lc}$ rather than to $p_{ad}$.

For a system using a predistorter, the SER simulation results and results through the analytical approach for 0.1 mW AWGN power are shown in Figure 3.20. The same bias points, 1.9 V, 2 V, and 2.1 V, are considered. Again the analytical curves clearly follow the behavior of the simulation curves. The SER performance of the 1.9 V and the 2 V is improved. For the 2 V bias point, the obtained SER is $4 \times 10^{-6}$ at 50 mW, however the SER increases to $1 \times 10^{-3}$ at 100 mW. The corresponding nonlinearity induced noise calculated analytically is shown in Figure 3.21. As expected, the lower clipping noise and the upper clipping noise curves are overlapping. For the 2 V bias point, the maximum contribution of the clipping noise at 300 mW is around 40 mW.
A predistortion using a unity slope linear curve passing through the three bias points under investigations is also considered and compared with the linearized curve demonstrated in \ref{sec:linearization}. Figure \ref{fig:linearization} shows the linear curves with a unity slope for the 2 V, 1.9 V, and 2.1 V bias points. The SER simulation results and results through the analytical approach for 0.1 mW AWGN power are shown in Figure \ref{fig:analytical}. The analytical curves clearly follow the behavior of the simulation curves. For analytical calculations, the equations (3.8) and (3.9) introduced in section \ref{sec:linearization} are used. Again, the 2 V bias point offers the best SER performance. At 50 mW, the obtained SER is $3 \times 10^{-4}$, however the SER increases to $5 \times 10^{-2}$ at 100 mW. The 2.1 V bias point has better performance compared to the 1.9 V bias point. The corresponding nonlinearity induced noise calculated analytically is shown in Figure \ref{fig:nonlinearity}.

Figure \ref{fig:nonlinearity} shows the nonlinearity induced noise of the two linearization techniques under investigations for the OFDM signal power range starting from 50 mW up to 100 mW. The obtained values confirms the advantage of the lower slope linearized-LED curve compared to the unity slope linearization curves.
3.3 The Analytical Approach

Figure 3.22: The difference between using the predistorter and a unity slope linear curve through the bias point.

Figure 3.23: Simulation and analytical SER for the OSRAM Golden DRAGON white LED with unity slope linearization and 0.1 mW AWGN power.
Figure 3.24: Analytical noise power for the OSRAM Golden DRAGON white LED with unity slope linearization.

Figure 3.25: [right] Clipping noise power using a predistorter [left] Clipping noise power using unity slope linearization.
3.4 Summary

In this chapter, the S-shaped LED model was proposed. The model describes the LED transfer function and levels off for voltages (across the LED) below the TOV and above the saturation voltage. The behavior looks a bit like the letter “S” and therefore the name “S-shaped LED model” was given to the model. The model is suitable for analyzing and/or simulating optical communication systems based on intensity modulation using O-OFDM signal variations. In a practical system, the model can also be used to condition the O-OFDM prior to LED modulation. Signal conditioning includes clipping of the lower peaks and/or the upper peaks.

The model with a unity slope linear rise was used to study the effects of several design parameters such as the bias point, clipping level, and the OFDM signal conditioning (as a function of the knee factor, $k$, to realize hard/soft-clipping of the upper peaks) on SER performance. It was shown that for all bias points, the obtained SER curves using a small value for the knee factor, i.e. $k = 2$, to demonstrate soft-clipping show better performance compared to hard-clipping, i.e., $k = 50$. This confirms that soft-clipping is a valid and effective approach to control nonlinearity induced distortion and to enhanced power efficiency, specially at high bias points.

Additionally, the S-shaped model fitted to a specific LED transfer function rather than having a unity slope linear rise was used to evaluate the SER performance through an analytical approach and Monte Carlo simulations. Furthermore, an analytical approach to evaluate O-OFDM systems with predistortion was proposed and a comparison with Monte Carlo simulation results was carried out. The analytical curves clearly follow the behavior of the simulation curves and a close estimation for the simulated curves was noticed. The obtained SER results confirmed that the pre-distorter offers better performance than a unity slope linear curve which results in a smaller dynamic range.
3 The S-shaped LED Model
4 Hardware Demonstrators

This chapter presents two hardware prototypes to demonstrate online short-range optical transmission. Visible light transmission is considered. The building blocks of the optical orthogonal frequency division multiplexing (O-OFDM) system are highlighted in section 4.1. The target of the conducted study was not to showcase high data rates; rather to study via a simple proof-of-concept hardware demonstrator achievable rates for phase-incoherent O-OFDM and to investigate performance for different electrical signal-to-noise ratios (SNRs). In this context, the influence of the transfer function of the considered light emitting diode (LED) on the bit-error-ratio (BER) performance will be experimentally confirmed. The first visible light communication (VLC) demonstrator utilizes a single off-the-shelf LED transmitter and is presented in section 4.2. The results show that for a directed-LOS (line-of-sight) scenario and without any channel coding, a BER of about $1 \times 10^{-4}$ for quadrature phase shift keying (QPSK) constellations at 0.5 m separation distance can be obtained. Also a reasonable BER of $2 \times 10^{-3}$ for 64-QAM (quadrature amplitude modulation) at 0.2 m is achieved. A 1/2 channel coding rate (convolutional encoder) is also considered and the performance is tremendously improved.

The second VLC demonstrator utilizes a 9-LED array of the same LED type used in the first demonstrator and is covered in section 4.3. Using this prototype, experimental results on the optical path-loss are obtained and compared with theoretical calculations. Also the influence of the LED beam angle on the horizontal coverage is highlighted. Moreover, the influence of the electrical SNR at the receiver, the constellation order, and the channel coding rate on the BER are investigated. Both VLC demonstrators are based on digital signal processing (DSP) development boards (DSP development board TMS320C6713 from Texas Instruments) [95].
Figure 4.1: A short-range simplex optical link demonstrator focusing on indoor broadcasting applications. The demonstrator consists of DSP evaluation boards, transmitter/receiver computers, and analog optical transmitter/receiver modules.

The hardware prototypes include two parts (see Figure 4.1). The digital part, which is the DSP development board to generate/decode the O-OFDM signal and to interface with the transmitter and receiver computers. The analog part, which includes the LED transmitter and the driver electronics. The analog part at the receiver includes a photodiode (PD), a transimpedance amplifier (TIA), DC blocking stage, and a preamplifier stage.

### 4.1 OFDM Building Blocks

The building blocks of the physical layer are depicted in Figure 4.2. The system uses a forward error correction (FEC) coding algorithm for data protection, namely a convolutional encoder. In addition, burst error protection is realized through time and frequency interleaving algorithms. One of several modulators (phase-shift keying (PSK) or multi-level quadrature amplitude modulation (M-QAM)), modulates the encoded bit stream into symbols. The generated serial stream of symbols at the modulator output is mapped into parallel streams and each stream is transmitted on a separate subcarrier. The inverse fast Fourier transform (IFFT) operation modulates and multiplexes the subcarriers. After generating the OFDM symbol, a cyclic prefix (CP) is added as a guard interval to avoid multipath induced inter-symbol interference (ISI) and to convert the linear convolution of the channel with the OFDM signal to a circular convolution. As a result, simple frequency
domain equalizer can be used.

At the receiver, time synchronization and symbol equalization are realized using a single OFDM symbol as block-type pilots [84]. The complex conjugate requirement is fulfilled while generating the pilots symbol. The OFDM frame consists of one OFDM symbol forming the block-type pilots and 20 OFDM symbols with data subcarriers. After the CP removal, the OFDM signal is converted back to the frequency domain by applying the fast Fourier transform (FFT) operation. Using the pilots symbol, the channel is estimated and frequency domain equalization is realized using a conventional OFDM zero-forcing (ZF) equalizer. The estimated bit stream is deinterleaved and then decoded by a hard decision Viterbi algorithm.

The parameters of the O-OFDM model is listed in Table 4.1.

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<thead>
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<th>Number of IFFT/FFT points</th>
<th>64</th>
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<tr>
<td>Number of data subcarriers</td>
<td>31</td>
</tr>
<tr>
<td>Number CP samples</td>
<td>16</td>
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<tr>
<td>Pilot symbols per frame</td>
<td>1</td>
</tr>
<tr>
<td>Data symbols per frame</td>
<td>20</td>
</tr>
</tbody>
</table>

### 4.2 VLC with Single LED

A single off-the-shelf white LED [96] and a single PD (SLD-70BG2) [97] are utilized in the analog front ends shown in Figure 4.3. The parameters of these analog front ends are listed in Table 4.2. This prototype allows investigating the influence of the electrical SNR at the receiver, the constellation order, and the channel coding rate on the BER performance. A directed-LOS (line-of-sight) scenario is considered.

<table>
<thead>
<tr>
<th>PD respositivity</th>
<th>0.53 [A/W]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Luminance intensity</td>
<td>11 [cd]</td>
</tr>
<tr>
<td>LED beam angle</td>
<td>20°</td>
</tr>
<tr>
<td>PD active area</td>
<td>9.8 [mm²]</td>
</tr>
<tr>
<td>PD FOV</td>
<td>60°</td>
</tr>
</tbody>
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Figure 4.2: The building blocks of the physical layer of the optical OFDM system, and IFFT bin assignment. The DC subcarrier should be avoided for data transmission.
4.2 VLC with Single LED

Figure 4.3: VLC using a single LED transmitter and a single PD receiver.

As the prototype is intended for indoor data transmission applications, the measurements are taken under ambient diffused sunlight inside a room. Figure 4.3 shows the SNR measured at the receiver up to 1 m. About 40 dB, 17 dB, and 5 dB electrical SNRs are measured at 15 cm, 0.5 m, and 1 m, respectively.

Figure 4.4 shows the BER curves versus the distance between the analog front ends. Measurements are taken for a distance up to 1 m. The BER for uncoded QPSK, uncoded 16-QAM transmission, as well as uncoded 64-QAM transmission are plotted. For the coded case (1/2 channel coding rate), QPSK and 16-QAM results are shown. The BER below $1 \times 10^{-7}$ is not recorded and that is why some graphs consist of fewer measurement points.

The results show that already without any channel coding, a BER of about $1 \times 10^{-4}$ at a separation distance of 0.5m can be achieved with QPSK modulation. For the same separation distance, channel coding improves the performance and BER of less than $1 \times 10^{-7}$ is noticed. It is also interesting to note, that the BER performance for coded QPSK transmission suddenly increases to $1 \times 10^{-1}$ at a distance of about 1m, while the BER performance is still about $1 \times 10^{-5}$ at 90cm separation. This distance can hence be considered as the maximum possible distance of the single LED transmitter. A further interesting observation is that at a distance of about 20 cm (20% of the maximum distance), uncoded 64-QAM still results in a reasonable BER performance ($2 \times 10^{-3}$).
Figure 4.4: SNR measurements at the receiver.

Figure 4.5: BER vs. the separation distance between the analog transmitter and receiver.
4.3 VLC with LED Array

In this prototype, the same LED and PD considered in section 4.2 are utilized, however a 9-LED array is realized as shown in Figure 4.6. A desktop lamp transmitter consists of 9 LEDs which are 1 cm separated and the optical receiver circuit employs a single PD. Experimental results on the optical path-loss are obtained and compared with theoretical calculations. The influence of the LED beam angle on the horizontal coverage is highlighted. Additionally, the influence of the electrical SNR at the receiver, the constellation order, and the channel coding rate on the BER performance are investigated.

4.3.1 Optical Channel

The bandwidth of the optical channel in a LOS configuration is reported higher than 88MHz [43]. Therefore, the optical pass-loss is the most important quantity to characterize the channel and relates the transmitted and received optical powers via [21],

\[ P_r = H(0)P_t \]  \hspace{1cm} (4.1)

where, \( P_t \) is the transmitted optical power, \( P_r \) is the received optical power, and \( H(0) \) is the optical path-loss. This approximation is particularly accurate in
directed-LOS links. Considering the LOS link geometry shown in Figure 4.7, the LOS channel path-loss is defined as [21],

\[ H(0)_{\text{LOS}} = \frac{A_{\text{PD}}}{d^2} R_0(\phi) T_s(\psi) g(\psi) \cos(\psi) \]  

(4.2)

where, \( A_{\text{PD}} \) is the PD active area, \( d \) is the distance between the transmitter and the receiver, \( \phi \) is the angle with respect to the transmitter, \( \psi \) is the angle with respect to the receiver, \( T_s(\psi) \) is the filter gain, \( g(\psi) \) is the concentrator gain, and \( R_0(\phi) \) is the transmitter radiant intensity given by [21],

\[ R_0(\phi) = \left[ \frac{n + 1}{2\pi} \right] \cos^n \phi \quad \text{(W/sr)} \]  

(4.3)

\[ n = \frac{\ln 2}{\ln(\cos \alpha)} \]  

(4.4)

where, \( \alpha \) is the transmitter beam angle.

### 4.3.2 Transmitted Optical Power

Most data sheets of white LEDs provide only the photometric power, namely luminous flux \( F \) in lumens or the luminous intensity \( I_v \) measured in candela, which are useful metrics for illumination design. However, the radiometric power in watts is more relevant parameter for wireless transmission. Therefore, measurements were conducted to determine the transmitted optical power \( P_t \) in watts for the considered LED. The LED operates with 20 mA bias current and dissipates 62 mW
of electrical power.

An optical power meter is used to measure the spectral power distribution $P(\lambda)$ in steps of 10 nm (starting 400 nm to 750 nm) [98]. The optical power meter is limited to 400 nm minimum wavelength and its photo detector has an active area of 1 cm$^2$. The obtained values are used to determine $P_t$ and $F$ using the following equations [21],

$$P_t = \sum_{400\text{nm}}^{750\text{nm}} P(\lambda)\Delta\lambda \quad (\text{W})$$  \hspace{1cm} (4.5)

$$F = \gamma \sum_{400\text{nm}}^{750\text{nm}} V(\lambda)P(\lambda)\Delta\lambda \quad (\text{lm})$$  \hspace{1cm} (4.6)

where, $\Delta\lambda$ is 10 nm, $V(\lambda)$ denotes the CIE 1931 (international commission on illumination) eye sensitivity function in the photonic vision regime, and $\gamma = 683 \text{ lm/W}$ is the peak luminous efficacy based upon the sensitivity of the eye at 555 nm [99].

A conversion factor $\xi$ relating the photometric power to the radiometric power can be obtained by using (4.5) and (4.6),

$$\xi = \frac{P_t}{F} \quad (\text{W/Im})$$  \hspace{1cm} (4.7)

The values for $I_v$ and the beam angle $\alpha$ can be obtained from the data sheet and used to calculate $F$ as follows [100],

$$\Omega = (1 - \cos(0.5\alpha))2\pi \quad (\text{sr})$$  \hspace{1cm} (4.8)

$$F = I_v\Omega \quad (\text{lm})$$  \hspace{1cm} (4.9)

where, $\Omega$ is the LED solid angle in steradian.

A conversion factor $\xi = 5.4 \text{ mW/Im}$ is calculated by using (4.5), (4.6), and (4.7). From the data sheet, $\alpha$ and $I_v$ are 20° and 11 cd, respectively. A 1.05 lm was calculated by using (4.8) and (4.9), and the corresponding $P_t$ is 5.7 mW. This corresponds to 51 mW total transmitted power from the 9-LED array.
4.3.3 Received Optical Power and Path-loss

Two coverage scenarios, namely vertical and horizontal scenarios, is considered, as shown in Figure 4.7. Experimental measurements is conducted to explore the received optical power and the path-loss for these coverage scenarios. The transmitter is directed downwards and emitting towards the floor. The receiver is directed upwards towards the ceiling. The separation distance between the transmitter and the receiver is denoted by \( d \). In a vertical coverage scenario, the receiver is moving vertically away from the transmitter (\( d = 50 \text{ cm} \rightarrow 225 \text{ cm}, \text{ in } 25 \text{ cm steps} \)). In a horizontal coverage scenario, the vertical distance between the transmitter and the receiver is fixed (\( b = 1 \text{ m} \)) and the receiver is moving horizontally (\( a = 0 \rightarrow 50 \text{ cm}, \text{ in } 10 \text{ cm steps} \)).

A PD with 9.8 mm\(^2\) active area is considered. The optical filter and the concentrator gain are set to 1. The average received optical power is obtained through measuring \( P(\lambda) \) and substituting in (4.5). The obtained values are scaled to correspond to the optical power on the 9.8 mm\(^2\) active area.

The calculated and the measured received optical power for the vertical coverage scenario are depicted in Figure 4.8. Theoretical and measured results match closely. Along the vertical separation, a path-loss variation of more than 12 dB is observed. The received average optical power at 0.5 m is around -24 dBm and reduces to -36 dBm (12 dB loss) at 2 m.

Theoretical and measured results for the horizontal coverage scenario are shown in Figure 4.9. A -30 dBm optical power is measured when the receiver is directly located under the transmitter. A 0.5 m horizontal displacement yields -37 dBm (7 dB loss) optical power. It can be seen that there is only a relative minor attenuation up to 20 cm away from the initial position (1 dB loss). A more pronounced attenuation is observed between 20 cm to 40 cm horizontal displacement (4 dB loss). This can be attributed to the field of view (FOV) mismatch between the transmitter and the receiver at \( d > 20 \text{ cm} \). With 20° LED beam angle, the horizontal coverage is calculated to be 17.6 cm as indicated in Figure 4.7. A horizontal displacement further above 17.6 cm from the initial position (at \( a = 0 \text{ cm} \)) places the PD out of the illumination coverage. This explains the aforementioned path-loss behavior. Therefore, a proper setting of the LED beam angle, the number of LEDs forming the array, the array geometry, and the FOV of the photodiode is essential to optimize the coverage [101]. This also highlights that these effects
4.3 VLC with LED Array

4.3.4 Bit-error Performance

All measurements are taken inside a room and ambient daylight through windows is considered. The electrical SNR is measured over one OFDM symbol. At least 10 data blocks of $10^6$ bits each are sent to measure the system bit-error performance. The BER below $10^{-6}$ is not recorded and that is why some graphs consist of fewer measurement points.

The BER performance versus distance for the vertical coverage scenario is presented in Figure 4.10. The additional $y$-axis at the right shows the measured electrical SNR at different receiver positions. For broadcasting applications using a reading lamp, it is practically valid to consider the target separation distance between the transmitter and the receiver to be around 1m, and $10^{-5}$ as the target BER for video broadcasting. For example, MPEG-4 video transmission has slight visible degradation at $10^{-5}$ BER [89]. Low order modulation schemes (BPSK and QPSK) can achieve these requirements even without any channel coding. For high order modulation schemes, namely 16-QAM and 64-QAM, the 16-QAM with 3/4
Figure 4.9: Measured and calculated SNR at the receiver for horizontal coverage scenario.

channel coding rate can achieve the required targets and BER less than $10^{-6}$ up to 2 m (18 dB SNR). However, the 64-QAM with 2/3 channel coding rate can only achieve the $10^{-5}$ BER target with higher SNR value (33 dB SNR). Finally, the 64-QAM 1/2 channel coding rate fulfils the requirements and achieves BER less than $10^{-6}$ up to 1.75 m (18 dB SNR). The uncoded modulation curves are included as references. The effect of the LED nonlinearity on the performance is clearly noticed. For example, 64-QAM with 3/4 channel coding rate achieves $2 \times 10^{-5}$ for 22 dB SNR in AWGN environment (see Figure 2.12). However, the measured BER for 64-QAM with 3/4 channel coding rate in the presence of the LED is around $2 \times 10^{-2}$ for 22 dB SNR.

The BER performance of QPSK modulation versus distance for the horizontal coverage scenario is depicted in Figure 4.11. A horizontal displacement of 0.5 m is considered to examine the cell coverage edge of an LED reading lamp. At 20 cm (25 dB SNR), BER less than $10^{-6}$ is achieved even without introducing any channel coding. However, at 30 cm the SNR value drops drastically to reach 9 dB and even with 2/3 coding rate, the maximum BER that can be achieved is $3 \times 10^{-4}$. To maintain the required BER performance one has to resort to 1/2 rate coded QPSK modulation.
Figure 4.10: Measured BER and SNR of vertical coverage scenario.

Figure 4.11: Measured BER and SNR of horizontal coverage scenario.
4.3.5 Illumination Requirements

The illuminance is measured and compared with the minimum required values according to lighting standards [77, 102]. The minimum illuminance required for different work spaces ranges from several hundred to thousand lux. Therefore, for VLC, a high SNR is obtainable, which is an indirect consequence of the illumination requirements. In order to determine the illuminance achieved using the square array of 9 LEDs lamp, measurements are conducted using a lux meter [103]. The obtained illuminance for the vertical coverage scenario is shown in Figure 4.12. At the target distance of 1m, only 20 lx are measured. From the obtained values, it is expected that with the appropriate number of LEDs to achieve sufficient illuminance, high SNR values can be achieved.

4.4 Summary

Although a directed-LOS link configuration was considered, the obtained BER vs. SNR is generally valid for OFDM based VLC systems because the illumination
requirements results in a dominant LOS component in most positions in the room. Even when multipath components exist, OFDM inherently combat multipath induced ISI with a proper CP length.

The DSP board used in VLC demonstrators has an onboard stereo codec with 96 kHz maximum sampling frequency. Therefore, the OFDM signal bandwidth is limited to 45 kHz. Consequently, the current bandwidth is sufficient for messaging or information services, several audio channels, low quality video streaming applications.

Preliminary measurements using the 9-LEDs desktop lamp showed promising results with illuminance of about 5 times below that required for work spaces. Therefore, it is expected that with the appropriate number of LEDs to achieve sufficient illuminance, the coverage area can be extended without any ISI and high SNR values can be achieved. It also has been found that signal coverage and CCI in a cellular network can be controlled via proper setting of the LED beam angle.
5 Conclusions and Future Work

5.1 Conclusions

In this thesis, the impact of the nonlinear characteristics of light emitting diodes (LEDs) on the bit/symbol error performance of optical communication systems based on intensity modulation (IM) using orthogonal frequency division multiplexing (OFDM) signals was studied. Also, a visible light communication (VLC) hardware demonstrators were realized. The main findings and conclusions can be summarized as follows:

- In this thesis, a digital pre-distorter is proposed to mitigate performance degradation due to LED nonlinearity. The performance of the compensated system can be greatly enhanced. For instance, using 64-QAM (quadrature amplitude modulation) constellations, it is shown in Section 2.4 that with a digital pre-distorter, only 5 dB power back-off is required to achieve the target bit-error ratio (BER) of $2.5 \times 10^{-5}$ for all bias points under investigation. For uncompensated system, power back-off of 8 dB was not enough to achieved the target BER with any bias point.

- For the first time, a model to describe the LED behavior which incorporates amplitude distortion and levels off for voltages that are below the LED turn-on voltage (TOV) and above the saturation voltage which corresponds to the maximum AC/pulsed current recommended by the LED manufacturer is proposed. The model allows searching for an optimum bias point and optimal OFDM signal power to modulate the LED intensity, in order to control the LED nonlinearity induced distortion.
• Analytical methodologies to evaluate error probability of uncompensated optical OFDM (O-OFDM) systems as well as for compensated O-OFDM systems are proposed. A comparison with Monte Carlo simulation results were carried out. The simulation curves and the analytical curves match closely in terms of symbol error ratio (SER).

• It is also noticed that the power efficiency can be enhanced through a smooth clipping of the upper peaks of the OFDM signal, i.e. soft-clipping, at the maximum permissible AC/pulsed current level rather than limiting the upper peaks at this level (hard-clipping). Consequently, a wider signal coverage can be achieved.

• It is shown that the introduced analytical approach and Monte Carlo simulations using the proposed LED model are practical ways to determine a suitable LED and to optimize the bias point, the O-OFDM signal power, and/or the knee factor $k$ for target application requirements.

• The obtained results confirmed that selecting an LED with high AC/pulsed current level, i.e. large dynamic range, enhances the performance. Moreover, SER performance is improved by considering an LED with low I-V slope characteristics. Therefore, a system design should consider maximizing the LED dynamic range and selecting an LED with a lower I-V slope characteristics to enhance the performance.

### 5.2 Limitations and Future work

The proposed model to describe the nonlinear behavior of an LED was based on a measured static transfer function of the LED. The modulation was achieved by varying the current with the OFDM signal. The author recognizes that the model can be further enhanced through the consideration of the LED frequency response. However, taking the LED frequency response into account is not straight forward, since the LED rate equation for the change in the carrier concentration should be considered. An interesting research issue for further study is to incorporate both the nonlinear behavior and the frequency response of the LED.

It was shown in this thesis that the bias point and the OFDM signal power modulating the LED have a significant impact on the induced nonlinearity distor-
5.2 Limitations and Future work

The nonlinearity distortion results in this research were obtained using mathematical models and Monte Carlo simulations. Future work should consider investigating LED nonlinearity through experimental measurements. However, this requires an analog transmitter front end capable of adjusting the LED bias point, the OFDM signal amplitude modulating the LED, and limiting the OFDM signal peaks at desired levels. For this reason, a modified transmitter unit is to be developed.

Two forms of optical OFDM, namely, DC biased optical OFDM (DCO-OFDM) and asymmetrically clipped optical OFDM (ACO-OFDM) were briefly covered in 13.2. However, in this thesis, the DCO-OFDM is considered to study LED nonlinearity and to demonstrate VLC. It would be interesting to consider for future work the evaluation of ACO-OFDM systems in the presence of LED nonlinearity. It is also important to compare the results with that obtained for DCO-OFDM systems.

Manufacturing analog front ends covering wider signal bandwidth, i.e. above 20 MHz, and providing higher optical-to-electrical and/or electrical-to-optical conversion efficiencies is subject matter of future work. In addition, the receiver front end needs to handle the high dynamic range of the room illumination and the influences from external light sources such as light bulbs (modulated or un-modulated) or sunlight.

The digital signal processing (DSP) board used in the current prototype has an on-board stereo codec with 96 kHz maximum sampling frequency. Therefore, the OFDM signal bandwidth is limited to 45 kHz. Clearly, this low bandwidth limits the achievable data rate. As part of another project at Jacobs University Bremen, an optical OFDM system is currently being implemented on field programmable gate array (FPGA) boards having fast on-board data converters running at 100 Msps. The target data rate of 100 Mbps would be sufficient for high quality audio and video broadcasting.

Furthermore, a challenging difficulty for VLC is the compatibility with the commonly used pulse-wide modulation (PWM) dimming. Therefore, further study is required to combine the PWM dimming signal with the OFDM signal or to find...
more suitable dimming technique for VLC systems based on OFDM.
A Publications

A.1 Patent Publications


A.2 Paper Under Review


A.3 Published Papers


Bibliography


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