POWER LINE COMMUNICATION SYSTEMS

A dissertation submitted to the University of Manchester
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in the Faculty of Engineering and Physical Sciences

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By
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Abstract

The remarkably increasing demand for communication systems has recently forced the research community to consider power line (PL) networks for data transmission, which is commonly referred to as power line communications (PLC). In particular, this technology becomes more attractive in harsh wireless environments where radio spectrum is scarce or/and propagation loss is high such as in underground structures and buildings with metal walls. PLC can support many applications such as home-networking, internet and smart grid. More specifically, PLC is considered the backbone of smart grids, not only because no extra wiring installation is required, but also because PLC is a through-grid technology which could reduce the reliance of the utility companies on third party connectivity and, consequently, overcome many security and privacy issues. On the other hand, since PLs were not designed for data transmission, communication signals over such channels can degrade tremendously. Contrary to many other communication channels, noise over PLs cannot be described as additive white Gaussian noise. It is rather categorized broadly into impulsive noise and background noise with the former being the most crucial element influencing PLC systems. With this in mind, this thesis will primarily focus on studying and developing advanced techniques and algorithms to reduce the severity of impulsive noise over PL channels.

The contributions of this thesis are described as follows. Initially, a thorough review is provided to introduce and compare the challenges facing PLC, PL channel and noise modelling schemes, as well as some noise mitigation techniques. Next, novel approaches are proposed, with different degrees of effectiveness and complexity, to reduce the effect of impulsive noise in orthogonal frequency-division multiplexing (OFDM)-based PLC systems. Firstly, an adaptive hybrid preprocessing system is introduced to improve the performance of the conventional hybrid approach. In this respect, closed-form
expressions for the output signal-to-noise ratio (SNR), probability of missed detection and probability of successful detection are derived. Secondly, and unlike existing works which are entirely based on countering impulsive noise at the receiver side, we show that if the OFDM signal is preprocessed at the transmitter side in such a way to minimize the signal peaks, the noise cancellation process can be made more efficient at the receiver. This is accomplished by applying a peak-to-average power ratio reduction scheme such as selective mapping. A closed-form expression for the probability of blanking error of this system is derived. Thirdly, to eliminate the estimation requirement problem of the short-term noise statistics associated with the aforementioned approaches, we propose the dynamic peak-based threshold estimation (DPTE) method. This method relies on utilizing the OFDM signal peak estimates with which optimal blanking is achieved irrespective of the noise parameters. In addition, to realize DPTE, a look-up table algorithm with uniform quantization is exploited and investigated in various system configurations.

Furthermore, this thesis explores the performance of multi-carrier code-division multiple access (MC-CDMA) systems over the multipath PL channel contaminated with Middleton class-A noise for various spreading codes, namely, Pseudonoise, Walsh-Hadamard and orthogonal poly-phase sequences. Different nonlinear preprocessors are implemented and the corresponding performance is evaluated in terms of the output SNR and symbol error rate.
Declaration

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• Last but not least, I gratefully acknowledge the funding received towards my PhD from the University of Manchester.
Preface

Mr. Khaled Maauf Rabie received the B.Sc. degree (with First Class Honors), coming top of his class, in Electrical and Electronic Engineering (EEE) from University of Tripoli, Libya, in 2008. He then continued his study at the University of Manchester and obtained an M.Sc. degree (with Distinction) in Communication Engineering in 2010, again coming top of his class, and received the best student award sponsored by Agilent Technologies. Since 2011, he has been pursuing his Ph.D. degree with the Microwave and Communication Systems (MCS) group at the University of Manchester with a focus on power line communication channel modelling, smart grid applications and signal processing for interference mitigation. His recent research interests also include energy harvesting, wireless power transfer and MIMO systems.

Mr. Rabie has served as a reviewer for several IEEE and IET journals and many international conferences in the area of power line communications and smart grids. He also chaired a conference session at the IEEE international conference on communications (ICC) in June 2015 and was invited in the same year to serve as a TCP member at the IEEE international conference on computer and information technology (CIT’15). He had three successful grant applications to attend summer and training schools in Bowness (UK, 2013), Trento (Italy, 2015) and Barcelona (Spain, 2015). In addition, he received two international invitations to deliver presentations by Khalifa University of Science, Technology and Research (UAE, 2013) and University of Klagenfurt (Austria, 2014).

Recently, Mr. Rabie has received the best student paper award at the IEEE International Symposium on Power Line Communications and its applications (ISPLC’15) in Austin, Texas, US.
To My Father,

For his endless support and encouragement
**List of Abbreviations**

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<tr>
<td>ARIB</td>
<td>Association of Radio Industries and Businesses</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<tr>
<td>BER</td>
<td>Bit Error Rate</td>
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<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
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<tr>
<td>BTG</td>
<td>Blanking Threshold Gain</td>
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<tr>
<td>CCDF</td>
<td>Complementary Cumulative Distribution Function</td>
</tr>
<tr>
<td>CE</td>
<td>Constant Envelope</td>
</tr>
<tr>
<td>CENELEC</td>
<td>European Committee for Electrotechnical Standardization</td>
</tr>
<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
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<tr>
<td>CP</td>
<td>Cyclic Prefix</td>
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<tr>
<td>CTF</td>
<td>Channel Transfer Function</td>
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<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
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<td>DPTE</td>
<td>Dynamic Peak-based Threshold Estimation</td>
</tr>
<tr>
<td>EMC</td>
<td>Electromagnetic Compatibility</td>
</tr>
<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency-Division Multiple Access</td>
</tr>
<tr>
<td>IDFT</td>
<td>Inverse Discrete Fourier Transform</td>
</tr>
<tr>
<td>IFDMA</td>
<td>Interleaved Frequency-Division Multiple Access</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter-Symbol Interference</td>
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<tr>
<td>LFDMA</td>
<td>Localized Frequency-Division Multiple Access</td>
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<tr>
<td>LUT</td>
<td>Look-up Table</td>
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<td>MAC</td>
<td>Media Access Control</td>
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<td>MC-CDMA</td>
<td>Multi-Carrier Code Division Multiple Access</td>
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<td>MMSE</td>
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<td>OFDM</td>
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<td>OPP</td>
<td>Orthogonal Poly Phase</td>
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<td>PTS</td>
<td>Partial Transmit Sequence</td>
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<td>PAPR</td>
<td>Peak-to-Average Power Ratio</td>
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<td>PL</td>
<td>Power Line</td>
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<td>PLC</td>
<td>Power Line Communications</td>
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<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
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<td>PDF</td>
<td>Probability Density Function</td>
</tr>
<tr>
<td>PMF</td>
<td>Probability Mass Function</td>
</tr>
<tr>
<td>PN</td>
<td>Pseudonoise</td>
</tr>
<tr>
<td>PRIME</td>
<td>PoweRline Intelligent Metering Evolution</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>SC-FDMA</td>
<td>Single-Carrier Frequency-Division Multiple Access</td>
</tr>
<tr>
<td>SER</td>
<td>Symbol Error Rate</td>
</tr>
<tr>
<td>SINR</td>
<td>Signal-to-Impulsive Noise Ratio</td>
</tr>
<tr>
<td>SLM</td>
<td>Selective Mapping</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>TF</td>
<td>Transfer Function</td>
</tr>
<tr>
<td>WH</td>
<td>Walsh-Hadamard</td>
</tr>
<tr>
<td>WS</td>
<td>Window Size</td>
</tr>
</tbody>
</table>
List of Mathematical Notations

\[ \prod \] product symbol
\[ \sinh (.) \] hyperbolic sine function
\[ \cosh (.) \] hyperbolic cosine function
\[ \tanh (.) \] hyperbolic tangent function
\[ \exp (x) \] exponential function \( e^x \)
\[ \sum \] summation symbol
\[ j \] imaginary unit \( j^2 = -1 \)
\[ \text{erfc}(.) \] complementary error function
\[ f(.) \] probability density function
\[ \mathbb{E} [.] \] expectation of a random variable
\[ ! \] factorial operator
\[ \mathcal{N} (\mu, \sigma^2) \] normal distribution with mean \( \mu \) and variance \( \sigma^2 \)
\[ \Pr(x) \] probability of \( x \)
\[ \log_x(.) \] logarithmic function to base \( x \)
\[ \ln(.) \] natural logarithm
\[ |.| \] magnitude of a complex number
\[ (. )^* \] conjugate
\[ \text{min} \] argument of the minimum
\[ \text{max} \] argument of the maximum
\[ (. )^T \] transpose
* convolution operation

$Q(.)$ Gaussian Q-function
List of Variables

\( H \)  Transfer function
\( V_s \)  Source voltage
\( V_L \)  Load voltage
\( Z_s \)  Source impedance
\( Z_L \)  Load impedance
\( Z_{eq} \)  Equivalent impedance
\( \rho \)  Propagation constant
\( d \)  Distance
\( R \)  Resistance
\( G \)  Conductance
\( C \)  Capacitance
\( L \)  Inductance
\( N_p \)  Number of paths
\( \tau \)  Time delay
\( m \)  Attenuation exponent factor
\( a_o/a_1 \)  Attenuation parameters
\( g_i \)  Weighting factor  \\
\( v_p \)  Velocity of propagation  \\
\( f \)  Frequency  \\
\( A \)  Impulsive index  \\
\( \Lambda \)  Inverse of impulsive index  \\
\( \Gamma \)  Gaussian-to-impulsive ratio  \\
\( \sigma^2_{m} \)  Variance of \( m^{th} \) impulsive noise pulse  \\
\( \sigma^2_{u} \)  Variance of total noise  \\
\( \sigma^2_{g} \)  Variance of Gaussian noise  \\
\( \sigma^2_{i} \)  Variance of impulsive noise  \\
\( \sigma^2_{s} \)  Variance of the transmitted signal  \\
\( \text{SNR}_{G} \)  Input signal-to-noise ratio  \\
\( \mu_z \)  Expected value of \( z \)  \\
\( p \)  Probability occurrence of impulsive noise  \\
\( P_{ber} \)  Bit error rate  \\
\( n_t \)  Total noise signal  \\
\( n_g \)  AWGN noise signal  \\
\( n_i \)  Impulsive noise signal  \\
\( P_{ser} \)  Symbol error rate  \\
\( d_c \)  Distance between QAM constellation points  \\
\( N \)  Number of sub-carriers
$T_b$  Blanking threshold

$T_b^{\text{opt}}$  Optimal blanking threshold

$T_c$  Clipping threshold

$T$  Adaptive hybrid threshold

$r_k$  Input of the nonlinear preprocessor

$y_k$  Output of the nonlinear preprocessor

$\overline{\gamma}$  Average output SNR

$s(t)$  Continuous-time transmitted signal

$s_k$  Discrete-time transmitted signal

$S_k$  Complex constellations of the data symbols

$K_o$  Scaling factor

$d_k$  The cumulative noise term

$\alpha$  Proportionality constant

$G_r$  Relative gain

$P_b$  Probability of blanking error

$P_m$  Probability of missed blanking/clipping

$P_s$  Probability of successful detection

$B$  Blanking event

$\bar{B}$  Absence of blanking event

$D$  Constellation Order

$H_0$  Null hypothesis
$H_1$ Alternate hypothesis

$T_s$ Active symbol interval

$U$ Number of phase sequences

$W$ Phase sequence vectors

$A_r$ Amplitude of the received signal

$h_k$ Channel impulse response

$k$ Time index

$e$ Estimation error

$N_q$ Quantization levels

$N_b$ Quantization bits representing OFDM symbol peaks

$e_q$ Quantization error

$P$ OFDM signal peak

$\hat{P}$ Estimated OFDM symbol peak

$\bar{P}$ OFDM-PTS symbol peak

$b$ Bernoulli process parameter

$\beta$ Phase weighting factor

$R_F$ Resolution factor

$s$ Output of PAPR reduction modulator (SLM and PTS)

$\bar{M}$ Number of total users

$M$ Number of disjoint sub-blocks

$\bar{N}$ Spreading code length
\( L \)  Over sampling rate

\( \phi \)  Number of OPP code phases
Chapter 1

Introduction

The congested wireless spectrum and its increasing cost have driven researchers, developers and investors to look for more sustaining and low-cost communication alternatives. No doubt that the existing power line (PL) network, which reaches almost every single building on the planet, can be a very promising solution. Using such networks for communications, commonly known as power line communication (PLC), becomes particularly more appealing in harsh wireless environments where radio spectrum is scarce or and propagation loss is high such as in underground structures and buildings with metal walls. The idea of data transmission over PLs is not new; in fact, the first communication attempt over such cables took place in 1900s for reading meters at remote locations [2]. A few decades later, PLs were considered for voice transmission using single-carrier narrow-band schemes operating in low frequency-bands and provided data rates in the range of few Kbps. Over the recent decades, the rising dependence on communications has increased significantly and owing to the advances in modulation techniques, error detection and correction schemes as well as signal processing, it has become feasible to exploit PLs for high-speed home-networking with data rates comparable to that provided by other wired and wireless technologies [3–6].

In addition, PLC has made many smart grid services and functionalities possible such as remote power grid configuration, sensing, dynamic pricing, smart metering and load control [7–9]. Although it is true that for more efficient realization of smart grids, a heterogeneous set of networks should be adopted since no single technology can be a perfect solution for all scenarios [8], the fact that PLC incurs no additional installation costs makes it more
attractive to smart grid developers than the other alternatives. Not only that, but also because PLC is a through-grid technology, privacy and security issues can be considerably reduced \[10\]. However, since PLs were not designed for data transmission, communication signals over these cables can suffer from various impairments, some of which are briefly discussed below.

### 1.1 Challenges Facing PLC

PL channels are a very harsh and noisy transmission medium suffering tremendously from significant non-Gaussian noise, frequency selectivity, high levels of frequency-dependent attenuation, electromagnetic interference and time-varying topology issues. Many studies, however, have concluded that noise remains the most crucial element influencing the reliability of PLC systems, both in narrow-band (NB) and broad-band (BB) PLC technologies \[11–13\].

#### 1.1.1 Non-Gaussian Noise

Noise over PLs is broadly classified into background noise and impulsive noise, more details are given in Chapter 2 \[11, 13, 14\]. The former type usually remains stationary over periods of seconds and minutes or even hours whereas the latter is time-variant in terms of microseconds to milliseconds, frequently exceeding the signal symbol length and can dramatically degrade communication signals \[13, 15\]. In addition, impulsive noise has very high power spectral density (PSD) always exceeding the PSD of background noise by at least 10–15 dB and may occasionally reach as much as 50 dB \[16\]. Typical sources of this noise include fluorescent, halogen lamps and switching power suppliers.

#### 1.1.2 Multipath Propagation

PL networks are usually made of a variety of conductor types joined almost randomly which results in discontinuity and impedance mismatching. This, consequently, leads to deep notches in the channel frequency response as will be presented in Chapter 2. In general, this phenomena becomes less observed
at low frequencies and can be overcome using multi-carrier modulation systems [17].

1.1.3 Frequency-Dependent Attenuation

In addition to the above impairments, the characteristic impedance of the PL channel between any transmitting and receiving nodes can significantly attenuate communication signals and this attenuation increases remarkably with increasing the operating frequency.

1.1.4 Electromagnetic Regulations

The other major burden of PLC is the electromagnetic compatibility (EMC) to other wireless systems. It is well acknowledged that electric wires may radiate electromagnetic waves at high frequencies. Therefore, in order to prevent any interference to other wireless systems operating in the same frequency band, the transmitted power over PLs is restricted and should comply with the regulations determined by independent and governmental regulatory agencies of that specific geographical area [18]. Other parameters that can affect the radiated emissions from PLs include the electrical characteristics of the cables and the network structure [19].

1.1.5 Varying Topology

The transfer function (TF) between any two outlets is both time and frequency variant, affected by plugging/unplugging and/or switching on/off appliances connected to the network. Generally, the load variations are significant and can range from a few ohms up to the order of Kilo-ohms [20].

Clearly, PL networks are not well-suited for communication signals [21]. Thus, it is essential to overcome the aforementioned inherent challenges of such networks. In particular, the focus of this thesis will be on impulsive noise mitigation since it is the most dominant factor [13, 22].

1.2 PLC Categories

PLC is generally divided into two main categories as briefly discussed below.
CHAPTER 1. INTRODUCTION

1.2.1 Narrow-Band PLC (NB-PLC)

NB-PLC operates in the frequency range (3 – 500 KHz) exploiting the low and medium voltage distribution networks and is able to deliver a few Kbps in single-carrier systems and up to 800 Kbps in multi-carrier systems. Figure 1.1 illustrates the frequency bands of NB-PLC technology defined by the main international standards which are

- Federal Communications Commission (FCC) in US.
- Industry Canada in Canada.
- Association of Radio Industries and Businesses (ARIB) in Japan.
- European Committee for Electrotechnical Standardization (CENELEC) in the EU.

Clearly, unlike the first three, the European CENELEC standard is more restrictive [23]. It is also worthwhile mentioning that the first industry-developed multi-carrier NB-PLC standards are the PoweRline Intelligent Metering Evolution (PRIME) and G3-PLC which can provide data rates of up to 130 Kbps and 208 Kbps, respectively [24, 25]. The main outstanding feature of G3-PLC over PRIME is its use of adaptive tone mapping where the system automatically uses the part of the spectrum which has the least amount of noise to maximize data rate. Moreover, ITU-T G.hnem and IEEE 1901.2 are two more advanced NB-PLC standards based on PRIME and G3-PLC, and are able to provide data rates of 1 Mbps and 500 Kbps, respectively.
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<table>
<thead>
<tr>
<th></th>
<th>NB-PLC</th>
<th>BB-PLC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data rate</td>
<td>Up to 800 Kbps</td>
<td>Over 1 Mbps</td>
</tr>
<tr>
<td>Frequency</td>
<td>Up to 500 KHz</td>
<td>Over 2 MHz</td>
</tr>
<tr>
<td>Modulation</td>
<td>FSK, S-FSK, BPSK, SS, OFDM</td>
<td>OFDM</td>
</tr>
<tr>
<td>Applications</td>
<td>Building Automations</td>
<td>Internet</td>
</tr>
<tr>
<td></td>
<td>Renewable Energy</td>
<td>HDTV</td>
</tr>
<tr>
<td></td>
<td>Advanced Metering</td>
<td>Audio</td>
</tr>
<tr>
<td></td>
<td>Street Lighting</td>
<td>Gaming</td>
</tr>
<tr>
<td></td>
<td>Electric Vehicle</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Smart Grid</td>
<td></td>
</tr>
</tbody>
</table>

Table 1.1: Comparison between NB-PLC and BB-PLC technologies.

1.2.2 Broad-Band PLC (BB-PLC)

BB-PLC functions in the frequency band \((1.8 - 250 \text{ MHz})\) utilizing the low voltage distribution network \([4,8]\) and providing data-rates of up to 1.2 Gbps for home-area networks. Table 1.1 shows a comparison between the NB-PLC and BB-PLC technologies in terms of data rates, frequency bands, modulation techniques and some applications.

1.3 Key Contributions

The major contributions of this thesis are briefly listed below.

- Derivation of the signal-to-noise ratio (SNR) expression at the output of the adaptive hybrid device as well as addressing the optimization problem of the threshold and proportionality constant. In addition, the constellation sensitivity of an orthogonal frequency-division multiplexing (OFDM) system with blanking, clipping, conventional hybrid and adaptive hybrid schemes is investigated.

- Derivation of the probability of blanking error expression for the OFDM-based system with nonlinear preprocessing at the receiver when selective mapping (SLM) is implemented at the transmitter. Various noise
and peak-to-average power ratio (PAPR) scenarios are studied and significant conclusions are drawn.

- Proposal of a novel method to intelligently reduce the effect of impulsive noise over PL networks without requiring any estimation of noise characteristics at the receiving end. More specifically, this only requires estimates of the signal peaks at the receiver. To achieve this, we also introduced a look-up table (LUT) based algorithm with uniform quantization. This method is referred to as dynamic peak-based threshold estimation (DPTE) method.

- Investigating the performance of various multi-carrier code division multiple access (MC-CDMA) systems over the multipath PLC channel in the presence of Middleton class-A noise. Different spreading codes are considered such as Pseudonoise (PN), Walsh- Hadamard (WH) and orthogonal poly-phase (OPP).

- Winning the best student paper award at the IEEE International Symposium on Power Line Communications and its applications (ISPLC’15) in Austin, Texas, US. This paper examines the performance of constant envelope OFDM (CE-OFDM) over the PLC channel and presents very interesting conclusions.

1.4 Thesis Organization

The remainder of this thesis is organized as follows. Chapter 2 presents a literature review on the relevant background of PLC fundamentals such as PLC channel characterization and noise modeling schemes. In addition, a number of impulsive noise cancellation techniques are reviewed including the application of three nonlinear pre-processors at the receiver’s front-end, namely, blanking, clipping and conventional hybrid (combined blanking-clipping). Chapter 3 proposes a new scheme to enhance the capability of the conventional hybrid-based system. Chapter 4 introduces a novel technique to reduce the effect of impulsive noise by preprocessing the OFDM signal at the transmitter side in order to make the noise cancellation process more efficient at the receiver. This is achieved by utilizing the well-known PAPR reduction
scheme, SLM. A closed-form mathematical expression for the probability of blanking error is derived. Chapter 5 demonstrates a new impulsive noise mitigation technique that does not rely on the noise statistics by utilizing estimates of the transmitted signal’s peak value. Furthermore, to realize this scheme, a LUT-based algorithm with uniform quantization is proposed and then various LUT sizes are investigated. Chapter 6 is dedicated to examine the performance of MC-CDMA systems over PL channels for various classes of spreading sequences such as PN, WH and OPP. Chapter 7 concludes the thesis and outlines prospects for the future extension of the work.

1.5 List of Publications

Journal Papers


CHAPTER 1. INTRODUCTION


Conference Papers


CHAPTER 1. INTRODUCTION

Other Papers


International Invited Presentations


Chapter 2

Literature Review

This chapter begins with reviewing different PL channel modeling schemes and then explores the impact of various network parameters on the channel transfer function (CTF). The noise types over PLs are then discussed and some noise modeling schemes are presented including the Middleton class-A and the two-component Gaussian models. The performance of single-carrier and multi-carrier modulation systems are also compared and then different impulsive noise mitigation techniques are investigated, namely, the application of nonlinear preprocessors at the front-end of the receiver such as blanking, clipping and conventional hybrid devices.

2.1 PLC Channel Modeling

As discussed previously, PL networks differ considerably from the other conventional communication medium representing unfavorable environment for data transmission at high frequencies [16, 17, 26]. Over the last years, a lot of research has been conducted investigating the indoor and outdoor PL channel properties [16, 17, 27–32]. Generally, PL channels are modeled using two approaches, namely, bottom-up and top-down, both are discussed below.

2.1.1 Bottom-Up Approach

This model is also referred to as the deterministic model where the PL CTF is determined using transmission line theories such as the $ABCD$ or the scattering matrices [32–35]. On one hand, this approach has two main disadvantages
High computational complexity.

- Perfect knowledge requirement of the network parameters such as cable lengths, characteristic impedance, loads etc.

On the other hand, this approach is relatively flexible since all network parameters can be formulated [26,36].

### 2.1.1.1 Analytical Modeling

For the sake of simplicity, in this section we compute the TF of a one-branch network which can then be easily extended to multiple-branch networks. It is commonly known that \(ABCD\) representation for a two-port network is a very convenient way of calculating TFs. Figure 2.1 shows the relationship between the input/output voltages and currents of a two-port network which can be represented mathematically as [26]

\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix}
A & B \\
C & D
\end{bmatrix} \begin{bmatrix}
V_2 \\
I_2
\end{bmatrix},
\]

(2.1)

where \(A, B, C\) and \(D\) are appropriately chosen constants.

The beauty of \(ABCD\) representation is that if there is a cascade of two-port circuits, as shown in Figure 2.2 for example, the overall TF is simply the product of all \(ABCD\) matrices for the individual two-port networks.

The TF \(H\) of the circuit presented in Figure 2.1 can therefore be easily calculated as

\[
H = \frac{V_L}{V_s} = \frac{Z_L}{AZ_L + B + C Z_L Z_s + D Z_s},
\]

(2.2)
Figure 2.2: A circuit consisting of a cascade of two-port circuits.

where $V_s$ and $V_L$ are the source and load voltages whereas $Z_s$ and $Z_L$ are the source and load impedances, respectively.

If, for instance, the PL has one tap bridge as illustrated in Figure 2.3a, then it can be simplified as in Figure 2.3b by replacing the tap bridge with its equivalent input impedance, $Z_{eq}$, at terminals $A$ and $B$.

Figure 2.3: A PL with one bridge tap connection and its equivalent circuit.

Hence, the overall TF of this circuit can be determined simply as

$$\varphi = \prod_{i=0}^{3} \varphi_i$$

where
\[ \varphi_0 = \begin{bmatrix} 1 & Z_s \\ 0 & 1 \end{bmatrix}, \] (2.4)

\[ \varphi_1 = \begin{bmatrix} \cosh(\gamma_1 d_1) & Z_1 \sinh(\gamma_1 d_1) \\ \frac{1}{Z_1} \sinh(\gamma_1 d_1) & \cosh(\gamma_1 d_1) \end{bmatrix}, \] (2.5)

\[ \varphi_2 = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \] (2.6)

and

\[ \varphi_3 = \begin{bmatrix} \cosh(\rho_2 d_2) & Z_2 \sinh(\rho_2 d_2) \\ \frac{1}{Z_2} \sinh(\rho_2 d_2) & \cosh(\rho_2 d_2) \end{bmatrix}, \] (2.7)

while \( Z_1, Z_2, \rho_1 \) and \( \rho_2 \) are the characteristic impedances and propagation constants for the second and fourth sub-circuits in Figure 2.3b, \( Z_{eq} \) is given by

\[ Z_{eq} = \frac{Z_c Z + Z_c \tanh(\rho d)}{Z_c + Z \tanh(\rho d)}. \] (2.8)

By substituting (2.4)–(2.7) into (2.3), it is straightforward to show that the \( ABCD \) parameters can be calculated as

\[ A = \cosh(\rho_2 d_2) \eta + \frac{\sinh(\rho_2 d_2) \kappa}{Z_2}, \] (2.9)

\[ B = Z_2 \cosh(\rho_2 d_2) \eta + \cosh(\rho_2 d_2) \kappa, \] (2.10)

\[ C = \cosh(\rho_2 d_2) \xi + \frac{\sinh(\rho_2 d_2) \vartheta}{Z_2}, \] (2.11)

\[ D = Z_1 \sinh(\rho_1 d_1) \xi + \cosh(\rho_2 d_2) \vartheta, \] (2.12)

where
\[ \eta = \cosh(\rho_1 d_1) + \frac{Z_s \sinh(\rho_1 d_1)}{Z_1}, \]  
\[ (2.13) \]

\[ \kappa = Z_1 \sinh(\rho_1 d_1) + Z_s \cosh(\rho_1 d_1), \]  
\[ (2.14) \]

\[ \xi = \frac{Z_1 \cosh(\rho_1 d_1) + Z_s \sinh(\rho_1 d_1) + Z_{eq} \sinh(\rho_1 d_1)}{Z_1 Z_{eq}}, \]  
\[ (2.15) \]

and

\[ \vartheta = \frac{Z_1 \sinh(\rho_1 d_1) + Z_s \cosh(\rho_1 d_1)}{Z_{eq} + \cosh(\rho_1 d_1)}. \]  
\[ (2.16) \]

The overall TF can now be calculated using (2.2) for which numerical examples are presented below.

### 2.1.1.2 Numerical Examples

In this section numerical results of the PL CTF are obtained for various configurations in the frequency range from 0 Hz to 30 MHz. All our investigations here are based on the per-unit length parameters illustrated in Table 2.1 where \( R, G, C \) and \( L \) denote the cable resistance, conductance, capacitance and inductance, respectively.

<table>
<thead>
<tr>
<th></th>
<th>( R )</th>
<th>( G )</th>
<th>( C )</th>
<th>( L )</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1.9884 ( \Omega )/m</td>
<td>0.01686 ( nS/m )</td>
<td>0.13394 ( nF/m )</td>
<td>362.8 ( nH/m )</td>
</tr>
</tbody>
</table>

Table 2.1: Pre-unit length parameters of the cables.

- Different Cable Lengths

The impact of cable length on the PL CTF without bridge taps is presented in Figure 2.4 for two different cable lengths, 7 m and 10 m when \( Z_s = 50 \Omega \) and \( Z_L = 60 \Omega \). It is noticeable from this figure that the attenuation increases as the cable length is increased and that the CTF shape also changes with varying the cable length.

- Impact of One Bridge Tap
The effect of bridge taps on the PL CTF is examined here. Figure 2.5 depicts the CTF with and without a bridge tap when $Z_s = 50 \, \Omega$, $Z_L = 50 \, \Omega$, cable length is 4 m, bridge tap length is 2 m and branch impedance is 50 \, \Omega$. Comparing the two figures, it is clearly visible that the presence of a bridge tap will significantly increase the attenuation in addition to changing the CTF shape.

- **Multiple Bridge Taps with Various Lengths and Impedances**

Now, we investigate the influence of increasing the number of bridge taps as shown in Figure 2.6. It is apparent that increasing the number of bridge taps considerably intensifies the attenuation. The network parameters used here are $Z_s = 50 \, \Omega$, $Z_L = 60 \, \Omega$, cable length = 10 m and the bridge taps characteristics are listed in Table 2.2.

<table>
<thead>
<tr>
<th>Tap</th>
<th>Tap 1</th>
<th>Tap 2</th>
<th>Tap 3</th>
<th>Tap 4</th>
<th>Tap 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Three Taps (Lengths, m)</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Three Taps (Impedance, ( \Omega ))</td>
<td>50 + j100</td>
<td>150</td>
<td>450</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Five Taps (Lengths, m)</td>
<td>2</td>
<td>5</td>
<td>1</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Five Taps (Impedance, ( \Omega ))</td>
<td>50 + j100</td>
<td>150</td>
<td>450</td>
<td>100 + j50</td>
<td>200</td>
</tr>
</tbody>
</table>

Table 2.2: Bridge taps parameters [1].
Figure 2.5: PL CTF with and without a bridge tap.

Figure 2.6: PL CTF with 3 and 5 bridge taps.
2.1.2 Top-Down Approach (Empirical)

This approach is the most common technique used to determine PL CTFs where the channel is considered as a black box and the system parameters are obtained from experimental measurements. Using this approach, a simple PL channel model was first proposed by Hensen in [37]. This model, however, did not take into consideration the multipath phenomenon which has a great influence in determining the PL CTF. Later on, Philipps in [27] introduced a more accurate model considering multipaths and is given by

$$H(f) = \sum_{i=1}^{N_p} g_i \exp \left( -j2\pi f \tau_i \right), \quad (2.17)$$

while $g_i$ is the weighting factor carrying the effect of the PL and reflections for each of the $N_p$ paths along route $i$, $f$ is the frequency and $\tau_i$ is the corresponding time delay. Since (2.17) does not account for the signal flow attenuation, Philipps’s model was extended by Zimmermann in [17] to have the following form

$$H(f) = \sum_{i=1}^{N_p} g_i \exp \left( (a_o - a_1 f^m) d_i \right) \exp \left( -j2\pi f \frac{d_i}{v_p} \right), \quad (2.18)$$

where

- $a_o$ and $a_1$ are the attenuation parameters.
- $m$ is the attenuation exponent factor ranges between 0.2 and 1.
- $d_i$ is the length of path $i$.
- $v_p$ is the velocity of propagation.

This model has been the most widely adopted by PLC researchers. In order to verify that (2.18) characterizes the PL CTF adequately, numerical results of this model were compared with real experimental results in [30]. For convenience, these results are reproduced in Figure 2.7. It should be pointed out that the same network sample is used to produce both results when $m = 1$, $a_o = -2.1 \times 10^{-3}$, $a_1 = 8.1 \times 10^{-10}$, in addition to the parameters shown in Table 2.3. Now, comparing the results in Figure 2.7, it can be concluded that this model covers the essential properties of the PL channel and can be effectively and reliably used to study and evaluate the performance of PLC systems by means of simulations.
2.2 Noise Over PL Channels

Unlike many other communication channels, noise over PLs can not be described as additive white Gaussian noise (AWGN) \([13, 14]\). In fact, according to \([14]\), noise over PLs is categorized into four types and this was later extended to five types in \([13]\) as demonstrated in Figure 2.8, namely, colored background noise, narrow-band noise, periodic impulsive noise asynchronous to the mains frequency, periodic impulsive noise synchronous to the mains frequency and aperiodic impulsive noise. The first two types vary

<table>
<thead>
<tr>
<th>Path number</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length (d_i \text{ (m)})</td>
<td>200</td>
<td>221</td>
<td>242</td>
<td>259</td>
<td>266</td>
<td>530</td>
</tr>
<tr>
<td>Weighting factor (g_i)</td>
<td>0.54</td>
<td>0.275</td>
<td>-0.15</td>
<td>0.08</td>
<td>-0.03</td>
<td>-0.02</td>
</tr>
</tbody>
</table>

Table 2.3: Parameters of the sample model.
slowly over time and remain stationary over periods of up to minutes and sometimes for hours; therefore they are summarized as background noise. However, the last three types are classified as impulsive noise having a short duration, random occurrence and high PSD which makes this noise the most dominant factor degrading communication signals over PLC channels. For example in [13], it was experimentally found that the PSD of impulsive noise always exceeds the PSD of background noise by at least 10 – 15 dB and may reach as much as 50 dB. The major sources of this noise are the devices connected to the PL network such as electronic drills, light dimmers and switching power supplies.

A considerable amount of research has been carried out to model the noise over PLs, see e.g. [21, 38–42] and the references therein. However, the main models are the Middleton class-A noise model, [39,40,43], and the two-component Gaussian model [44,45]. It is important to mention that these models incorporate both the background noise and impulsive noise components.

### 2.2.1 Middleton Class-A Noise Model

This model is widely used to evaluate the performance of PLC systems [46], in which the amplitudes of background and impulsive noise are Gaussian-distributed and its probability density function (PDF) is given as

\[
f_Z(z) = \sum_{m=0}^{\infty} \exp\left(-A\right) \frac{A^m}{m!} \frac{1}{\sqrt{2\pi}\sigma_m^2} \exp\left(-\frac{z^2}{2\sigma_m^2}\right), \tag{2.19}
\]
where

\[ \sigma^2_m = \sigma^2_u \left( \frac{m}{A} + \Gamma \right), \quad (2.20) \]

\[ \sigma^2_u = \sigma^2_g + \sigma^2_i, \quad (2.21) \]

\[ \Gamma = \frac{\sigma^2_g}{\sigma^2_i}, \quad (2.22) \]

while \( \sigma^2_m \) denotes the variance of the \( m^{th} \) considered impulsive noise source, \( \sigma^2_u \), \( \sigma^2_g \) and \( \sigma^2_i \) represent the total noise, Gaussian noise and impulsive noise variances, respectively. Also, \( A \) is the average number of impulsive sources simultaneously active and is referred to as \textit{impulsive index}. Having a closer look at (2.19), it can easily be noticed that the PDF of Middleton class-A model is basically a sum of weighted normal distributions. It is also clear that the three parameters \( A, \Gamma \) and \( \sigma^2_u \) specify the statistical characteristics of this model and that when \( A \) is large, impulsive noise will become continuous and, therefore, Middleton class-A noise becomes more likely as Gaussian noise; while conversely, low values of \( A \) mean rare and highly structured noise pulses. In addition, the mean and variance of Middleton class-A noise are, respectively, defined as [18]

\[
\mathbb{E}[z] = \int z f_Z(z) \, dz = \sum_{m=0}^{\infty} \frac{\exp(-A) A^m}{m! (2\pi \sigma^2_m)} \int z \exp \left( -\frac{z^2}{2\sigma^2_m} \right) \, dz = 0, \quad (2.23)
\]

and

\[
\sigma^2_z = \mathbb{E}[z^2] = \frac{\exp(-A) \sigma^2_g}{\Gamma} \sum_{m=0}^{\infty} \frac{A^m}{m! \left( \frac{m}{A} + \Gamma \right)}. \quad (2.24)
\]

Furthermore, it is found that the bit error rate (BER) performance of binary phase shift keying (BPSK) in the presence of Middleton class-A can be calculated as [39]

\[
P_{ber} = \frac{\exp(-A)}{2} \sum_{m=0}^{\infty} \frac{A^m}{m!} \text{erfc} \left( \sqrt{\frac{(1 + \Gamma) E_b}{m (m + \Gamma) N_0}} \right). \quad (2.25)
\]
In order to demonstrate the impact of Middleton class-A noise model parameters on its PDF and BER performance, we plot in Figures 2.9 and 2.10 some numerical results of (2.19) and (2.25), respectively. It is apparent that when $\Gamma$ is kept constant at 0.001 and the impulsive index is varied from a large value to a small value, two main cases can be highlighted. Firstly, when $A$ is large, the characteristic distribution of the noise is very similar to Gaussian distribution, see Figure 2.9a, and so is the BER performance, see Figure 2.10a. Secondly, however, when $A$ becomes small, the distribution shows very impulsive characteristics and therefore the corresponding BER performance degrades dramatically. Similarly, same observations can be noticed from Figures 2.9b and 2.10b when $\Gamma$ is changed while keeping $A$ constant. In general, clearly impulsive noise can significantly degrade the BER performance compared to the case of Gaussian noise.

2.2.2 Two-Term Approximation

In [47], (2.19) was truncated from the infinite series form to two terms as in (2.26), i.e. $m \in \{0, 1\}$.

$$f_Z(z) = \frac{\exp(-A)}{\sqrt{2\pi \sigma_o^2}} \exp\left(-\frac{z^2}{2\sigma_o^2}\right) + \frac{1 - \exp(-A)}{\sqrt{2\pi \sigma_1^2}} \exp\left(-\frac{z^2}{2\sigma_1^2}\right). \quad (2.26)$$

Hence, the corresponding error probability of the BPSK system in (2.25) becomes

$$P_{ber} = \frac{\exp(-A)}{2} \text{erfc}\left(\sqrt{\frac{E_b}{2\sigma_o}}\right) + \frac{1 - \exp(-A)}{2} \text{erfc}\left(\sqrt{\frac{E_b}{2\sigma_1}}\right). \quad (2.27)$$

It is also shown that for small $A$ and $\Gamma$ values, only an error of less than 2% occurs due to this truncation. To highlight this, a comparison of the BER performance when $A = 0.1$ and $\Gamma = 0.1$ is illustrated in Figure 2.11 for the following systems

- Using the infinite series, (2.25).
- Using a 2nd order approximation, $(m = 0, 1)$.
- Using a 3rd order approximation, $(m = 0, 1, 2)$. 
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Figure 2.9: The effect of Middleton class-A noise model parameters on its PDF.

(a) Middleton class-A PDFs for different values of $A$ when $\Gamma = 0.001$.

(b) Middleton class-A PDFs for different values of $\Gamma$ when $A = 0.01$. 

PDF

Random Variable ($z$)
Figure 2.10: The effect of the Middleton class-A noise model parameters on the BER performance.
Clearly, the $2^{\text{nd}}$ order approximation produces a noticeable error whereas the $3^{\text{rd}}$ order approximation represents the series more accurately. According to [47], the latter scenario accounts for more than 99% of the probability mass function of (2.19).

### 2.2.3 Two-Component Gaussian Model

In this model, the total noise is represented as \([44, 48-51]\)

$$n_t = n_g + b n_i$$  \hspace{1cm} (2.28)

where $n_t$ denotes the total noise, $n_g$ is the AWGN component with zero mean and variance $\sigma_g^2$, $n_i$ is Gaussian-distributed with mean zero and a large variance $\sigma_i^2$, ($\sigma_i^2 > \sigma_g^2$), and $b \in \{0, 1\}$ is the Bernoulli process with parameter $p$ such that
Pr(b) = \begin{cases} p, & b = 1 \\ (1 - p), & b = 0 \end{cases} \quad (2.29)

where \( p \) represents the probability occurrence of impulsive noise. Therefore, the PDF of the two-component Gaussian noise model can be expressed as

\[ f_{n_t}(z) = \frac{1 - p}{\sqrt{2\pi \sigma^2_g}} \exp \left( -\frac{z^2}{2\sigma^2_g} \right) + \frac{p}{\sqrt{2\pi \left( \sigma^2_g + \sigma^2_i \right)}} \exp \left( -\frac{z^2}{2 \left( \sigma^2_g + \sigma^2_i \right)} \right). \quad (2.30) \]

Now, given the information signal variance is \( \sigma^2_s \), \( \sigma^2_g \) and \( \sigma^2_i \) will define the input signal-to-noise ratio (SNR\(_G\)) and the signal-to-impulsive noise ratio (SINR), respectively, as follows

\[ \text{SNR}\(_G\) = 10 \log_{10} \left( \frac{\sigma^2_s}{\sigma^2_g} \right), \quad (2.31) \]

and

\[ \text{SINR} = 10 \log_{10} \left( \frac{\sigma^2_s}{\sigma^2_i} \right). \quad (2.32) \]

### 2.3 Single-Carrier and Multi-Carrier Modulation

In this section we review and compare the performance of single-carrier and multi-carrier systems over impulsive noise PLC channels [44, 52, 53]. These comparisons adopt the two-component Gaussian noise model.

#### 2.3.1 Single-Carrier Modulation

Many researchers have studied the performance of single-carrier systems over impulsive noise channels. For instance, the author of [44] derived a closed-form expression for the SER performance of a quadrature amplitude modulation (QAM) system, with \( 2^M, M = 2, 4, 6, \ldots \) constellation points, and this is
reproduced in (2.33) for convenience.

\[ P_{\text{ser}}^{(QAM)} = 1 - \frac{(1 - p)}{2^{M-2}} \left[ \left( (1 - A_o) + \left( 2^{\frac{M}{2}-1} - 1 \right) (1 - 2A_o) \right)^2 \right] \]

\[ - \frac{p}{2^{M-2}} \left[ \left( (1 - A_N) + \left( 2^{\frac{M}{2}-1} - 1 \right) (1 - 2A_N) \right)^2 \right]. \tag{2.33} \]

where

\[ A_o \triangleq Q \left( \frac{d_c}{\sqrt{2\sigma_g}} \right), \tag{2.34} \]

\[ A_N \triangleq Q \left( \frac{d_c}{\sqrt{2 \left( \sigma_g^2 + \frac{\sigma_i^2}{p} \right)}} \right), \tag{2.35} \]

and \( d_c \) represents the distance between the constellation points whereas \( Q(\cdot) \) is the Gaussian Q-function defined as

\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp \left( -\frac{x^2}{2} \right) dx. \tag{2.36} \]

To illustrate the impact of noise parameters on this system, some numerical results of (2.33) are plotted in Figure 2.12 as a function of SINR for various values of \( p \). It is clearly noticeable from these results that as \( p \) increases the SER performance degrades. In addition, increasing SINR improves performance for all the pulse probabilities. The other observation one can see is that when SINR is small (SINR \( \rightarrow 0 \)), the symbol probability of error becomes equal to the corresponding noise probability which is intuitive.

### 2.3.2 Multi-Carrier Modulation

In multi-carrier modulation systems, such as OFDM, the transmitted signal, assuming perfect synchronization, is given by

\[ s_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} S_k \exp \left( j\frac{2\pi nk}{N} \right), \quad k = 0, 1, 2, \ldots, N - 1 \tag{2.37} \]
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Figure 2.12: SER performance of 16-QAM single-carrier system in an impulsive noise environment for various noise scenarios when $\text{SNR}_{G} = 30$ dB.

where $S_k$ is the complex constellations of the data symbols, $N$ is the number of sub-carriers and $k$ is the time index. In this case, a mathematical expression for the SER performance in the presence of impulsive noise can be written as \[44\]

\[
P_{\text{ser}}^{\text{MCM}} = 1 - \frac{1}{2^{M-2}} \sum_{m=0}^{N} \binom{N}{m} p^m (1 - p)^{N-m} 
\]

\[
\times \left( (1 - A_m) + \left( 2^{M-1} - 1 \right) (1 - 2A_m) \right)^2
\]

(2.38)

where

\[
A_m \triangleq Q \left( \frac{d_c}{2\sigma_m} \right),
\]

(2.39)

\[
\sigma_m^2 = \frac{\sigma_g^2}{2} + \frac{m\sigma_i^2}{2pN},
\]

(2.40)

and
\[
\binom{N}{m} = \frac{N!}{(N-m)!m!}
\] (2.41)

Figure 2.13 depicts the SER performance of an OFDM system with impulsive noise for different values of \( p \). The base-band modulation used here is 64-QAM and the number of sub-carriers is 128 \((N = 128)\). For comparison’s sake, the SER performance of a single-carrier system using same base-band modulation is also included on the same plot. It is obvious that at relatively high SINR values, the multi-carrier system outperforms the single-carrier one. This can be justified by the fact that in the case of single-carrier modulation a noise impulse will affect one data symbol whereas in multi-carrier schemes the energy of this impulse is spread over \( N \) data symbols due to the discrete Fourier transform (DFT) operation, hence reducing its severity. However, when the opposite it true, i.e. SINR is low, multi-carrier modulation suffers more since all \( N \) symbols are now subject to high interference even after spreading the impulses over \( N \) symbols resulting in a worse performance than that of single-carrier systems. In other words, multi-carrier systems can provide a code diversity effect which is able, somewhat, to reduce the impact of impulsive noise; interestingly enough, however, this advantageous effect could turn into a disadvantage if the noise energy exceeds a certain threshold. In such scenarios, additional signal processing techniques must be deployed. To achieve this, many techniques have been reported in the literature with different degrees of effectiveness and complexity but the simplest and most efficient of which remains the application of nonlinear preprocessors at the receiver’s front-end.

### 2.4 Nonlinear Preprocessors

In order to further improve the impulsive noise cancellation process especially in high impulsive noise environments, nonlinear preprocessing devices are usually applied at the receiver’s front-end to blank or/and clip the received signal when it exceeds a certain threshold value(s), see Figure 2.14. In general, there are three main nonlinear preprocessors namely: blanking, clipping and conventional hybrid (joint blanking-clipping) [51, 54–60]. The reason for choosing these techniques is because of their simplicity and ease of
Figure 2.13: SER performance of multi-carrier and single-carrier systems as a function of SINR for several values of $p$ when $\text{SNR}_G = 30$ dB. In both systems, 64-QAM modulation is used.

implementation which makes them widely implemented in practice [61–64]. The basic principles of these devices are

- **Blanking**
  
  $$y_k = \begin{cases} r_k, & |r_k| \leq T_b \\ 0, & |r_k| > T_b \end{cases} \quad k = 0, 1, \ldots, N - 1$$  

  while $T_b$ is the blanking threshold, $|.|$ is the absolute value operator, $r_k$ and $y_k$ are the input and output of the nonlinear preprocessor, respectively.

- **Clipping**
  
  $$y_k = \begin{cases} r_k, & |r_k| \leq T_c \\ T_c \exp(j \text{arg}(r_k)), & |r_k| > T_c \end{cases} \quad k = 0, 1, \ldots, N - 1$$  

  where $T_c$ is the clipping threshold.

- **Conventional Hybrid**
Figure 2.14: Block diagram of an OFDM system with nonlinear preprocessing at the receiver’s front-end.

\[ y_k = \begin{cases} 
  r_k, & |r_k| \leq T_c \\
  T_c \exp(j \text{arg}(r_k)), & T_c < |r_k| \leq T_b \\
  0, & |r_k| > T_b 
\end{cases} \quad (2.44) \]

where \( T_b = 1.4 T_c \) in (2.44). It is clear from (2.42)–(2.44) that only the amplitude of the received signal is processed while the phase is always kept unchanged. The blanking or/and clipping threshold(s) should be carefully selected to optimize the system performance. For instance, if the threshold is too small, many unaffected samples of the OFDM signal will be blanked/clipped resulting in poor BER performance; whereas for very large threshold(s), impulsive noise will be overlooked and will become part of the detected signal leading also to performance degradation.

Many studies have evaluated the performance of OFDM-based receivers with blanking [56,61–63,65]. For instance, the authors in [61] and [56] discussed the performance of such systems using computer simulations whereas numerically-based results were obtained first in [62]. However, theoretical performance analysis to find an analytical expression for the SNR at the output of the blanking nonlinear preprocessor first appeared in [65] as

\[ \bar{\gamma} = \frac{2}{\mathbb{E}[A_n^2]} \quad (2.45) \]

where \( \bar{\gamma} \) is the average output SNR and \( \mathbb{E}[A_n^2] \) is given by (2.46). In addition, a closed-form expression for the optimal blanking threshold, \( T_{b_{\text{opt}}} \), that maximizes the output SNR is formulated as a function of the noise parameters which is given by (2.47) [65].

Later on, the same author extended this analysis in [66,67] to include the other two nonlinear preprocessing schemes. In these studies, it is shown that
\[ E \left[ A_n^2 \right] = 2 (1 - p) \left[ \sigma_g^2 (1 - \sigma_i^2) \left( \frac{T_b^2}{2 (1 + \sigma_i^2)} + 1 \right) \right. \]
\[ \times \exp \left( -\frac{T_b^2}{2 (1 + \sigma_i^2)} \right) + 2 p \left[ (\sigma_g^2 + \sigma_i^2) + (1 - \sigma_g^2 - \sigma_i^2) \right. \]
\[ \left. \times \left( \frac{T_b^2}{2 (1 + \sigma_g^2 + \sigma_i^2)} + 1 \right) \exp \left( -\frac{T_b^2}{2 (1 + \sigma_g^2 + \sigma_i^2)} \right) \right]. \] (2.46)

\[ T_b^{\text{opt}} = \sqrt{\frac{2 (1 + \sigma_g^2) (1 + \sigma_g^2 + \sigma_i^2)}{\sigma_i^2}} \ln \left( \frac{\left[ 1 + \sigma_g^2 + \sigma_i^2 \right]^2 (p - 1) (1 - \sigma_g^2)}{p (1 - \sigma_g^2 - \sigma_i^2)} \right). \] (2.47)

the output signal of the nonlinear device can be given as \( y_k = K_o s_k + d_k \) where \( K_o \) is an appropriately selected scaling factor, \( s_k \) is the transmitted OFDM signal and \( d_k \) is the cumulative noise term. This decomposition is justified by the application of Bussgang’s theorem [68]. It is also found that when \( K_o \) is chosen as \( K_o = (1/2) E \left[ |y_k s_k^*|^2 \right] \), the SNR at the output of the blanking, clipping and conventional hybrid nonlinear preprocessors can be expressed as

\[ \gamma = \frac{E[|K_o s_k|^2]}{E[|y_k - K_o s_k|^2]} = \frac{2K_o^2}{E_{out} - 2K_o^2}, \] (2.48)

where \( K_o^2 \) and \( E_{out} \), depending on the nonlinear device used, can be given as

- **Blanking**

\[ K_o^{(b)} = 1 - \sum_{l=0}^{\infty} \left( 1 + \frac{T_b^2}{2 (1 + \sigma_i^2)} \right) p_l \exp \left( -\frac{T_b^2}{2 (1 + \sigma_i^2)} \right). \] (2.49)

\[ E_{out}^{(b)} = 2 + 2 \sum_{l=0}^{\infty} p_l \left( \sigma_i^2 - \frac{T_b^2}{2} + (1 + \sigma_i^2) \right) \exp \left( -\frac{T_b^2}{2 (1 + \sigma_i^2)} \right). \] (2.50)

- **Clipping**
\[
K^{(c)}_o = 1 - \sum_{l=0}^{\infty} p_l \left[ \exp \left( -\frac{T_c^2}{2(1+\sigma_l^2)} \right) - \sqrt{\frac{\pi}{2}} \frac{T_c}{\sqrt{1+\sigma_l^2}} Q \left( \frac{T_c}{\sqrt{1+\sigma_l^2}} \right) \right]. 
\]

\[
E^{(c)}_{out} = 2 + \sum_{l=0}^{\infty} p_l \left( 2\sigma_l^2 - (1 + \sigma_l^2) \exp \left( -\frac{T_c^2}{2(1+\sigma_l^2)} \right) \right). 
\]

• Conventional Hybrid

\[
K^{(ch)}_o = 1 - \sum_{l=0}^{\infty} p_l \left[ \exp \left( -\frac{T_c^2}{2(1+\sigma_l^2)} \right) + \frac{T_c T_b}{2(1+\sigma_l^2)} \exp \left( -\frac{T_b^2}{2(1+\sigma_l^2)} \right) \right] 
\]

\[
+ \sum_{l=0}^{\infty} p_l \sqrt{\frac{\pi}{2}} \frac{T_c}{\sqrt{1+\sigma_l^2}} \left[ Q \left( \frac{T_c}{\sqrt{1+\sigma_l^2}} \right) - Q \left( \frac{T_b}{\sqrt{1+\sigma_l^2}} \right) \right]. 
\]

\[
E^{(ch)}_{out} = 2 + \sum_{l=0}^{\infty} p_l \left( \sigma_l^2 - (1 + \sigma_l^2) \exp \left( -\frac{T_c^2}{2(1+\sigma_l^2)} \right) - \frac{T_c^2}{2} \exp \left( -\frac{T_b^2}{2(1+\sigma_l^2)} \right) \right). 
\]

From (2.48)-(2.54), it is clear that when \( T_{b/c} \to \infty \), i.e. no blanking/clipping is performed (typical OFDM receiver), (2.48) can be simplified to

\[
\gamma \left( T_{b/c} \to \infty \right) = 10 \log_{10} \left( \frac{\sigma_s^2}{\sigma_g^2 + p\sigma_i^2} \right). 
\]

To visualize the importance of the threshold selection, we plot in Figure 2.15 some numerical examples of (2.48) for the three nonlinear preprocessing-based systems. As expected, it can be observed that too low and too high threshold values will significantly degrade performance for the reasons mentioned above and that good selection of the threshold is the key to achieve best performance. In this respect, to optimize the output SNR, it is more convenient to rewrite (2.48) as

\[
(\gamma)^{-1} = \frac{E_{out}}{2K^2_o} - 1. 
\]

It is clear that the optimal blanking/clipping thresholds cannot be expressed in closed-forms; hence only numerical results can be obtained by satisfying the following argument
Figure 2.15: Output SNR versus blanking/clipping thresholds for the blanking, clipping and conventional hybrid systems when $p = 0.01$, SINR = $-10$ dB and SNR$_G$ = $25$ dB. For the conventional hybrid system: $T_c = T$, $T_b = 1.4T$.

$$\arg\min_{T_{b/c}} \left\{ \frac{E_{out}}{K_o^2} \right\}. \quad (2.57)$$

Now, to highlight this phenomena, we demonstrate in Figure 2.16 the optimal blanking/clipping thresholds and the corresponding maximum achievable output SNR versus SINR. For comparison sake, results of the typical OFDM receiver (2.55) are also included on this plot from which it is clearly seen that this system always has the worst performance. Notably, the conventional hybrid system has the best performance compared to the other two nonlinear preprocessing systems. Nonetheless, the former can be further improved as will be discussed in the next chapter.
Figure 2.16: Optimal blanking/clipping threshold and the corresponding maximum achievable output SNR versus SINR when $p = 0.01$ and $\text{SNR}_G = 25$ dB.
2.5 Chapter Summary

This chapter focused on PL channel characterization and investigated the impact of various network parameters on the system TF. Two PL channel modeling methods were analyzed, namely, the bottom-up and top-down approaches. It was shown that PL channels are frequency-dependent and time-varying affected by the devices connected to the outlets as well as by the number of branches between the transmitting and receiving points.

This chapter also reviewed the different noise types over PLs and presented some noise modeling schemes that are commonly used to evaluate the performance of PLC systems. Noise over PLs is broadly classified into background noise and impulsive noise with the latter being the most dominant factor degrading communication signals. It was shown that multi-carrier systems are generally more robust against impulsive noise than single-carrier systems due to the spreading effect of the noise energy over $N$ sub-carriers. It was however presented that when the impulsive noise energy becomes relatively high, preceding the OFDM demodulator with a nonlinear preprocessor such as blanking, clipping or conventional hybrid device can further improve the system performance. Moreover, the conventional hybrid preprocessor was found to always outperform the blanking and clipping based systems.
Chapter 3

Adaptive Hybrid Nonlinear Preprocessing

This chapter is dedicated to enhance the capability of the conventional hybrid preprocessing based system presented in the previous chapter. This proposed system will be referred to as adaptive hybrid. Closed-form expressions for the output SNR, probability of missed detection and probability of successful detection are derived for this system and compared with that of the conventional hybrid scheme. Furthermore, the SER performance of the four nonlinear preprocessors will be evaluated.

3.1 Introduction

As shown in the previous chapter, the conventional hybrid system is characterized by two thresholds $T_1$ and $T_2$ ($T_2 = \alpha T_1$), where $\alpha$ is a proportionality constant. This system assumes a fixed value for the proportionality constant and it was found that optimizing the threshold $T_1$ is the key to improve performance. In contrast, in this chapter we show that the performance of the hybrid method is sensitive not only to the threshold but also to the proportionality constant. With this in mind, we propose to enhance the capability of this method by optimizing the two parameters. Hence, the contributions of this chapter are as follows. First, a mathematical expression for the output SNR as a function of the threshold and proportionality constant is formulated and then used to optimize the performance. Second, the probability of missed
detected, probability of successful detection of impulsive noise and the SER performance of the optimized system are analyzed. Results reveal that using an adaptive hybrid technique with an optimally selected threshold and proportionality constant can outperform other nonlinear schemes.

The basic principle of the adaptive hybrid device is given as

\[
y_k = \begin{cases} 
  r_k, & |r_k| \leq T \\
  T \exp(j \text{arg}(r_k)), & T < |r_k| \leq \alpha T \\
  0, & |r_k| > \alpha T
\end{cases} \\
\text{for } k = 0, 1, \ldots, N - 1
\]

(3.1)

where \(T\) is the hybrid threshold and \(\alpha > 1\).

### 3.2 Performance Analysis of the Output SNR

The SNR performance at the output of the adaptive hybrid device can also be calculated using (2.48) where \(K_o\) and \(E_{out}\) for this system are given by (3.2) and (3.3), respectively. These parameters are found by, respectively, replacing \(T_c\) and \(T_b\) with \(T\) and \(\alpha T\) in (2.53) and (2.54).

\[
K_o^{(ah)} = 1 - \sum_{l=0}^{\infty} p_l \left[ \exp \left( -\frac{T^2}{2(1 + \sigma_l^2)} \right) + \frac{T^2}{2(1 + \sigma_l^2)} \exp \left( -\frac{\alpha^2 T^2}{2(1 + \sigma_l^2)} \right) \right] - \sum_{l=0}^{\infty} p_l \sqrt{\frac{\pi}{2}} \frac{T}{\sqrt{1 + \sigma_l^2}} \left[ Q \left( \frac{T}{\sqrt{1 + \sigma_l^2}} \right) - Q \left( \frac{\alpha T}{\sqrt{1 + \sigma_l^2}} \right) \right] .
\]

(3.2)

\[
E_{out}^{(ah)} = 2 + 2 \sum_{l=0}^{\infty} p_l (\sigma_l^2 - (1 + \sigma_l^2)) \times \exp \left( -\frac{T^2}{2(1 + \sigma_l^2)} \right) - \frac{T^2}{2} \exp \left( -\frac{\alpha^2 T^2}{2(1 + \sigma_l^2)} \right) .
\]

(3.3)
To illustrate the impact of $T$ and $\alpha$ on the output SNR, some numerical results are presented in Figure 3.1. This figure shows a 3D surface plot of the output SNR as a function of $T$ and $\alpha$, using (2.48), (3.2) and (3.3) for $p = 0.01$, and 0.1 when $\text{SNR}_G = 25$ dB and $\text{SINR} = -30$ dB. From this figure, a general trend can be observed, for the two pulse probabilities, that when $T$ and $\alpha$ are too small \{$T \ll 2$ and $\alpha \ll 3$\} the system performance worsens radically. At the other extreme, when $T$ and $\alpha$ are too high \{$T \to \infty$ and $\alpha \to \infty$\}, this will also cause performance deterioration and the output SNR approaches $-10$ dB and $-20$ dB for $p = 0.01$, and 0.1, respectively. Comparing Figures 3.1a and 3.1b, it is evident that the output SNR is more sensitive to the proportionality constant when $p$ is relatively high. It is also interesting to note that a good choice of both $T$ and $\alpha$ will maximize the output SNR.

### 3.3 System Performance Optimization

In this section we present some numerical results for the optimal threshold and optimal proportionality constant that satisfy the following

$$\arg \min_{T, \alpha} \left\{ \frac{E_{\text{out}}}{K_0^2} \right\}.$$  

(3.4)

In addition, the corresponding maximum achievable output SNR is investigated under various noise conditions. These analytical results are validated with simulations which, from this point onward unless it is stated otherwise, are based on an OFDM system with $N = 256$ sub-carriers, 16-QAM modulation, $\sigma_s^2 = (1/2)\mathbb{E}[|s|^2] = 1$, $\sigma_g^2 = (1/2)\mathbb{E}[|n_g|^2]$, $\sigma_i^2 = (1/2)\mathbb{E}[|n_i|^2]$, and $\text{SNR}_G = 25$ dB. Figure 3.2 illustrates the numerical and simulated results of the optimal adaptive threshold when $\alpha$ is optimized. Note that results for the optimal blanking, clipping and conventional hybrid thresholds are also included on this graph. The optimal proportionality constant corresponding to the optimized adaptive hybrid threshold is plotted versus SINR in Figure 3.3 for various impulsive noise probabilities. In both figures, it is clearly visible that the simulated results match the analytical ones.

In general, it is obvious that the behavior of the optimal values for $T$ and $\alpha$ for the adaptive system can be divided into two regions during which these parameters behave differently. These regions can be defined as the high SINR
Figure 3.1: 3D surface plot of the SNR at the output of the adaptive hybrid device as a function of $T$ and $\alpha$ when $\text{SNR}_G = 25$ dB and $\text{SINR} = -30$ dB for $p = 0.01$ and 0.1.
region \( \{ 0 \to -6 \text{ dB} \} \) and low SINR region \( \{-6 \text{ dB} \to -\infty \} \). In the former region, it is interesting to observe that the optimal adaptive threshold matches the optimal clipping threshold and this corresponds to the high dependency of the optimal \( \alpha \) on the impulsive noise characteristics as shown in Figure 3.3. To elaborate, having a large value for \( \alpha \) means that the blanking threshold \( (\alpha T) \) will be too high and therefore the vast majority of the received samples will not be blanked. As a consequence, clipping becomes the dominant process which justifies why the optimal \( T \) of the adaptive system approaches the clipping threshold. Moreover, it should be pointed out that the variation in the optimal \( \alpha \) increases as \( p \) becomes higher making the selection of this parameter even more crucial in heavily disturbed impulsive noise environments. On the other hand, in the low SINR region the optimal \( T \) of the proposed system starts to diverge from the clipping threshold and approaches the conventional hybrid threshold. However, the corresponding optimal \( \alpha \) drops sharply and remains almost constant at about 1.4 which is equal to the proportionality constant value of the conventional hybrid technique (2.44). This clearly explains
CHAPTER 3. ADAPTIVE HYBRID NONLINEAR PREPROCESSING

Figure 3.3: The optimal proportionality constant (with optimized $T$) versus SINR for various values of $p$ when $\text{SNR}_G = 25$ dB.

why the optimal $T$ of the adaptive system approaches that of the conventional hybrid technique for very low SINR values.

In order to calculate the maximum achievable SNR at the output of the adaptive hybrid device, the numerically found optimal $T$ and optimal $\alpha$ values are substituted into (2.48), (3.2) and (3.3). Figure 3.4 depicts the maximum achievable output SNR versus SINR for the proposed system with different impulsive noise probabilities. Additionally, the output SNR curves of the blanking, clipping and conventional hybrid systems along with the typical OFDM receiver are also included on this plot. It can be seen that the adaptive hybrid technique offers the best performance which can be best quantified in terms of the relative gain. This gain, given by (3.5), is defined as the gain in the output SNR obtained by the adaptive hybrid technique, $\gamma_{ah}$, over the conventional hybrid technique, $\gamma_{ch}$, and is plotted in Figure 3.5 for various values of $p$.

$$G_r = 10 \log_{10} \left( \frac{\gamma_{ah}}{\gamma_{ch}} \right).$$  \hspace{1cm} (3.5)
Figure 3.4: Maximum achievable output SNR versus SINR for the blanking, clipping, conventional hybrid, adaptive hybrid and typical OFDM systems with various noise probabilities when $\text{SNR}_G = 25$ dB.
As apparent, the relative gain is directly proportional to $p$ and can be as high as 0.62 dB at about SINR = −4 dB when $p = 0.1$. The intuitive explanation of this is that when $p$ is high the decision accuracy of whether to blank or clip becomes more critical and this is where the adaptive hybrid technique is most effective as it optimizes the blanking/clipping proportionality constant which guides the decision process. This can also be extracted from Figure 3.3 where the optimal $\alpha$ variation increases for higher noise probabilities. On the other hand, when $p$ decreases the gain becomes negligible and hence the conventional hybrid technique could be applied instead since it is simpler. Furthermore, it is evident that, irrespective of the noise probability, the conventional and adaptive hybrid systems perform similarly when SINR is very low. This can be easily explained as follows: in such an environment the impulsive noise amplitude is so high, compared to the OFDM signal, that it can be identified perfectly with either technique. In other words, the decision accuracy of whether to blank or clip will have less influence on the overall performance. For better understanding of the impact of the proposed technique on the system performance we investigate two important performance
measures which highly depend on the impulsive noise characteristics. These measures are the probability of miss \((P_m)\) and the probability of success \((P_s)\).

### 3.4 Probability of Miss and Probability Success

\(P_m\) is defined as the probability that the affected received signal samples, \(A_r = |r_k|\), are not blanked/clipped and is given by the joint probability \(\Pr (\bar{B}, \mathcal{H}_1)\) where \(\bar{B}\) denotes the absence of blanking/clipping event and the alternate hypothesis \((\mathcal{H}_1)\) indicates the presence of impulsive noise. Hence, \(P_m\) can be mathematically expressed as

\[
P_m = \Pr (A_r < T \mid \mathcal{H}_1) \Pr (\mathcal{H}_1).
\]  

On the other hand, \(P_s\) is defined as the probability of correctly blanking/clipping the contaminated samples and is given by the joint probability \(\Pr (B, \mathcal{H}_1)\) where \(B\) denotes the presence of blanking and/or clipping. Therefore, \(P_s\) can also be expressed as

\[
P_s = \Pr (A_r > T \mid \mathcal{H}_1) \Pr (\mathcal{H}_1).
\]

In the presence of impulsive noise, the amplitude of the received signal has Rayleigh distribution with parameter \(\sigma^2 = \sigma_s^2 + \sigma_g^2 + \sigma_i^2\). Hence, its PDF is

\[
f_{A_r} (r \mid \mathcal{H}_1) = \frac{r}{(\sigma_s^2 + \sigma_g^2 + \sigma_i^2)} \exp \left( -\frac{r^2}{2 (\sigma_s^2 + \sigma_g^2 + \sigma_i^2)} \right).
\]

Therefore, \(P_m\) can be simply found as

\[
P_m = \int_{-\infty}^{T} f_{A_r} (r \mid \mathcal{H}_1) \, dr
= p \left( 1 - \exp \left( -\frac{T^2}{2 (\sigma_s^2 + \sigma_g^2 + \sigma_i^2)} \right) \right)
\]

whereas \(P_s\) is determined as
\[ P_s = \int_{T}^{\infty} f_{A_r}(r | \mathcal{H}_1) \, dr = p \exp \left( -\frac{T^2}{2(\sigma^2_s + \sigma^2_w + \sigma^2_i)} \right). \] (3.10)

Figure 3.6 depicts some numerical results of (3.9) and (3.10) as a function of SINR for an OFDM receiver with an adaptive hybrid device when \( \text{SNR}_G = 25 \) dB for various values of \( p \). Again, results for the blanking, clipping and conventional hybrid systems are included. In this evaluation, \( T \) in both (3.9) and (3.10) is replaced with the optimal blanking/clipping thresholds obtained previously. In general, it is seen that the behaviors of \( P_m \) and \( P_s \) are inversely proportional. It can also be noted that as the noise becomes more impulsive (SINR \( \rightarrow -\infty \)), \( P_m \) is minimized and \( P_s \) is maximized for the four systems; whereas at the other extreme, for high SINR values (SINR \( \rightarrow 0 \)), \( P_m \) becomes very high and \( P_s \) worsens. Moreover, it is interestingly noticeable that the adaptive hybrid technique, in terms of \( P_s \), always outperforms the conventional hybrid technique but in terms of \( P_m \) this is true only in the intermediate SINR region; however, in the low SINR region both techniques perform similarly. Comparing Figures 3.6a and 3.6b, it can be observed that the enhancement in both \( P_m \) and \( P_s \) increases as \( p \) becomes higher and this justifies why the output SNR, in the previous section, improves for higher values of \( p \). It is to be emphasized at this point that \( P_m \) and \( P_s \) can not be stand-alone measures of the system performance but a good tradeoff of both is more crucial. This encourages us to have a closer look at the signal constellation after the OFDM demodulator and investigate the SER performance more insight.

### 3.5 Constellations’ Sensitivity Discussion

In this section we investigate the effect of the nonlinear preprocessing techniques on the signal constellation at the output of the OFDM demodulator. Two different modulation schemes are considered here namely, quadrature phase shift keying (QPSK) and 16-QAM. However, to achieve communication systems with very high levels of spectral efficiency, very dense QAM
Figure 3.6: Probability of miss and probability of success versus SINR for the blanking, clipping, conventional hybrid and adaptive hybrid systems for $p = 0.01$ and 0.1 when $\text{SNR}_G = 25$ dB.
constellations is usually employed. For example, the current HomePlug AV2 PL standard supports 4096-QAM modulation (12 bits/symbol) [69] which is 4 times higher than the previously released HomePlug AV standard [70,71]. Using higher-order QAM without increasing the BER requires increasing the signal energy, reducing noise, or both. In our investigation here we keep the signal power unchanged and examine the impact of reducing impulsive noise using the four nonlinear preprocessors on the QPSK and 16-QAM constellation points.

Figure 3.7 illustrates the constellation diagram of the QPSK signal in the presence of impulsive noise after the demodulator for the typical OFDM receiver, blanking, clipping, conventional hybrid and adaptive hybrid systems.
A noiseless ideal QPSK signal will have a constellation consisting of distinct points \( \{ \pm 1 \} \) symmetric around the real and imaginary axes—the blue points in Figure 3.7. However, as can be seen from Figure 3.7a impulsive noise intensively dislocates the constellation points. It is observed that when a nonlinear device is applied, Figures 3.7b–d, the scattering is constrained considerably and four clusters centered around the original four points are formed. It can also be noted that the adaptive hybrid technique provides slightly better constellation precision than the conventional hybrid system. It is worthwhile mentioning the fact that the lower the scattering the more accurate the detection decision becomes.

It is commonly known that by moving to a higher-order constellation, it
becomes possible to transmit more bits per symbol. However, if the mean energy of the constellation is kept fixed for a fair comparison, which is the case in this section, the points will be closer together and will therefore be more susceptible to noise. This can be observed by comparing Figures 3.7a and 3.8a even though the pulse probability is 10 times lower compared to that of the QPSK system. In addition, the improvement offered by the nonlinear devices becomes insignificant as the constellation size is increased. It is also interesting to see from Figures 3.7d and 3.8d that the adaptive hybrid system provides negligible enhancement relative to that obtained with the QPSK system. For more quantitative and meaningful results, we assess next the SER performance of the two aforementioned QAM constellations under different noise conditions.

3.6 SER Performance

This section is dedicated to analyze the minimum achievable SER performance of the four nonlinear preprocessing methods. To analytically evaluate the SER performance in OFDM receivers with nonlinear preprocessors, two assumptions are usually made.

1. The noise at the output of the OFDM demodulator approaches Gaussian distribution if the number of sub-carriers is sufficiently large ($N = 8192$ subcarriers in this case) [72–74].

2. Sufficient number of samples within an OFDM symbol should be contaminated with impulsive noise such that $N p \gg 1$.

With this in mind, conventional SER prediction techniques can be used. Figures 3.9a and 3.9b illustrate the SER performance versus SINR for the adaptive hybrid system when $p = 0.1$ and $0.03$ for QPSK and 16-QAM modulation, respectively. Again for comparison-sake, we have included the SER results of the blanking, clipping, conventional hybrid and typical OFDM receivers. The analytical results are straightforwardly obtained by substituting the maximum achievable output SNR calculated previously into [75]

$$P_{ser} = 1 - \left[ 1 - 2 \left( 1 - \frac{1}{\sqrt{D}} \right) Q \left( \sqrt{\frac{3\gamma}{D-1}} \right) \right]^2$$

(3.11)
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Figure 3.9: Minimum achievable SER performance versus SINR for the blanking, clipping, conventional hybrid, adaptive hybrid and typical OFDM systems with various impulse probabilities for QPSK and 16-QAM schemes when $\text{SNR}_G = 25$ dB.
where $D$ is the constellation order which is in this case 4 and 16 for QPSK and 16-QAM, respectively.

As a first observation, one can see from Figure 3.9 that the adaptive hybrid technique establishes the lower bound performance in all impulsive noise scenarios whereas the typical OFDM receiver always has the worst performance. In general, and as anticipated, systems with 16-QAM modulation are more sensitive to impulsive noise compared to QPSK systems for the reasons discussed earlier. In both constellations, however, it is worth stressing that the adaptive hybrid system enhances the performance at the expense of slightly higher computational complexity at the receiver since two parameters are optimized. Besides, it is interesting to note that the SER performance of the typical OFDM receiver follows that of the systems with nonlinear preprocessors when $\text{SINR} \gtrsim -4 \text{ dB}$ and it diverges as $\text{SINR}$ becomes smaller. This is simply because the OFDM demodulator provides a code diversity effect which is able, to some extent, to mitigate some of the impulsive noise impact. On the other hand, when $\text{SINR} \rightarrow -\infty$, the SER performance approaches 1 regardless of the noise probability of occurrence which implies that in such environments the exploitation of a nonlinear preprocessor becomes even more appealing.

Finally, it should be highlighted that all the OFDM nonlinear preprocessing-based systems discussed so far are sensitive to the optimal thresholds selection, which are mainly dependent on noise statistics. In other words, to achieve best performance, noise characteristics, in the form of $\text{SINR}$ and $p$, must be made available at the receiver-end which can be very challenging to accomplish in practice due to the dynamic nature of the PLC channel. If this condition is not satisfied, such systems will suffer from dramatic performance degradation as will be shown later in this thesis.

3.7 Chapter Summary

This chapter showed that the performance of the hybrid system is not purely threshold-dependent but is also sensitive to the proportionality constant. In light of this, a new technique was proposed to further enhance the capability of the conventional hybrid scheme by jointly optimizing the threshold and the proportionality constant. Closed-form expressions for the output SNR,
probability of miss and probability of success were derived and validated with simulations. Furthermore, the constellations’ sensitivity and SER performance of the proposed system were analyzed. Results have revealed that the adaptive system can yield up to 0.6 dB SNR improvement and a considerably enhanced SER performance in comparison to the conventional hybrid scheme.
Chapter 4

SLM-Based Impulsive Noise Reduction Scheme

Unlike existing works on PL impulsive noise mitigation, which are based entirely on countering impulsive noise at the receiver side, in this chapter, we propose to preprocess the OFDM signal at the transmitter side in such a way to further enhance the noise cancellation process at the receiver. Note that, for conciseness and without loss of generality, only the performance of blanking nonlinear preprocessors is evaluated in this chapter\(^1\); for the clipping and hybrid schemes, refer to [77].

4.1 Introduction

As briefly pointed out at the end of Chapter 3, the main disadvantage of the blanking method is that in order to optimally suppress impulsive noise, the noise statistics must be accurately known apriori, generally in the form of SINR and noise probability of occurrence. This method will be referred to from now on as conventional blanking. It is known that imperfect recognition of the impulsive noise signals will lead to nulling uncorrupted signal samples resulting in blanking errors and hence performance deterioration [78]. To minimize this, we propose to preprocess the OFDM signal at the transmitter side by simply applying a PAPR reduction technique such as amplitude

\(^1\)Note that part of this work is published in [76].
clipping [79–82], tone reservation [83–86], coding [87–90], partial transmit sequence (PTS) [91–97] and SLM [96, 98–106]. For the work in this chapter, however, we exploit the SLM scheme, as it is well-known for its robustness, and combine it with blanking at the receiver to reduce impulsive noise. Hence, the contribution of this chapter is twofold. First we derive a closed-form expression for the probability of blanking error \( P_b \) and demonstrate how it can be reduced considerably with the proposed system. For more quantitative characterization, the corresponding output SNR is also considered. The second contribution resides in addressing the problem of blanking threshold optimization under various PAPR and impulsive noise scenarios. It will be shown that significant improvements can be achieved with the proposed system relative to the conventional blanking-based receivers.

### 4.2 System Model

The basic system diagram considered in this study is shown in Figure 4.1a and the conventional blanking-based system is illustrated in Figure 4.1b. In both systems, the information bits are first mapped into 16-QAM symbols which are then passed through either an inverse discrete Fourier transform (IDFT) or IDFT/PAPR reduction to produce a time domain signal, \( s(t) \) or \( \bar{s}(t) \), respectively. \( s(t) \) is expressed as in (4.1) whereas \( \bar{s}(t) \) is defined later.

\[
s(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_k \exp\left(\frac{j2\pi kt}{T_s}\right), \quad 0 < t < T_s
\]  

(4.1)

where \( S_k \) is the complex constellations of the data symbols and \( T_s \) is the active symbol interval. The PAPR of the transmitted signal can, therefore, be given as

\[
PAPR = 10 \log_{10} \left( \frac{\max |s(t)|^2}{E[|s(t)|^2]} \right) \quad 0 < t < T_s
\]  

(4.2)

In order to get accurate estimates of the actual PAPR, oversampling by 4
CHAPTER 4. SLM-BASED IMPULSIVE NOISE REDUCTION SCHEME

Figure 4.1: Block diagram of an OFDM system with and without PAPR reduction at the transmitter and blanking at the receiver.

times is deployed in all our investigations in this chapter since such oversampling rate was shown to be sufficient to approximate the true PAPR\(^2\) [83].

As mentioned above, the PAPR reduction scheme used here is SLM. This scheme is based on phase rotations in which the transmitter generates a set of different data blocks representing the same information as the original data block and then selects the one with the minimum PAPR for transmission [98]. Assuming that the data stream is defined as \( S = [S_0, S_1, \ldots, S_{N-1}]^T \), then each data block, \( S \), is multiplied by \( U \) different phase sequence vectors \( (W) \) of length \( N \)

\[
W^{(u)} = \begin{bmatrix} W_0^{(u)}, W_1^{(u)}, \ldots, W_{N-1}^{(u)} \end{bmatrix}^T \quad u = 1, 2, \ldots, U
\]

This multiplication yields \( U \) modified data blocks

\[
\bar{S}^{(u)} = \begin{bmatrix} S_0^{(u)} W_0^{(u)}, S_1^{(u)} W_1^{(u)}, \ldots, S_{N-1}^{(u)} W_{N-1}^{(u)} \end{bmatrix}^T \quad u = 1, 2, \ldots, U
\]

where \([.]^T\) denotes the matrix transpose. The modified blocks are then passed

\(^2\)It should be mentioned that such process may significantly increase the computational complexity of the transmitter as more processing is performed [107].
通过IDFT，SLM-OFDM信号有N子载波。其可以写为

\[ s(n)(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} \bar{s}^{(n)}_k \exp\left( \frac{j2\pi kt}{T_s} \right) \quad 0 < t < T_s \] (4.5)

最终，复数数据块中最小PAPR的选择用于传输，\( \bar{s}(t) \)。该信号然后通过PLC信道传输，该信道会受到背景和脉冲噪声的干扰。我们考虑两种成分的高斯模型（2.30）来描述这些噪声类型。在完美同步条件下，取决于使用常规方法还是SLM方法，接收信号将具有以下形式

\[
r_k = \begin{cases} 
  s_k / \bar{s}_k + n_{g,k}, & H_0 \\
  s_k / \bar{s}_k + n_{g,k} + b n_{i,k}, & H_1 
\end{cases} \quad k = 0, 1, \ldots, N - 1
\] (4.6)

而\( s_k \)和\( \bar{s}_k \)是\( s(t) \)和\( \bar{s}(t) \)的离散形式，\( n_{g,k} \)和\( n_{i,k} \)是独立的。

### 4.3 例解

直观地认为，如果OFDM符号的平均PAPR较小，那么会使得脉冲噪声更容易与有用信号区分，从而可以在接收器上更有效地被屏蔽。这可以通过部署已知的SLM方案来实现。为了进一步澄清，图4.2中给出了传统OFDM、SLM-OFDM和脉冲噪声信号的示意图。该图显示了两种不同的情况。首先，在常规系统中，当阈值考虑时，两个噪声脉冲{IN2和IN3}将被识别，而IN1则未被检测并成为信号的一部分，然后被OFDM解调器接收。之后，符号检测阶段将进行。

\(^3\)信号\( s_k / \bar{s}_k, n_{g,k} \)和\( n_{i,k} \)被假定为相互独立。
Figure 4.2: Improved blanking threshold for SLM-based OFDM system with $N = 64$ and $U = 32$.

demodulator. In contrast, if $T_{b2}$ is used, the blanker will be able to identify \{IN1, IN2 and IN3\}; however, the unaffected samples \{S1, S2 and S3\} will also trigger the blanker and, consequently, will be set to zero causing blanking errors. On the other hand, the SLM-based system allows using $T_{b2}$ without any blanking errors (leaving the unaffected samples untouched) in addition to eliminating \{IN1, IN2 and IN3\}. The amount of reduction in the blanking threshold is referred to as blanking threshold gain ($BTG = T_{b2} - T_{b1}$). It will be shown later that the higher the BTG, the more performance enhancement is achieved in terms of the output SNR.

4.4 Complementary Cumulative Distribution Function (CCDF)

The amount of PAPR reduction is usually measured in terms of the CCDF, which provides the probability that the PAPR of a data block exceeds a given
threshold (PAPR₀). The CCDF of an SLM-based OFDM system with \( U \) statistically independent frames is expressed as \([108, 109]\)

\[
CCDF = (1 - \Pr \{ \text{PAPR} \leq \text{PAPR}_0 \})^U
\]

\[
= \left( 1 - (1 - \exp(-\text{PAPR}_0))^N \right)^U. \tag{4.7}
\]

The amount of PAPR reduction of this system depends on two parameters, namely, \( U \) and \( W \). Throughout our investigations in this chapter, \( W \) will be chosen from the set \( \{\pm 1, \pm j\} \). Now, a plot of (4.7) is shown in Figure 4.3 along with simulation results for a 16-QAM SLM-based signal for different values of \( U \) when \( N = 64 \). It can clearly be seen that the PAPR performance improves as \( U \) is increased and this improvement becomes less significant when \( U \) goes beyond 8 sequences. This reduction in the PAPR implies that more of the transmitted signal energy is contained close to the average value offering larger BTG value and hence lower \( P_b \) will be obtained as presented below.

![Figure 4.3: CCDF plot for the conventional and SLM-based OFDM systems with various values of \( U \) when \( N = 64 \).](image-url)
4.5 Probability of Blanking Error

$P_b$ is defined as the probability that the amplitude of the received sample, $A_r$, exceeds $T_b$ when it is unaffected by impulsive noise and is mathematically given by the following joint probability

\[
P_b = \Pr (B, \mathcal{H}_0) = \Pr (A_r > T_b | \mathcal{H}_0) \Pr (\mathcal{H}_0).
\]  

Equation (4.8) can also be rewritten in terms of the cumulative distribution function (CDF) as

\[
P_b = [1 - F_{A_r} (T_b | \mathcal{H}_0)] \Pr (\mathcal{H}_0)
\]  

where $F_{A_r} (T_b | \mathcal{H}_0)$ is the conditional CDF of the received signal and is simply related to its PDF as

\[
F_{A_r} (T_b | \mathcal{H}_0) = \int_{-\infty}^{T_b} f_{A_r} (r | \mathcal{H}_0) dr.
\]  

4.5.1 Conventional Blanking System

In the absence of impulsive noise, the received signal amplitude for the conventional OFDM system has Rayleigh distribution with parameter $\sigma^2 = \sigma_s^2 + \sigma_g^2$, hence

\[
 f_{A_r}^{\text{(conv)}} (r | \mathcal{H}_0) = \frac{r}{(\sigma_s^2 + \sigma_g^2)} \exp \left( -\frac{r^2}{2(\sigma_s^2 + \sigma_g^2)} \right).
\]

By substituting (4.11) into (4.10) and then (4.10) into (4.9), we get

\[
P_b^{\text{(conv)}} = \exp \left( -\frac{T_b^2}{2(\sigma_s^2 + \sigma_g^2)} \right) (1 - p)
\]
4.5.2 SLM-based Blanking System

In this case, the PDF of the transmitted signal is derived in [110] as a function of \( N \) and \( U \). This expression is reproduced here for the received signal in the absence of impulsive noise as written as

\[
f^{(SLM)}_{A_r}(r \mid \mathcal{H}_0) = U f^{(conv)}_{A_r}(r \mid \mathcal{H}_0) \left( 1 - \exp \left( -\frac{r^2}{2(\sigma_s^2 + \sigma_g^2)} \right) \right)^N U^{-1} \times \left( 1 - \left[ 1 - \exp \left( -\frac{r^2}{2(\sigma_s^2 + \sigma_g^2)} \right) \right] \right) \quad (4.13)
\]

Similar to the conventional blanking system, the conditional CDF of the received SLM-based signal can be determined as

\[
F^{(SLM)}_{A_r}(T_b \mid \mathcal{H}_0) = \int_{-\infty}^{T_b} f^{(SLM)}_{A_r}(r \mid \mathcal{H}_0) \, dr

= \left[ 1 - \left[ 1 - \exp \left( -\frac{T_b^2}{2(\sigma_s^2 + \sigma_g^2)} \right) \right]^N \right] \frac{1}{N} \quad (4.14)
\]

Now, using the definition of \( P_b \) in (4.9), we can write the probability of blanking error for the SLM-based system as

\[
P^{(SLM)}_b = \left( 1 - F^{(SLM)}_{A_r}(T \mid \mathcal{H}_0) \right) \Pr(\mathcal{H}_0)

= \left[ 1 - \left( 1 - \left[ 1 - \exp \left( -\frac{T_b^2}{2(\sigma_s^2 + \sigma_g^2)} \right) \right] \right) \right] \frac{1}{N} \quad (4.15)
\]
Some numerical results obtained from (4.12) and (4.15) are shown in Figure 4.4 along with simulation results for the conventional and SLM-based blanking systems for various values of $U$ when $N = 64$ and $\text{SNR}_G = 25$ dB. It is clear that the simulated curves closely match the analytical ones when $U$ is small and deviate slightly for large values of $U$. It is also obvious that the behavior of $P_b$ can be divided into two regions. The first region is when $T_b \lesssim 2$ during which the SLM-based system does not provide much probability reduction compared to that of the conventional system. When $T_b = 2$, about 10% of the signal samples will exceed this threshold regardless of the number of phase sequences being used. This can also be clearly observed from Figure 4.2 where about 7 samples out of 64 for each system exceed 2, which represent about 10% of the total samples. In the second region ($T_b > 2$), however, it is noticeable that the SLM-OFDM system minimizes $P_b$ in comparison to conventional blanking. In addition, it is evident that $P_b$ is inversely proportional to $U$ and $T_b$. For instance when $U = 16$ and at $T_b = 2.5$, the probability is reduced by about 0.5 order of magnitude whereas for $T_b = 3$, the probability
is minimized by about 3 orders of magnitude. This implies that the system performance will improve for higher values of $U$ as will be further discussed later.

### 4.6 Performance Analysis of the Output SNR

In this section the output SNR performance of the proposed SLM-based technique is examined. In addition, the optimal blanking threshold is investigated in several PAPR and noise scenarios. The simulation parameters used here are: $N = 64$, 16-QAM modulation, $\sigma_s^2 = (1/2)\mathbb{E}[|\bar{s}_k|^2] = 1$, $\sigma_g^2 = (1/2)\mathbb{E}[|n_{g,k}|^2]$ and $\sigma_i^2 = (1/2)\mathbb{E}[|n_{i,k}|^2]$. Similar to (2.48), the output SNR of this system is determined as

$$\gamma_{SLM}^{(U)} = \frac{\mathbb{E}[|K_o \bar{s}_k|^2]}{\mathbb{E}[|y_k - K_o \bar{s}_k|^2]}$$  \hspace{1cm} (4.16)

where $\bar{s}_k = \bar{s}(kT_s/N)$ and $K_o = (1/2)\mathbb{E}[|y_k \bar{s}_k|^2]$.

#### 4.6.1 The Output SNR versus Blanking Threshold

The output SNR of the SLM-based blanking system with $U = 1, 2, 4, 8, 16, 32$ and 64 is plotted in Figure 4.5 as a function of $T_b$ for $p = 0.01$ and 0.001. Results for the conventional blanking system are also added to this plot, the analytical results of which are obtained from (2.48)–(2.50). As expected, when $U$ is increased the output SNR improves and this gain becomes less significant as $U$ goes beyond 8. Also, this improvement is inversely proportional to $p$; for instance, when $p = 0.01$ the gain in the output SNR when $U = 64$ is about 2.25 dB whereas when $p = 0.001$ the gain becomes about 3 dB at the same value of $U$.

For the two impulsive noise probabilities under consideration, there is a general trend that when $T_b$ is too small ($T_b \lesssim 2$) the system performance degrades dramatically. Similarly, if $T_b$ is too high, more noise energy is allowed to be part of the detected signal which deteriorates performance. In this case, the output SNR approaches 10 dB and 20 dB when $p = 0.01$ and 0.001, respectively. Interestingly, for each value of $U$ there exists an optimal blanking threshold at which the output SNR is maximized. Besides, as $U$ increases the
**CHAPTER 4. SLM-BASED IMPULSIVE NOISE REDUCTION SCHEME**

4. SLM-BASED IMPULSIVE NOISE REDUCTION SCHEME

![Blanking Threshold vs Output SNR](image1.png)

**Figure 4.5**: The output SNR versus blanking threshold for different values of $U$ and $p$ when SINR = $-10$ dB and SNR$_G$ = 25 dB.

optimal blanking threshold is decreased and higher output SNR is achieved. The optimization of the blanking threshold of the SLM-based blanking system is investigated in more insight next.

### 4.6.2 Blanking Threshold Optimization

Here, an extensive search for the optimal blanking threshold is carried out under various PAPR and noise conditions. For given $p$, SINR and SNR$_G$ values the optimal blanking threshold is found by satisfying the following

$$T_{b,U}^{(opt)} = \arg \max_{0 \leq T_b < \infty} \{ \gamma^{(U)}_{SLM} (T_b, p, \text{SINR}, \text{SNR}_G) \}. \quad (4.17)$$

Figures 4.6a and 4.6b depict the optimal blanking threshold versus SINR for $p = 0.01$ and 0.001, respectively, with various values of $U$. It can be seen, generally, that when $U \lesssim 8$, the optimal threshold is larger for small pulse probabilities compared to that when $p$ is high. One common observation one
Figure 4.6: Optimal blanking threshold versus SINR for several values of $U$ and $p$ when $\text{SNR}_G = 25$ dB.

can clearly see for all $p$ values is that when $U$ is increased, the optimal threshold decreases. The intuitive explanation of this is that when $U$ increases, the useful signal energy will be contained within lower level and, hence, smaller blanking threshold will allow more effective blanking of the impulsive noise. Furthermore, it is interesting to note that when $U \lesssim 8$, the optimal threshold tends to be very large when the impulsive noise amplitude is either extremely low or extremely high. This is because in the first scenario when the SINR approaches zero, impulsive noise becomes comparable to the useful signal variance; as a result blanking does not provide any enhancement. On the other hand, for extremely high noise amplitudes, impulsive noise becomes easily identifiable and, therefore, large blanking threshold can still provide optimal performance. It is important to point out the fact that for large values of $U$, e.g. $U \geq 16$, the optimal threshold levels off, i.e it becomes independent of SINR, and this applies to all noise probabilities. Interestingly enough, when $U = 64$, the optimal threshold is almost equal for all the SINR and $p$ values which means that if an SLM-based OFDM system is deployed with a sufficiently large number of phase sequences, it will become possible to optimally
blank the impulsive noise independently of its characteristics. This phenomena is of great importance and will therefore be investigated thoroughly in Section 4.7.

### 4.6.3 Maximum Achievable Output SNR

The maximum achievable output SNR corresponding to the optimal blanking threshold found above is presented in Figure 4.7. It is seen that the proposed SLM-based system can always outperform the conventional blanking scheme and this enhancement is always proportional to $U$ irrespective of the noise probability. It is also evident that this improvement becomes more pronounced for low impulsive noise probabilities, e.g. $p = 0.001$. To highlight this gain, we have plotted the relative gain (4.18), versus SINR in Figure 4.8.

\[
G_r = 10 \log_{10} \left( \frac{\bar{\gamma}^{\{U\}}_{SLM} \left( T_b = T_{b,U}^{\{opt\}} \right)}{\bar{\gamma} \left( T_b = T_{b}^{\{opt\}} \right)} \right). \tag{4.18}
\]
Figure 4.8: Relative gain versus SINR for different values of $U$ and $p$ when $\text{SNR}_G = 25$ dB.

It can be seen from this figure that the largest gain is reached in the intermediate SINR region ($-5$ dB → $-15$ dB) where gains of up to 2.75 dB and 3.6 dB are achieved over the conventional blanking system when $U = 64$, for $p = 0.01$ and 0.001, respectively. It is worthwhile mentioning that even for a small number of phase sequences, e.g. $U = 2$, the proposed technique can still provide about 1 dB and 1.5 dB SNR enhancement when $p = 0.01$ and 0.001, respectively. Additionally, this gain becomes negligible in the low SINR region. It is important at this stage to stress the fact that recovering the side information of the SLM scheme is particularly crucial and in order to achieve best performance, such information must be protected by using powerful channel codes. This, consequently, not only will cause data rate loss but can also significantly increase the system complexity and transmission delay. Therefore, other PAPR reduction schemes which eliminate the requirement for side information such as the blind approaches [92, 99, 111–114], can be an attractive candidate in practice.
4.7 Blind Blanking Technique (Fixed Threshold)

Similar to the conventional blanking system and although it maintains better performance, the proposed SLM-based approach requires perfect estimation of noise parameters at the receiver in order to achieve best performance. However, the fulfillment of such an assumption can be difficult in practice. As presented previously, see Figure 4.6, the optimal blanking threshold reaches a plateau when $U$ is sufficiently large irrespective of the noise characteristics. For instance, when $U = 64$ the optimal threshold remains constant at about 2.9 for all the given pulse probabilities and SINR values.

Motivated by this, in this section we assign a predetermined and fixed blanking threshold value for the SLM-OFDM system such that $T_b = 2.9$ while $U = 64$. The output SNR of a such scenario is plotted versus SINR in Figure 4.9 for various pulse probabilities ranging from highly to weakly disturbed. This method is referred to as blind blanking. In order to provide comparative results, results of the conventional blanking system are also included. It is clearly seen that the blind blanking technique always outperforms the
conventional one in addition to the fact that no previous knowledge about the noise is required at the receiver to achieve best performance. To make this enhancement clearer, the relative gain obtained by this technique over the conventional one is presented in Figure 4.10. It is evident that the gain can be as high as 3.5 dB in a weakly disturbed impulsive noise environment \((p = 0.001)\) and about 1.25 dB in a heavily disturbed environment \((p = 0.1)\).

Therefore, it can be summarized that the blind blanking technique has two advantageous properties. Firstly and unlike the existing techniques, it does not require any noise estimations to combat the impulsive noise and hence avoids estimation errors and reduces the receiver complexity. Secondly, a better performance is obtained relative to the conventional blanking system. However, it is important to point out that these advantages are achieved at the expense of increased complexity in the transmitter. In order to reduce this complexity, other modulation schemes which have inherently low PAPR can be utilized in conjunction with nonlinear preprocessing at the receiver such as single-carrier frequency-division multiplexing access (SC-FDMA)⁴ [117–121].

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⁴In this respect, we have also examined the performance of two SC-FDMA systems:Localized FDMA (LFDMA) and Interleaved FDMA (IFDMA), see [115,116].
4.8 Chapter Summary

In this chapter we introduced and elucidated a new method to improve the conventional blanking process in OFDM-based PLC systems by employing a PAPR reduction scheme, namely, SLM. Closed-form expressions for the probability of blanking error were derived for both the conventional and SLM-based blanking systems. Results clearly demonstrated the robustness and superiority of the proposed system in the form of minimized probability of blanking error and significantly increased output SNR. Furthermore, it was found that when \( U \geq 64 \), it becomes feasible to optimally blank impulsive noise without the need for any prior knowledge about its characteristics. This improvement, however, would be achieved at the expense of some increased computational complexity at the transmitter.
Chapter 5

Dynamic Peak-Based Threshold Estimation Method

In this chapter we propose a novel impulsive noise mitigation technique that is based on utilizing the transmitted signal peak estimates; this method is referred to as dynamic peak-based threshold estimation (DPTE). To accomplish this, we also introduce a look-up table (LUT) based algorithm with uniform quantization. In addition, it will be shown that implementing the partial transmit sequence (PTS) scheme at the transmitter will make the performance of the proposed system more superior. The DPTE system not only eliminates the need for any noise estimations, but it is also found to provide better output SNR performance compared to the optimized conventional systems1.

5.1 Introduction

Determining the optimal threshold of a blanking nonlinear preprocessor is the key for achieving the best performance. This however requires accurate knowledge of the noise statistics. Even though most studies on this topic assume that the long-term impulsive noise characteristics can be made available at the receiver, short-term variations may still lead to misestimating the optimal threshold and hence causing significant performance degradation. In

1The results in this chapter have mostly been presented in [78, 122, 123].
this chapter, we will first assess the impact of short-term variations of impulsive noise on the conventional blanking method and then propose a different criterion for estimating the optimal threshold independently of the noise characteristics. In this scheme, a one-to-one direct relationship between the optimal threshold and the PAPR of the OFDM symbols is found and then utilized for enhancing the performance of the blanking-based technique. This implies that if the OFDM signal peaks can be accurately determined at the receiver, this will allow perfect noise cancellation at the receiver without any noise estimation requirement.

The question that arises here is, however, how can the signal peaks be determined for every single OFDM symbol at the receiver. To achieve this, we exploit a LUT-based algorithm with uniform quantization, and to distinguish it from the DPTE method, it will be referred to as DPTE-LUT. In the DPTE-LUT system, the OFDM symbol peaks are first quantized and then the corresponding bits are transmitted to the receiver as side information. The other contribution of this chapter resides in further improving the capability of the DPTE scheme by simply implementing the PTS scheme at the transmitter side.

5.2 Performance Loss Due to Estimation Errors

In this section, we investigate the effect of using the non-optimal blanking threshold on the output SNR for the conventional blanking method. It is worthwhile highlighting that estimation errors of the impulsive noise parameters could arise due to short-term channel time variations as well as noisy estimates. Figure 5.1 shows the loss in the output SNR due to misestimation of the optimal threshold for different values of \( p \). This loss is calculated as 

\[
10 \log_{10} \left( \frac{\gamma_{cb}}{\gamma_{\tilde{cb}}} \right),
\]

where \( \gamma_{cb} \) is obtained from (2.45) and (2.46) when replacing \( T_b \) with \( T_b^{\{opt\}} \) given in (2.47) whereas \( \gamma_{\tilde{cb}} \) is obtained by replacing \( T_b \) with \( \left( T_b^{\{opt\}} + e T_b^{\{opt\}} \right) \) where \( e \) denotes the estimation error.

It is clear from these results that as the deviation from the exact optimal threshold increases, the loss in the output SNR becomes greater and more so for negative errors. For instance, for \( p = 0.01 \) and an error of \(+35\%\) from the exact optimal threshold value, the worst case scenario will be a loss of about 1.7 dB at SINR = −15 dB, whereas for an error of \(-35\%\) the highest
Figure 5.1: The loss in the output SNR due to positive and negative errors when $\text{SNR}_G = 25$ dB.
loss will be about 6.2 dB at SINR = −16 dB. The intuitive explanation of this phenomena is that as the blanking threshold goes below the optimal value, more of the useful signal energy will be blanked whereas when it is above, less signal energy is blanked; this phenomena can also be extracted from the results in Figure 2.15 where it is obvious that the output SNR curves slope is higher (more sensitive to errors) in the region below the optimal threshold than that of above the optimal value.

Also, there is always more loss in the intermediate SINR region which clearly states that this region is the most sensitive to estimation errors. On the other hand, for extremely low SINR values, the effect of the deviation from the optimal threshold becomes less serious especially for the positive errors.

5.3 DPTE Method

In this system, the information bits are first mapped into 16-QAM base band symbols, $S_k$, which are then passed through an IDFT modulator to produce a time domain signal, $s_k$. The PAPR of the transmitted signal is given by (4.2). However, since we assume here that the signal power is normalized to unity, the PAPR simply indicates the peak value of the signal, i.e. $P = 10 \log_{10}(\max |s_k|)$. The noise in this investigation is modeled using the two-component Gaussian model (2.28). Therefore, under perfect synchronization condition, the received signal can be expressed as

$$r_k = s_k + n_{g,k} + b n_{i,k} \quad k = 0, 1, 2, \ldots, N - 1 \quad (5.1)$$

where $s_k$, $n_{g,k}$ and $n_{i,k}$ are assumed to be mutually independent. At the receiver, before the OFDM demodulator, blanking is applied. The basic principle of this blanking device is based on the OFDM symbol peak estimates as follows

$$y_k = \begin{cases} r_k, & |r_k| \leq P \\ 0, & |r_k| > P \end{cases} \quad k = 0, 1, \ldots, N - 1 \quad (5.2)$$

where $P$ is the exact peak value of the associated OFDM symbol.
5.3.1 DPTE Algorithm Development

It is intuitive to think that there is a direct relationship between the optimal threshold and the OFDM peak signal values because anything that exceeds the signal peak signifies unwanted noise or interference. To evaluate this relationship we conducted an extensive search for the optimal threshold for signals with different PAPR values based on the flowchart illustrated in Figure 5.2. The number of OFDM symbols considered here is \( n = 10^6 \) symbols with \( N = 256 \) sub-carriers. An OFDM symbol is first generated \( \{s(j)\} \) and its peak value is calculated \( \text{Peak}(j) \) before it is passed through the PLC channel where the noise vector \( \{n(j)\} \) is added to it to produce \( \{r(j)\} \). \( \{s(j)\}, \{n(j)\} \) and \( \{r(j)\} \) are vectors each of length 256 and \( j = 0, 1, \ldots, n \). Blanking is applied on the \( j^{th} \) received symbol with blanking threshold \( T_b \) being varied from 0 to 10 using the index \( \{v\} \) in step of 0.01 and the corresponding SNR \( (v) \) is determined. After that the blanking threshold \( T_b(v) \) that maximizes the output SNR of the \( j^{th} \) OFDM symbol is assigned to \( T_{opt}(j) \). This procedure is repeated for all the symbols and finally the vector \( T_{opt} \) is plotted versus the vector Peak in Figure 5.3 in the absence and presence of AWGN.

It is clear that there exists a one-to-one linear relation between the signal peaks and the optimal blanking threshold. This implies that if the peak of every individual OFDM symbol can be determined accurately at the receiver, it will be possible to optimally blank impulsive noise on symbol by symbol basis without the need to know the impulsive noise characteristics. It is also apparent from these figures that the optimal threshold is equal to the peak signal value.

5.3.2 Simulation Results

Computer simulations are conducted in this section to analyze the achievable gain in terms of SNR at the output of the blanker, in the presence of impulsive noise. These simulations are based on an OFDM system consisting of \( N = 256 \) sub-carriers with 16-QAM modulation. It is assumed that the transmitter and receiver are synchronized. The number of OFDM symbols considered in our study is \( 10^6 \), the OFDM signal power is normalized to unity \( \sigma_s^2 = (1/2) E[|s_k|^2] = 1 \) and the noise is generated using the two-component Gaussian noise model. Various pulse probabilities are considered, \( p = 0.1 \),
Figure 5.2: The flowchart for finding the relationship between the optimal blanking threshold and signal peaks.
Figure 5.3: The relationship between the optimal blanking threshold and the OFDM symbol peaks with AWGN, without AWGN and with AWGN and flat fading channel.
0.03, 0.01 and 0.003 which implies that 10%, 3%, 1% and 0.3% of the received OFDM samples will be affected by impulsive noise, respectively. Therefore, in this system with 256 sub-carriers, the average number of noise pulses received within each OFDM symbol will be about 26, 8, 3 and 1 pulses per OFDM symbol when $p = 0.1, 0.03, 0.01$ and 0.003, respectively. It is also worthwhile mentioning that the positions of impulsive noise pulses vary randomly within the OFDM symbols.

To establish the lower bound performance of this method we assume that the peak of each OFDM symbol is known precisely at the receiver. Under a such assumption, Figure 5.4 illustrates the output SNRs using both DPTE and conventional blanking methods for various probabilities. It can be observed that the DPTE method always outperforms the conventional blanking one and this gain increases as $p$ decreases. Notably, the highest gain for all the probabilities $p$ is always within the intermediate region of SINR. For instance, when $p = 0.003$, there is a gain of about 2.5 dB over using conventional blanking at $\text{SINR} = -10$ dB and about 0.8 dB for high impulse probability $p = 0.1$ at the same SINR value.
Figure 5.5: Block diagram of the proposed DPTE-LUT system.

5.4 Implementation of DPTE Method

The question that arises when implementing the DPTE method in practice is how can the signal peaks be determined for every single OFDM symbol at the receiver. In this section, we propose and implement a technique to accomplish this by exploiting a LUT-based algorithm with uniform quantization as presented in Figure 5.5. In this system, the OFDM symbol peaks are first quantized and then the corresponding bits are transmitted to the receiver as side information. At the receiver, the peak estimator will extract the peak value of the associated symbol and adjust the blanking threshold of the nonlinear device accordingly; hence

\[ y_k = \begin{cases} r_k, & |r_k| \leq \tilde{P} \\ 0, & |r_k| > \tilde{P} \end{cases} \quad k = 0, 1, \ldots, N - 1 \quad (5.3) \]

where \( \tilde{P} \) is the estimated OFDM symbol peak value.

5.4.1 PAPR Quantization

For better realization of the DPTE-LUT method, it is important to analyze the peaks distribution of the OFDM signal. Therefore, a bar-chart for the signal peaks distribution is plotted in Figure 5.6. This figure provides very useful information. For instance, it can be seen that 99.5\% of the symbol peaks, i.e. window size (WS = 99.5\%), are concentrated within the range from 2.2 to 4.3 whereas 80\% of the peaks lie in the range between 2.6 and 3.4. Such information is advantageous for determining the best trade-off between the
number of quantization bits, hence the size of the LUT, and WS.

The LUT size depends on the required accuracy of the signal peak estimate at the receiver. Practically, the symbol peak amplitudes can take on any value on a continuous range, following the probabilistic model in Figure 4.3 when $U = 1$, and therefore must be discretized into a finite number of quantized levels ($P_q$), where $q = \{1, 2, \ldots, N_q\}$, ranging from predetermined minimum and maximum values $P_{\min} = P_1$ and $P_{\max} = P_{N_q}$, respectively, see Figure 5.6. $N_q$ depends on the size of the LUT being used ($N_q = 2^{N_b}$), where $N_b$ is the number of bits representing each OFDM symbol peak. Since the DPTE-LUT technique exploits uniform quantization the resolution factor ($R_F$), i.e. the spacing between quantization levels, can be defined as

$$R_F = \frac{P_{\max} - P_{\min}}{N_q} \quad (5.4)$$

It should be pointed out that the smaller the $R_F$, the better the precision of the signal peak estimates achieved. It is important to ensure that signal peaks which fall between two quantization levels are assigned to the upper
level as this will minimize the possibility of not blanking the useful signal energy for that specific symbol. However, symbols with peaks larger than $P_{\text{max}}$ are mapped into $P_{\text{max}}$ and similarly all symbols having peaks below $P_{\text{min}}$ will be mapped into $P_{\text{min}}$. After that, the quantized peaks are represented by $N_b$ bits per OFDM symbol which are transmitted to the receiver as side information. Figure 5.7a depicts the exact ($P$) and quantized ($P_q$) signal peaks from which it is clear that as the LUT size increases, the resolution becomes higher and consequently the quantization error, $e_q = P - P_q$, is minimized as demonstrated in Figure 5.7b. This implies that more accurate estimation of the signal peaks can be obtained at the receiver, hence more accurate blanking threshold will be used resulting in a more efficient impulsive noise suppression.

5.4.2 Results and Discussion

This section looks into the impact of both the LUT size and quantization concentration, i.e. $WS$, on the DPTE-LUT method in terms of output SNR and SER. Simulation parameters here are $N = 64$ sub-carriers, 16-QAM modulation and $\text{SNR}_G = 30$ dB. For better clarity, it is important to define in advance the different DPTE scenarios considered here.

- Ideal DPTE: exact signal peaks are determined precisely at the receiver. This establishes the lower bound performance of DPTE method.

- Ideal DPTE-LUT: the quantized signal peaks are detected at the receiver error-free, i.e. assuming that the side information is not contaminated with noise.

- Practical DPTE-LUT: the side information is passed through the PLC channel and experiences impulsive noise impairments.

5.4.2.1 The Effect of LUT Size

We examine here the impact of the LUT size on the performance of both the ideal and practical DPTE-LUT techniques. System parameters considered here are $WS = 99.9\%$, $P_{\text{min}} = 2$ and $P_{\text{max}} = 5$, see Figure 5.6.

- Ideal DPTE-LUT
Figure 5.7: Uniform quantization and quantization error of the OFDM symbol peaks.
The output SNRs of the ideal DPTE-LUT is shown in Figures 5.8a and 5.8b for \( p = 0.001 \) and 0.1, respectively, with different LUT sizes. Results for ideal DPTE and conventional blanking are also included on these plots. As anticipated, it can be seen that as the LUT size increases the performance of the DPTE-LUT scheme becomes closer to that of the ideal DPTE system. It is also evident that for low impulsive noise probabilities, e.g. \( p = 0.001 \), the DPTE-LUT technique always outperforms the conventional blanking method irrespective of the LUT size. On the other hand, however, for heavily disturbed impulsive noise channel, \( p = 0.1 \), the importance of the LUT size becomes more significant. For instance, when a LUT size of only 2 bits is used, the proposed DPTE-LUT scheme slightly under-performs the conventional blanking method in the intermediate SINR region. This clearly states that higher resolution is required when \( p \) is relatively high. In addition, it is worth pointing out that for all impulsive noise scenarios a LUT of size 4 or 5 bits is sufficient to achieve a near-ideal performance. Clearly, for the ideal DPTE system gains of up to 3 dB and 0.8 dB can be attained when \( p = 0.001 \) and 0.1, respectively. This represents the highest achievable gain which the proposed system approaches with a LUT of size 4 or 5 bits. Furthermore, the SER performance corresponding to the output SNR results in Figures 5.8a and 5.8b is shown in Figures 5.8c and 5.8d, from which same trends can be observed.

- Practical DPTE-LUT

The realization of the proposed scheme requires transmitting the side information associated with each OFDM symbol peak over the PLC channel which may lead to receiving some of such information in error. We therefore investigate here the impact of the practical implementation on the DPTE-LUT system. Our investigations here will adopt a 4-bit LUT as such LUT size is found above to provide sufficiently accurate peak estimation. Figures 5.9a and 5.9b compare the output SNR for the ideal DPTE, ideal DPTE-LUT and practical DPTE-LUT techniques in addition to the conventional blanking system with various values of \( p \). As anticipated, the performance of the practical DPTE-LUT technique becomes closer to that of the ideal DPTE case as \( p \) becomes smaller. This is because when \( p \) is high, the side information is more likely to be detected in error resulting in using the inaccurate blanking threshold and,
CHAPTER 5. DYNAMIC PEAK THRESHOLD ESTIMATION METHOD

Figure 5.8: Maximum achievable output SNR and minimum achievable SER performance of the ideal DPTE, ideal DPTE-LUT and conventional blanking methods versus SINR for different values of $p$. 
therefore, causing inefficient impulsive noise reduction. Generally, the loss due to the practical impact of impulsive noise on the side information is insignificant. Hence it can be concluded that the proposed system is promising and can be reliably implemented in practice.

Similarly as in the previous section, the SER performance in correspondence to the output SNR curves, shown in Figures 5.9a and 5.9b, is illustrated in Figures 5.9c and 5.9d from which similar observations can be seen. It is worthwhile stressing the fact that the robustness of the DPTE-LUT scheme can be further enhanced by applying powerful coding techniques to make the side information more resistant to impulsive noise.

5.4.2.2 The Effect of Window Size

The concentration effect of quantization on the ideal DPTE-LUT scheme is assessed here in two different noise environments, weakly and heavily disturbed impulsive noise.

- Weakly Disturbed Impulsive Noise Environment

The output SNR of the DPTE-LUT technique is shown in Figure 5.10 for different WSs when $p = 0.01$ and $N_b = 2$ and 5 bits. To obtain comparative figures, results for ideal DPTE and conventional blanking are also presented on this plot. As can be seen, there is a general trend that as the WS increases, the DPTE-LUT system performance becomes closer to that of the ideal DPTE scheme. In general, for a given WS, the loss due to the limited WS is smaller for a system with a LUT of 5 bits in relation to that with a LUT of only 2 bits.

- Heavily Disturbed Impulsive Noise Environment

Similar procedure as in the previous section is followed here for a heavily disturbed environment, $p = 0.3$, and the output SNR performance of this scenario is demonstrated in Figure 5.11. It is interesting to note that for a small LUT size, Figure 5.11a, in the intermediate SINR region, $R_F$ becomes more crucial to the overall performance than WS. For instance, we can see that considering only 80.0% of the peaks spectrum, in the intermediate range of SINR, yields about 0.8 dB better SNR performance than that of a wider WS
Figure 5.9: Maximum achievable output SNR and minimum achievable SER performance of the ideal DPTE, ideal DPTE-LUT, practical DPTE-LUT and conventional blanking methods versus SINR for various values of $p$. 
Figure 5.10: Maximum achievable output SNR versus SINR when $N_b = 2$ and 5 bits in a weakly disturbed impulsive noise environment for different WS values.
Figure 5.11: Maximum achievable output SNR versus SINR for $N_b = 2$ and 5 bits in a heavily disturbed impulsive noise environment with several WS values.
such as 99.5% and 99.9%. The intuitive explanation of this is the fact that in this SINR region the amplitude of noise pulses are slightly higher than that of the signal peaks, and consequently any slight quantization error in the peak values will result in a severe effect on the output SNR. In contrast, when the LUT size is relatively big, e.g. $N_b = 5$ bits, Figure 5.11b, the performance is almost independent of the WS since quantization resolution becomes sufficiently high for all WSs. Therefore, under such conditions, the application of non-uniform quantization seems appropriate as the quantization intervals can be made smaller where the majority of peaks are concentrated; hence, higher quantization accuracy can be achieved.

### 5.5 Improved DPTE Method

The DPTE scheme can be further improved if the OFDM signal is preprocessed at the transmitter to maintain the average OFDM peaks below a certain threshold and this is done here by applying the PTS scheme [124]. This system will be referred to as DPTE-PTS, the block diagram of which is shown in Figure 5.12. This configuration will make impulsive noise more distinguishable at the receiver and hence minimize $P_b$ and $P_m$ while improving $P_s$ as will be presented later.

#### 5.5.1 System Model

In this system, the information bits are first mapped into 16-QAM symbols which are then grouped into vectors each of length $N$, denoted as $S_k$, where $k = 1, 2, \ldots, N$. $S_k$ is then partitioned into $M$ disjoint sub-blocks $S_k^{(m)} = [S_0^{(m)}, S_1^{(m)}, \ldots, S_{N-1}^{(m)}]$, $m = 1, 2, \ldots, M$, and all sub-carriers which are already represented in another sub-block are set to zero so that $S_k = \sum_{m=1}^{M} S_k^{(m)}$. Then the IDFT is employed for each sub-block to produce $s_k^{(m)} = \text{IDFT}\{S_k^{(m)}\}$. After that, each sub-block is multiplied by a different phase weighting factor $b^{(m)}$. The peak value optimization block iteratively searches for the optimal combination of the phase weighting factors that offer the minimum PAPR. Once the optimal weighting factor is determined, all the sub-blocks are summed $\bar{s}_k = \sum_{m=1}^{M} b^{(m)} s_k^{(m)}$ and then transmitted. At the receiver’s front-end, blanking is applied as
Figure 5.12: Block diagram of an OFDM-PTS system with DPTE-based blanking at the receiver.
\[
y_k = \begin{cases} 
  r_k, & |r_k| \leq \bar{P} \\
  0, & |r_k| > \bar{P}
\end{cases} \quad k = 0, 1, \ldots, N - 1 \tag{5.5}
\]

where \( \bar{P} < P \) is the reduced OFDM symbol peak value when PTS is implemented. After the blanking device, \( y_k \) is passed through the DFT to produce \( Y_k = \text{DFT}\{y_k\} \) which is then partitioned into \( M \) disjoint sets \( \{Y_k^{(m)} : m = 0, 1, \ldots, N - 1\} \) and zero padding is performed such that \( Y_k = \sum_{m=1}^{M} Y_k^{(m)} \). Using the inverse phase weighting factors \( b^{(m)*} \), we get \( \bar{S}_k^{(m)} = b^{(m)*} Y_k^{(m)} \) and the signal after summing and parallel-to-serial (P/S) device can then be given as \( \bar{S}_k = \sum_{m=1}^{M} \bar{S}_k^{(m)} \).

### 5.5.2 CCDF and \( P_b \)

In the PTS scheme, a set of phase weighting factors is usually selected for generating the phase weighting sequences. Assuming that there are \( W \) phase weighting factors in this set, the optimal PAPR is found after checking \( W^{M-1} \) different combinations and the number of bits required to represent the side information is \( \log_2 (W^{M-1}) \). Compared to the conventional OFDM system which only requires one IDFT operation, in the PTS scheme \( M \) IDFT operations are performed. It should also be highlighted that the amount of PAPR reduction for this scheme depends on both \( M \) and \( W \). To illustrate this, we plot in Figure 5.13 the CCDF of the PAPR for this system when \( M = 2, 3, 4, \) and \( 5 \) and \( W \in \{\pm 1, \pm j\}^2 \). It is evident that the amount of PAPR reduction increases as \( M \) is increased and that at CCDF = 10^{-4}, there is a PAPR reduction of about 4.5 dB and 3.5 dB when \( M = 5 \) and \( M = 3 \), respectively. This reduction in the PAPR implies lower probability of blanking error as shown in Figure 5.14. The analytical CCDF and \( P_b \) results of the conventional OFDM scheme are obtained from (4.12) and (4.7), respectively. It is noticeable that the PTS-based system minimizes \( P_b \) compared to the conventional OFDM system and that \( P_b \) is inversely proportional to \( M \) which means that the system performance will improve for higher values of \( M \).

\(^2\)It is found in [124] that a restriction to four phase weighting factors can provide a significant peak reduction.
Figure 5.13: CCDF plot for the conventional and PTS-based OFDM systems with different values of $M$ when $W = 4$.

### 5.5.3 $P_m$ and $P_s$

We explore here $P_m$ and $P_s$ for the conventional DPTE\(^3\) and DPTE-PTS techniques. For the conventional system, $P_m$ and $P_s$ are determined, respectively, as

\[
P_{m}^{\text{Conv. DPTE}} = \Pr (A_r < P | \mathcal{H}_1) \Pr (\mathcal{H}_1),
\]

(5.6)

\[
P_{s}^{\text{Conv. DPTE}} = \Pr (A_r > P | \mathcal{H}_1) \Pr (\mathcal{H}_1),
\]

(5.7)

whereas for the DPTE-PTS scheme, $P_m$ and $P_s$ are calculated as follows

\[
P_{m}^{\text{DPTE-PTS}} = \Pr (A_r < \bar{P} | \mathcal{H}_1) \Pr (\mathcal{H}_1),
\]

(5.8)

\(^3\)Conventional DPTE refers to the DPTE system without PAPR reduction which is eventually the ideal DPTE method.
CHAPTER 5. DYNAMIC PEAK THRESHOLD ESTIMATION METHOD

Figure 5.14: Probability of blanking error versus blanking threshold for different values of $M$ when $W = 4$ and $\text{SNR}_G = 30$ dB.

$P_s^{\text{(DPTE-PTS)}} = \Pr (A_r > \bar{P} \mid \mathcal{H}_1) \Pr (\mathcal{H}_1), \quad (5.9)$

where $P$ and $\bar{P}$ are the peak values of the associated OFDM symbols for the conventional DPTE and DPTE-PTS systems, respectively. Figure 5.15 depicts some simulation results of (5.6)-(5.9) as a function of SINR when $\text{SNR}_G = 30$ dB with several values of $M$. It is clear that the DPTE-PTS system offers the best performance which improves as $M$ increases. On the other hand, the optimized conventional blanking system has the worst performance.

### 5.5.4 Output SNR and SER Performance

For more quantitative characterization of the system performance, we now analyze the maximum achievable output SNR and minimum achievable SER performance. Figure 5.16 shows these parameters versus SINR for the DPTE-PTS, conventional DPTE and optimized conventional blanking systems with
Figure 5.15: Probability of miss and probability of success versus SINR for the DPTE-PTS, conventional DPTE and optimized conventional blanking systems with various values of $M$ when $W = 4$. 
Figure 5.16: Maximum achievable output SNR and minimum achievable SER performance versus SINR for the DPTE-PTS, conventional DPTE and optimized conventional blanking systems with different values of $M$ when $p = 0.01$. 
different values of $M$ when $p = 0.01$. Notably, the proposed DPTE-PTS technique always outperforms the other two systems for all impulsive noise probabilities and, as anticipated, this enhancement increases as $M$ becomes larger and $p$ becomes smaller.

5.6 Chapter Summary

This chapter began with investigating the impact of estimation errors on the performance of the conventional blanking scheme. After that, to avoid these estimation errors, a new blanking technique, which does not rely on noise characteristics, referred to as DPTE scheme, was introduced. Also, for implementing the DPTE method, we proposed a LUT-based algorithm with uniform quantization. All in all, results reveal that

- The conventional blanking method may lead to a dramatic degradation in the output SNR if the impulsive noise characteristics can not be accurately obtained at the receiver.

- There is a direct relationship between the peaks of OFDM symbols and the optimal blanking threshold.

- DPTE method is not only independent of impulsive noise measurements, but can also offer an output SNR gain of about 2.5 dB if the signal peaks can be precisely determined at the receiver.

- Better performance is achieved as the LUT size is increased and, in general, a LUT size of 32 is sufficient to achieve near-ideal DPTE performance. Moreover, in heavily disturbed environments, more quantization accuracy is required specially in the intermediate SINR region.

- Finally, implementing a PTS modulator at the transmitter is found to further improve the DPTE method.
Chapter 6

MC-CDMA Over PL Channels

Unlike the previous chapters that focused on OFDM-based PLC systems, this chapter explores the application of multi-carrier code division multiple access (MC-CDMA) over the multipath fading PL channel contaminated with Middleton class-A noise\(^1\). The proposed MC-CDMA system is implemented with a minimum mean-square error (MMSE) equalizer and nonlinear preprocessing to overcome the effects of bursty noise and multipath frequency-selective fading\(^2\).

6.1 Introduction

While both MC-CDMA and OFDM transmit symbols in parallel, they differ in the sense that the latter transmits different symbols over the available subcarriers whereas in the former each symbol is spread over the entire number of sub-carriers. The motivation for using CDMA for PLC resides in its flexibility as a multiple access and resilience to impulsive noise, as well as its lower PAPR features, compared to OFDM, for certain types of sequences. To realize MC-CDMA, [128–131], many classes of spreading sequences can be utilized such as Pseudonoise (PN), Gold, Kasami, Walsh- Hadamard (WH) and orthogonal poly-phase (OPP) sequences, which have different properties particularly with respect to auto-correlation and cross-correlation functions.

As for this thesis, for conciseness, but without loss of generality, we will

\(^1\)For DS-CDMA systems, you may refer to [125].

\(^2\)Some of the results in this chapter are presented in [126, 127].
focus generally on the PN, WH and OPP sequences. The rationale for selecting these codes is, firstly, because they are commonly used in CDMA-based systems. Secondly, such codes represent the majority of spreading sequences which are mainly categorized into orthogonal and non-orthogonal codes \[132\]; in addition to their auto-correlation and cross-correlation properties. For example, PN codes \[133\], have Gaussian-like auto-correlation function, which is an advantage, but exhibit poor cross-correlation between the codes. In contrast, binary WH codes, \[134\], have zero cross-correlation, in the case of perfect synchronization, making it very attractive for the down-link transmission. However, the resultant large peak power when the number of users is very small and the limited number of available codes constrained by the code length, as well as their sensitivity to time-misalignments, remain the main drawbacks of WH sequences \[135,136\].

As for OPP sequences, they are the non-binary extension of the binary WH sequences. As well as having similar cross-correlation properties as the binary WH, their advantages include firstly, robustness against timing misalignment which can significantly degrade the performance of the binary WH codes. Secondly, unlike the WH codes, OPP codes are not limited by the sequence length hence can accommodate much more users/data. Thirdly, but most importantly, such codes tend to have lower PAPR properties as the constellation size of these codes is increased. The utmost importance of this property, as shown previously, resides in the fact that minimizing the PAPR of the transmitted signal will ultimately flatten the envelop of the signal, hence allowing more efficient noise cancellation. Motivated by these features, this chapter will particularly focus on MC-CDMA with OPP codes. It should also be mentioned that an MMSE equalizer is employed at the receiver to compensate for the inter-symbol interference (ISI) caused by the frequency-selective channel without increasing the noise power.

The main contributions of this chapter are as follows. First, it investigates the PAPR performance of PN, WH, and OPP based MC-CDMA systems under full-loading and half-loading system scenarios to establish the critical relationship between the system loading, PAPR of the transmitted signal and OPP code constellation size. The second contribution resides in evaluating the SNR value of the MC-CDMA-OPP system at the output of the most three popular nonlinear preprocessors for reducing the impact of bursty noise in multipath
fading. Finally, the problem of blanking and clipping thresholds optimization is addressed in different noise environments and the corresponding maximum achievable output SNR and minimum SER performances are evaluated. Although this chapter is mainly dedicated to analyze the performance of MC-CDMA, for the purpose of comparison, OFDM-based results are also included throughout this chapter.

### 6.2 System Model

The system model under consideration in this chapter is presented in Figure 6.1. First, the 16-QAM symbol of each user, $S_m$, is spread using the user-specific code $c_m = \left[ c_m^{(0)}, c_m^{(1)}, ..., c_m^{(N-1)} \right]$, where $N$ denotes the code length, $m = [1, 2, ..., M]$ and $M$ is the total number of users. Throughout this chapter, unless explicitly stated otherwise, we use a fully loaded system (i.e. $M = 64$ users). After that, the spread signals are multiplexed to produce $d = [d_0, d_1, ..., d_{N-1}]$ which is then passed through a serial-to-parallel (S/P) converter. The S/P output is fed to an IDFT, the size of which is equal to the code length, then applied to a P/S converter and a CP is added before transmission. The transmitted signal for one MC-CDMA block is expressed mathematically as

$$ s(t) = \sum_{k=0}^{N-1} \sum_{m=0}^{M-1} S_m c_m^{(i)} \exp \left( \frac{j2\pi kt}{T_s} \right) $$

(6.1)

where $T_s$ is the MC-CDMA symbol duration. Given (6.1), the PAPR of the MC-CDMA transmitted signal can also be calculated using (4.2). After $s(t)$ is passed through the multipath PLC channel, which is assumed to follow Rayleigh fading distribution, the received signal will have the following form

$$ r_k = h_k * s_k + n_{t,k}, \quad k = 0, 1, ..., N - 1 $$

(6.2)

where $s_k = s (kT_s/N)$. In this study, we adopt a standard Rayleigh multipath fading channel for both MC-CDMA and OFDM systems and assume that the number of channel taps is lower than the CP duration to avoid the ISI
Figure 6.1: Block diagram of the MC-CDMA system over the multipath PLC channel with nonlinear preprocessing at the receiver.
problem, i.e. orthogonality is maintained by the CP$^3$.

Middleton class-A noise model is used here to emulate the total noise characteristics (2.19). In order to diminish the deleterious impact of impulsive noise, the received signal is passed through a blanking (2.42), clipping (2.43) or adaptive hybrid nonlinear preprocessor (3.1). After the nonlinear device, the next stage is the CP removal after which the resultant signal, $z_k$, is fed to the S/P converter and then to an $\bar{M}$-point DFT to produce $G = [G_0, G_1, \ldots, G_{\bar{N}-1}]$. In order to compensate the channel distortion, equalization is performed using the MMSE equalizer as [132]

$$\tilde{G}(k) = \frac{H^*(k)}{H(k)H^*(k) + \gamma^{-1}}$$  \hspace{1cm} (6.3)

where $\gamma$ represents the SNR at the output of the nonlinear preprocessor and $H(k)$ is the channel frequency response. The output of the MMSE equalizer is now passed through a P/S converter and then multiplied by the spreading codes to produce estimates for the data symbols of the different users, $\tilde{S}_m$. Finally, 16-QAM demodulation takes place and the SER can be calculated.

### 6.3 CCDF of Various MC-CDMA Systems

In this section we examine the impact of the different spreading codes under consideration on the PAPR performance. Figure 6.2 depicts some simulation results for the CCDF of the MC-CDMA-PN, MC-CDMA-WH and MC-CDMA-OPP systems for full-loading (64 users) and half-loading (32 users) scenarios$^4$. For full-loading, it is noticeable that MC-CDMA-WH and MC-CDMA-OPP have considerably lower PAPR compared to OFDM while the OPP-based system offers the best performance. In addition, it can be deduced that MC-CDMA-PN exhibits similar performance as OFDM which could be a result of the high cross-correlation property of such codes. Furthermore, as the number of OPP codes phases ($\phi$) is increased, the performance enhances providing a gain of up to $2.5 \text{ dB}$ when $\phi = 8$, compared to OFDM, at CCDF $= 10^{-3}$. To highlight this, we plot in Figure 6.3 the PAPR performance of the

---

$^3$Perfect channel estimation is assumed and hence the results will serve as an upper bound for the performance.

$^4$The reason why we consider here different loadings is because in practice such scenarios can take place since not all the users are always active, i.e. continuously transmitting.
MC-CDMA-OPP signal against $\phi$ for the two loading scenarios when CCDF $= 10^{-1}, 10^{-2}$ and $10^{-3}$. The first observation one can see is that the PAPR value is reduced with increasing $\phi$. Comparing the two loadings, clearly full-loading always has lower PAPR with respect to half-loading at the same CCDF and $\phi$ values. It is also apparent that the amount of PAPR reduction is more significant in full-loading than that with half-loading. For example, in the former case at CCDF $= 10^{-3}$, a PAPR reduction of about 6 dB can be achieved when $\phi = 64$ relative to the system with $\phi = 2$ whereas only about 4.5 dB is attained from the half-loading scenario at the same CCDF value. This enhancement can be intuitively justified by the increase in the phase randomization across the sub-carriers which is related to the minimum-distance decoding of the sequences [89]. Moreover, it is noteworthy to mention that the PAPR tends to level off for very large constellation sizes of OPP codes. Although in this section we looked into both half-loading and full-loading scenarios, in the rest of this chapter, the focus will be primarily on the latter not only because it relatively has better PAPR properties but also because
full-loading is a more common configuration in practice.

### 6.4 Preprocessor Threshold Impact on Output SNR

In this section we assess the performance of MC-CDMA-OPP with different constellation sizes in terms of the output SNR. Our investigations here are based on: $\bar{M} = 64$ users, $\sigma_u = 0.05$, $A = 0.0025$ and $\Gamma = 0.001$ which means that impulsive noise is 1000 times, or 30 dB, greater than the background level. Figure 6.4 presents the output SNR of the proposed system with different OPP codes phases $\{\phi = 2, 4, 8, 64\}$ in a multipath fading channel. In addition, the performance of OFDM-based schemes is included with and without multipath fading. Note that for fair comparisons same channel is used for both MC-CDMA and OFDM systems. The analytical results of the OFDM system W/O multipath for the blanking, clipping and adaptive hybrid techniques are found using (2.48)-(2.52), (3.2) and (3.3).

As a first common observation on the results shown in Figure 6.4, it is
Figure 6.4: Output SNR as a function of the thresholds, for MC-CDMA-OPP with several values of $\phi$, and OFDM in the presence of Middleton class-A noise. MC-CDMA-OPP results are obtained for a multipath channel. Noise parameters: $\sigma_u = 0.05$, $\Gamma = 0.001$ and $A = 0.0025$. 
obvious that, in a multipath channel and irrespective of the nonlinear device utilized, the performance of the MC-CDMA-OPP system always outperforms the OFDM approach even with a small number of phases, e.g. $\phi = 2$, and this improvement becomes higher as the code constellation size is increased. Clearly, the proposed system with $\phi = 64$ can offer output SNR gains of up to 4 dB, 2 dB and 4 dB relative to the OFDM-based one when blanking, clipping and adaptive hybrid are employed, respectively. Interestingly enough, however, when the constellation size of OPP codes is sufficiently large, e.g. $\phi = 64$, MC-CDMA-OPP can always achieve exact performance as OFDM W/O multipath. Notably, in both OFDM and MC-CDMA-OPP, the adaptive hybrid scheme serves as an upper bound for the performance. As a final remark on these results, it is interesting to observe that for every phase value there is always an optimal blanking/clipping threshold that maximizes the output SNR.

### 6.5 Performance Optimization

We now conduct extensive computer simulations to find the maximum achievable output SNR and the minimum achievable SER performance for the MC-CDMA-OPP and OFDM systems. The rest of our results will only consider blanking and clipping, not only because adaptive hybrid was shown to offer insignificant improvement compared to the other two nonlinear devices, but also because it is more complex to achieve optimal performance as it requires optimizing two parameters ($T$ and $\alpha$), instead of one.

#### 6.5.1 Fixed $\Gamma$

In this section, we examine the optimized system performance when $\sigma_n^2$ is varied while fixing $\Gamma$. With this in mind, the maximum achievable output SNR is plotted versus $10 \log_{10} (\sigma_n^2)$ in Figures 6.5a and 6.5b for MC-CDMA-OPP and OFDM systems with blanking and clipping, respectively. Comparing these figures, a few common observations can be seen. Firstly, the typical receiver always has the worst performance followed by the OFDM system with multipath. Secondly, the MC-CDMA-OPP approach always outperforms the OFDM-based schemes even with a small value of $\phi$ and this enhancement
Figure 6.5: Maximum achievable output SNR versus $10 \log_{10} (\sigma_u^2)$ for the MC-CDMA and OFDM systems in a multipath channel for various codes phases with blanking and clipping. Noise parameters: $A = 0.0025$ and $\Gamma = 0.001$. OFDM with no multipath: (dotted line → analytical) and (◦ → simulation).
becomes higher as \( \phi \) is increased. The third common trend, and the most interesting, is that the proposed system with \( \phi = 64 \) approaches the performance of OFDM W/O multipath over the entire noise spectrum. In general, blanking always yields better performance than clipping under same noise and modulation features.

It is also clear that when \( \sigma_u^2 \) is very low, around 25 dB below the transmitted signal level, the performance is very good for both blanking and clipping which is intuitive. At the other extreme, however, when \( \sigma_u^2 \) is very high, around 10 dB above the signal level, performance deteriorates dramatically. This is because in this noise region, the system becomes Gaussian limited making the nonlinear preprocessing-based schemes inefficient. However, in the intermediate \( \sigma_u^2 \) region: \(-25 \leq \sigma_u^2 \leq 10 \) dB, for the blanking-based systems the output SNRs degrade as \( \sigma_u^2 \) increases until it reaches about \(-15 \) dB after which performance begins to improve reaching its peak at around \( \sigma_u^2 = 0 \) dB with an output SNR of 24 dB before it declines rapidly again. On the other hand, for clipping, the performance consistently decreases as \( \sigma_u^2 \) increases over the entire intermediate region of \( \sigma_u^2 \).

For more quantitative characterization of the proposed system, the SER performance in correspondence to the output SNR results discussed above is also found and presented in Figure 6.6 from which similar observations can be seen. Notably, blanking has generally lower SER performance but when \( \sigma_u^2 \) is extremely high, SER is affected badly for all considered systems indicating that the nonlinear devices are no longer efficient and, consequently, more sophisticated techniques should be deployed instead.

### 6.5.2 Varying \( \Gamma \)

We now keep \( \sigma_g^2 \) fixed such that \( \text{SNR}_G = 10 \log_{10} \left( \frac{\sigma_s^2}{\sigma_g^2} \right) = 30 \) dB while varying \( \sigma_u^2 \). Under these conditions, the output SNRs of the optimized MC-CDMA-OPP and OFDM systems versus \( \Lambda = 1/\Gamma \) are plotted in Figure 6.7 with different values of \( \phi \). Similar trends as in Figure 6.5 can be seen except the fact that when \( \Lambda \) is very large, meaning that \( \sigma_u^2 \) is also very large as shown on the top \( x \) axes, the output SNR does not decline as in the previous section. This is owing to the fact that the system is no longer Gaussian limited (since \( \sigma_g^2 \) is kept fixed) and impulsive noise becomes the dominant parameter in this region. Therefore, when impulsive noise is very large it becomes more
Figure 6.6: Minimum achievable SER performance corresponding to the maximum achievable output SNR versus $10 \log_{10} (\sigma_u^2)$, for both MC-CDMA with different values of $\phi$, and OFDM systems in a multipath channel for various codes phases with blanking and clipping. Noise parameters: $A = 0.0025$ and $\Gamma = 0.001$. 
Figure 6.7: Maximum achievable output SNR against $\Lambda$ for both MC-CDMA with various values of $\phi$ and OFDM systems in a multipath channel with blanking and clipping. Noise parameters: $A = 0.0025$ and $\text{SNR}_G = 30$ dB. OFDM with no multipath: (dotted line $\rightarrow$ analytical) and ($\circ$ $\rightarrow$ simulation).
distinguishable from the useful transmitted signal and, consequently, more efficient banking and clipping process is accomplished. The corresponding SER performance is illustrated in Figure 6.8 and similar trends can be noticed.

6.6 Impulsive Index Impact

To illustrate the effect of impulsive index on the performance of the optimized system, we set the noise power as \( \sigma_u^2 = 0.05 \) and plot the output SNR of the MC-CDMA-OPP system with blanking and clipping versus \( \phi \) for \( A = 0.001, 0.003, 0.005, 0.01 \) and 0.05 as shown in Figure 6.9a. It is clearly visible that the output SNR is inversely proportional to \( A \) but, as expected, directly proportional to \( \phi \). For instance, when \( A = 0.001 \), gains of up to 3 dB and 2 dB are attained when \( \phi = 64 \) relative to the case when \( \phi = 2 \) for the blanking and clipping, respectively; whereas when \( A = 0.005 \) these gains are reduced to around 2.5 dB and 1.5 dB for the same features. It is also obvious that the OPP blanking-based scheme performs better than the clipping one when \( A \) is low, e.g. \( A = 0.001, 0.003 \) and 0.005, whereas clipping offers slightly better performance when \( A \) is relatively high, e.g. \( A = 0.01 \) and 0.05. Moreover, in a heavily disturbed environment, \( A = 0.05 \), the output SNR becomes almost independent of \( \phi \).

For better visualization of the achievable gain over the OFDM-based scheme, we plot in Figure 6.9b the gain in the output SNR over the OFDM system versus \( \phi \) for several values of \( A \). It is evident that when \( A = 0.001 \), gains of up to 5.25 dB and 2.25 dB can be achieved when \( \phi = 64 \) for the blanking and clipping scenarios, respectively. In a heavily disturbed impulsive noise environment, the gain becomes negligible for clipping and zero for blanking indicating that under such impulsive noise conditions OFDM should be implemented instead since it is less complex.

6.7 Chapter Summary

In this chapter we looked into the performance of MC-CDMA approach over frequency-selective PL channels contaminated with Middleton class-A noise combined with MMSE and nonlinear preprocessing at the receive. Three
Figure 6.8: Minimum achievable SER performance corresponding to the maximum achievable output SNR against $\Lambda$ for both MC-CDMA with different values of $\phi$ and OFDM systems in a multipath channel with blanking and clipping. Noise parameters: $A = 0.0025$ and $\text{SNR}_G = 30$ dB.
Figure 6.9: Maximum achievable output SNR and SNR gain over the OFDM system versus $\phi$ with different values of $A$ for both blanking and clipping in a multipath channel. Noise parameters: $\sigma_u = 0.05$ and $\Gamma = 0.001$. 
spreading sequences, namely: PN, WH and OPP codes were investigated under various system loadings with more focus on OPP codes. Three burst noise reduction schemes were considered, namely, blanking, clipping and adaptive hybrid. In comparison to the OFDM-based scheme, it was found that some MC-CDMA-OPP systems are more effective in tackling the PLC channel impairments providing output SNR gains of up to 5 dB. It was also demonstrated that the performance of MC-CDMA-OPP scheme improves as we increase the constellation size of the OPP codes due to the inverse relationship between the constellation size and the signals’ PAPR.
Chapter 7

Conclusions and Future Work

7.1 Conclusions

This thesis highlighted the fundamentals of PLC and the challenges facing this technology with a focus on impulsive noise cancellation and security issues over such channels. Chapter 1 briefly discussed the PLC impairments such as multipath propagation, frequency-dependent attenuation, electromagnetic regulations and the non-Gaussian noise. Then, the literature review in Chapter 2 presented the main PLC channel characterization and noise modelling schemes. In addition, this chapter compared the performance of single-carrier and multi-carrier modulation systems over the impulsive noise PLC channel; it was shown that the latter can generally perform better than the former. Moreover, some impulsive noise mitigation techniques reported in the literature such as blanking, clipping and hybrid nonlinear preprocessors, were discussed and compared.

In Chapter 3, a new method was proposed to enhance the capability of the conventional hybrid nonlinear preprocessor by jointly, and adaptively, optimizing the threshold and the proportionality constant in such a way to maximize the output SNR performance. This system was referred to as adaptive hybrid. As such, a closed-form analytical expression for the output SNR was formulated and then the problem of threshold and proportionality constant optimization was addressed. Other aspects of the achievable performance were also analyzed such as the probability of miss, probability of success and SER performance. The obtained results demonstrated that the adaptive hybrid technique is able to yield up to 0.6 dB output SNR improvement relative
to the conventional hybrid approach and can considerably minimize the probability of miss, maximize the probability of success as well as improving the SER performance. This improvement is however accomplished at the cost of higher complexity at the receiver since two parameters are now optimized instead of one.

Chapter 4, unlike the aforementioned systems which mainly counter impulsive noise at the receiver, proposed to preprocess the OFDM signal at the transmitter in such a way to improve the noise cancellation process at the receiver. This was done by deploying an SLM modulator at the transmitter. A closed-form expression for the probability of blanking error of this system was derived and validated with computer simulations. The system performance was evaluated with and without multipath fading. Results clearly demonstrated the robustness and superiority of the proposed SLM-based system in the form of minimized probability of blanking error and a significant increase in the output SNR. It was also found that choosing $U \geq 64$ makes it feasible to optimally cancel impulsive noise without the need for any prior knowledge about the noise statistics. In practice, this system however would require sending some side information which could lead to an overall reduction in the system capacity. Therefore, to avoid this, other blind PAPR reduction schemes can be instead used in practice.

Chapter 5 began with addressing the impact of noise statistics misestimations on the performance of conventional blanking systems. In particular, it was shown that conventional blanking may cause dramatic performance degradations if the noise parameters can not be accurately obtained at the receiver. To overcome this, we introduced a new novel blanking technique which is not only independent of the noise statistics, but can also provide about 2.5 dB output SNR gains over conventional blanking. This method only requires estimates of the OFDM symbol peaks at the receiver and was referred to as DPTE. Additionally, and in order to realise this method, we proposed a LUT-based algorithm with uniform quantization. In this regard, it was presented that a LUT size of 32 is sufficient to achieve near-ideal performance and that in heavily disturbed impulsive noise environments, quantization accuracy becomes more important. Next, DPTE was combined with PTS to further enhance its capability and results indicated that increasing the number of partitions of the PTS scheme will significantly enhance performance.
Unlike all the above chapters which considered OFDM, in Chapter 6, we investigated the performance of MC-CDMA over the multipath PL channel when contaminated with Middleton class-A noise. In this chapter, three spreading sequences were investigated, namely: PN, WH and OPP codes, under various system loading and noise scenarios. To reduce the impact of impulsive noise, we utilised the most common three nonlinear preprocessors. Results demonstrated that under full-loading condition, MC-CDMA with WH and OPP codes always outperforms the conventional OFDM-based scheme irrespective of the noise scenario. MC-CDMA-OPP was also shown to offer the best performance and this improvement becomes more pronounced as the constellation size of the OPP codes is increased. This is basically due to the inverse relationship between the constellation size and the signals’ PAPR.

7.2 Future Work

Here we suggest possible research extensions to the work done in this thesis.

1. In Chapter 4, we proposed to preprocess the OFDM signal at the transmitter to enhance the noise cancellation process at the receiver. We particularly derived a closed-form expression for the probability of blanking error and only simulation results were presented for the output SNR. However, it would be worthwhile deriving accurate mathematical expressions for the output SNR and SER performance for this system. Additional work could then be done by optimising some parameters such as the number of phase sequences.

2. It was also found that modulation techniques characterized by low PAPR can considerably improve communication performance over PLC channels when combined with a nonlinear preprocessor at the receiving end. Motivated by these results, we have recently investigated the performance of CE-OFDM which has the lowest achievable PAPR of 0 dB. Initial results of a such system were presented at the IEEE ISPLC’15 conference in Texas, see Appendix B. Although the organising committee of this conference selected this paper to receive the best student paper award, we believe that some other important aspects should still be
examined and this may include
- The effect of the multipath PLC channel on the BER/SER performance.
- The degradation caused by the sudden phase changes due to the non-linear process at the receiver and whether or not the overall gain obtained from the $0 \, \text{dB} \, \text{PAPR}$ property outweighs these losses.
- The impact of modulation index on the system performance etc.

3. In addition, PLC/wireless diversity combining has recently received considerable attention and this was briefly considered in this thesis in the context of improving the security of PLC systems. However, as far as some of our proposed noise mitigation techniques are concerned, PLC/wireless diversity can be exploited to ensure that the sensitive information, such as quantization bits in DPTE-LUT and side information in SLM-based systems, can be more reliably delivered to the receiving end. Interesting comparisons can then be made between the improvements attained with the PLC/wireless and PLC-alone systems.

4. Although most of the work in this thesis is based on either computer simulations or/and theoretical analysis, it would be interesting to carry out practical experiments to examine the performance of the various proposed techniques in this thesis, which will indicate how practical such systems can be in real-life scenarios.

5. Electromagnetic emissions from PLs is another potential area of research. It is agreed that for higher data-rates over PLs, higher bandwidth should be utilised. However, high-frequency transmissions over such cables implies not only higher attenuation but also more electromagnetic radiations since PLs tend to behave more as antennas at high frequencies. Therefore, to avoid interference with other wireless systems, the possible transmitted power is very restricted. Clearly, there exists a trade-off between several parameters. This motivates us to conduct extensive measurement campaigns to determine a practical relationship between the frequency, transmitted power, attenuation and EMC while maintaining a reliable communication performance. This could be examined for various environments such as flats, offices, labs etc.
Bibliography


Appendix A

Sample Codes

This appendix presents some samples of the MATLAB codes used to produce chosen results in this thesis.

A.1 CCDFs of Conventional & SLM-based OFDM

```matlab
clc
clear
% ------------- Analytical ---------
N = 64; % fft size
PAPRo = [0:0.1:80];
U = [ 1 2 4 8 16]; % Number of phase sequences (U)
for ii=1:length(U)
    CCDF_SLM_(ii,:) = ( 1 - (1-exp(-(PAPRo))).^(N) ).^(U(ii)); % eq (4.7)
end
CCDF_th = 10*log10(PAPRo);
% ------------ Simulation -----------
n = 10000; % Number of OFDM symbols
M = 16; % constellation size
v = sqrt(1/((2/3)*(M-1))); % normalizing factor
a = [1:sqrt(M)/2]; % alphabets
alphaMqam = [-2*a-1 2*a-1];
PAPRosim = [0:0.25:10];
for mm=1:length(U)
p = [1 -1 j -j]; % phase factor possible values
```

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B = randsrc(N,U(mm),p); % generate N-point phase factors for each OFDM candidates
for ii=1:n
    ipMod = randsrc(1,N,alphaMqam) + 1i*randsrc(1,N,alphaMqam);
    xF1 = sqrt(2)*v*reshape(ipMod,N,1); % grouping into multiple symbols
    for sml=1:U(mm)
        xF(:,sml) = xF1 .* B(:,sml);
        sk(sml,:) = N/sqrt(N) *ifft(fftshift(xF(:,sml)),N);
        sk1(sml,:) = reshape(sk(sml,:),1,N);
        PAPR_dB_smlxx(sml) = 10*log10( max(abs(sk1(sml,:)).^2)./
                                   mean(abs(sk1(sml,:)).^2));
    end
    PAPR_dB(mm,ii) = PAPR_dB_smlxx(1);
    PAPR_dB_sml(mm,ii) = min(PAPR_dB_smlxx);
end
% ---------------- calculating CCDF ----------------
for tt=1:1:length(PAPRosim)
    CCDF_sim_slm(mm,tt) = length(find(PAPR_dB_sml(mm,:) >=
                                    PAPRosim(tt)) ) / length(PAPR_dB_sml(mm,:));
end
figure
semilogy(CCDF_th,CCDF_SLM_(1,:),'-'); hold on;
semilogy(PAPRosim,CCDF_sim_slm(1,:),'-o'); grid on;
semilogy(CCDF_th,CCDF_SLM_(2,:),'-');
semilogy(CCDF_th,CCDF_SLM_(3,:),'-');
semilogy(CCDF_th,CCDF_SLM_(4,:),'-');
semilogy(CCDF_th,CCDF_SLM_(5,:),'-');
semilogy(PAPRosim,CCDF_sim_slm(2,:),'--o');
semilogy(PAPRosim,CCDF_sim_slm(3,:),'--o');
semilogy(PAPRosim,CCDF_sim_slm(4,:),'--o');
semilogy(PAPRosim,CCDF_sim_slm(5,:),'--o');
axis([ 4 12 10^-3 1]);
hleg1 = legend('Analytical','Simulation');
xlabel('PAPR_o (dB)'); ylabel('CCDF');
A.2 Principle of DPTE Method

clc
clear
p = 0.01;
SNR = 25;
SINR = -10;
sigmag2 = 1 / 10^(SNR/10); % Gaussian noise variance
sigmai2 = 1 ./ 10.^(SINR/10); % Impulsive noise variance
N = 64; % fft size
n = 100; % Number of OFDM symbols
Tb = 0:0.01:7; % Blanking threshold vector
% M-QAM modulation
M = 16; % constellation size
v = sqrt(1/((2/3)*(M-1))); % normalizing factor
a = [1:sqrt(M)/2]; % alphabets
alphaMqam = [-(2*a-1) 2*a-1];
for w=1:n
xF = (randsrc(1,N,alphaMqam) + 1i*randsrc(1,N,alphaMqam));
xF = v*reshape(xF,N,1); % grouping into multiple symbols
sk1 = N/sqrt(N)*ifft(fftshift(xF),N); % Taking ifft,
sk = sqrt(2)*reshape(sk1,1,N);
Peak(w) = max(abs(sk)); % Calculating PAPR
% Generating Gaussian and Impulsive noise
ng = sqrt(sigmag2)*(randn(1,N) + 1i*randn(1,N));
nii = sqrt(sigmai2)*(randn(1,N) + 1i*randn(1,N));
m = rand(1,N) > (1-p);
ni = m .* nii;
rk = sk + ng + ni; % Received signal
% Finding the optimal blanking threshold
for t=1:length(Tb)
rk_abs = abs(rk);
rk_ang = angle(rk);
index1 = find(rk_abs > Tb(t));
rk_abs(index1) = 0;
yk(t,:) = rk_abs .* exp(1i*rk_ang);
nk = yk(t,:) - sk;
SNR(t) = 10*log10(mean(abs(sk).^2)/mean(abs(nk).^2));
end
[C I] = max(SNR);
Topt(w) = Tb(I);
end

% Finding best fit
Peak1 = polyfit(Peak,Topt,1)
Topt1 = polyval(Peak1,Peak);
figure
plot(10*log10(Peak),10*log10(Topt),'.'); hold on;
plot(10*log10(Peak),10*log10(Topt1),'-b'); grid on;
axis([3 6.5 3 6.5]);
xlabel('Peak value (dB)');
ylabel('Optimal Blanking Threshold (dB)');
Appendix B

Best Student Paper at ISPLC’15

Constant Envelope OFDM Transmission Over Impulsive Noise Power-Line Communication Channels

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Abstract

Signal blanking is a simple and efficient method to reduce the effect of impulsive noise over power-line channels. The efficiency of this method, however, is found to be not only impacted by the threshold selection but also by the average peak-to-average ratio (PAPR) value of the orthogonal frequency division multiplexing (OFDM) signals. As such, the blanking capability can be further enhanced by reducing the PAPR value. With this in mind, in this paper we evaluate the performance of constant envelope OFDM (CE-OFDM) which has inherently the lowest achievable PAPR of 0 dB; therefore, the proposed system is expected to provide the lower bound performance of the blanking-based
method. In order to characterize system performance, we consider the probability of blanking error and signal-to-noise ratio (SNR) at the output of the blanking device. The results reveal that the proposed system can achieve significant improvements over the conventional OFDM blanking-based scheme in terms of minimized probability of blanking error. It will also be shown that output SNR gains of up to 6 dB can be attained over the conventional OFDM blanking-based systems.

B.1 Introduction

It is commonly known that impulsive noise (IN) is the main factor responsible for degrading communication signals over power-line communication (PLC) channels [13, 44]. Multi-carrier modulation (MCM) systems, such as orthogonal frequency division modulation (OFDM), are found to be more robust to this noise compared to single-carrier systems due to the property of spreading the noise energy over the available sub-carriers. Such a property, however, could turn into a disadvantage if the noise energy is very high where all sub-carriers become contaminated with high levels of interference. In such noise environments, other techniques, with varying degrees of effectiveness and complexity, should be utilized; the simplest and most effective of which remains the blanking scheme. In this scheme, the receiver is preceded with a blanking device to zero the incoming signal when it exceeds a certain threshold value. Blanking the unaffected useful signals will cause blanking errors and hence performance will degrade. In fact, one of the major causes of blanking errors is the high signal peaks associated with conventional OFDM systems, also known as the peak-to-average power ratio (PAPR) problem.

To reduce the PAPR of OFDM systems, many techniques have been introduced in the literature which are, generally, classified into distortionless techniques such as coding [87], tone reservation [83] and selective mapping (SLM) [98]; and non-distortionless techniques such as signal clipping and peak cancellation [79, 80]. In general, although distortionless techniques are relatively more complex than the non-distortionless ones, they are still more attractive because the distortion caused by non-distortionless schemes can outweigh the benefit of the reduced PAPR. The other alternative technique is based on transforming the OFDM signal prior to transmission and applying
the inverse transformation at the receiver prior to demodulation. For instance, a companding transform is studied in [137, 138] where companded signals have an increased average power and, consequently, attains lower PAPR compared to conventional OFDM but, relatively, it is still large compared to single-carrier modulation. On the other hand, the authors in [139–141] proposed a phase modulator transform where it is shown that such a system can offer the lowest possible PAPR value of 0 dB. This system is referred to as constant envelope OFDM (CE-OFDM). Unlike conventional OFDM which amplitude modulates the carrier, CE-OFDM uses the OFDM signal to phase modulate the carrier and the inverse transform is performed at the receiver prior to the OFDM demodulator. Phase modulation transforms the amplitude variations into a constant amplitude signal by phase modulating the carrier which results in transforming the high PAPR OFDM signal to a 0 dB PAPR constant envelope waveform. It should be highlighted that transmitting OFDM by ways of angle modulation to reduce PAPR has been in the literature for a while but it was only studied in the context of power amplifier nonlinearity. In contrast to these studies, in this paper we utilize this property to enhance the capability of the impulsive noise mitigation process in PLC systems.

Therefore, the contribution of this paper is threefold. Firstly, the PAPR performance of the CE-OFDM and its impact on the probability of blanking error are studied. The second contribution resides in evaluating the signal-to-noise ratio (SNR) at the output of the blanking device. Then the problem of blanking threshold optimization is addressed and the corresponding maximum achievable output SNR is presented. For the sake of comparison and completeness, throughout our investigations results for the conventional OFDM blanking-based scheme are included. Results show that CE-OFDM with a blanking device can significantly outperform the OFDM-based system in terms of the probability of blanking error. Furthermore, it is shown that the proposed system is able to achieve a gain of up to 6 dB in the output SNR relative to the conventional OFDM scheme.

The rest of the paper is organized as follows. Section B.2 highlights the key differences between the OFDM and CE-OFDM systems. In Section B.3, the system model is presented. Section B.4 evaluates the complementary cumulative distribution function (CCDF) and probability of blanking error performances for both the CE-OFDM and OFDM systems. In Section B.5 the
output SNR is studied as a function of the blanking threshold whereas the problem of blanking threshold optimization is addressed in Section B.6. Finally, conclusions are drawn in Section B.7.

## B.2 CE-OFDM Versus OFDM Signaling

In this section we present the main differences between the CE-OFDM and OFDM systems. To start with, a typical OFDM waveform is represented by

\[
x(t) = \sum_{k=0}^{N-1} X_k e^{j2\pi kt/T_B}, \quad 0 \leq t \leq T_B
\]  

(B.1)

where \(X_k\) is the complex constellations of the data symbols, \(T_B\) is the active symbol interval, \(N\) is the number of sub-carriers. Therefore, the real and imaginary parts of the OFDM signal can be expressed as

\[
\mathfrak{R}\{x(t)\} = \sum_{k=0}^{N-1} \mathfrak{R}\{X_k\} \cos(2\pi kt/T_B)
\]

\[
- \mathfrak{I}\{X_k\} \sin(2\pi kt/T_B)
\]

(B.2)

and

\[
\mathfrak{I}\{x(t)\} = \sum_{k=0}^{N-1} \mathfrak{R}\{X_k\} \sin(2\pi kt/T_B)
\]

\[
+ \mathfrak{I}\{X_k\} \cos(2\pi kt/T_B),
\]

(B.3)

respectively, and the instantaneous signal power is given by

\[
|x(t)|^2 = \mathfrak{R}^2\{x(t)\} + \mathfrak{I}^2\{x(t)\}
\]  

(B.4)

This clearly indicates that the signal power fluctuates over time. On the other hand, the general constant envelope (CE) signal has the following form

\[
s(t) = Ae^{j\phi(t)}
\]  

(B.5)

where \(A\) is the signal amplitude and \(\phi(t)\) is its phase. CE-OFDM combines the advantages of both OFDM and CE modulation in which the high PAPR OFDM signal is converted into a constant envelope waveform. CE-OFDM has
the form of (B.5) where the phase signal is a real-valued OFDM waveform which can be generated in various ways. For instance, in [142] the complex OFDM signal is divided into its real and imaginary components prior to the phase modulator whereas in [143] discrete cosine transform (DCT) is used. For this paper, to generate the real-valued OFDM signal we use the Hermitian symmetry to map the symbols prior to the OFDM modulator as will be illustrated in the next section. Since the OFDM signal is used to phase modulate the carrier, unlike the conventional OFDM which amplitude modulates the carrier, the phase signal of CE-OFDM can be expressed as

\[
\phi(t) = \Re\{x(t)\} = \sum_{k=0}^{N-1} \Re\{X_k\} \cos\left(\frac{2\pi kt}{T_B}\right) - \Im\{X_k\} \sin\left(\frac{2\pi kt}{T_B}\right) \tag{B.6}
\]

The CE-OFDM signal power is simply \(|s_{ce-ofdm}(t)|^2 = A^2\), and hence its PAPR is 0 dB. For better clarity, illustrative examples of the instantaneous power of the OFDM and the mapped CE-OFDM signals are presented in Fig. B.1 from which it is clear that the peak and the average values of the CE-OFDM signal are the same hence its PAPR is 0 dB. It should be mentioned here that the spectral efficiency of CE-OFDM is not as good as conventional OFDM since real-valued signals are required at the input of the phase modulator [140]. Similar to conventional OFDM, CE-OFDM uses cyclic prefix to simplify equalization over multipath fading channels.

### B.3 System Model

The CE-OFDM system model under consideration is shown in Fig. B.2 where the unshaded blocks basically represent the conventional OFDM system. In both systems, the information bits are first mapped into 16-QAM symbols, \(X = [X_0, X_1, \ldots, X_{N/2-1}]\), which are then serial-to-parallel (S-to-P) converted. For the OFDM system, these symbols are passed directly through an inverse discrete Fourier transform (IDFT) to produce complex valued sequence, \(x_n^{\text{ce}}\), given by
Figure B.1: Instantaneous power of CE-OFDM and OFDM signals.

\[ x_n^{cw} = \sum_{k=0}^{N-1} X_k e^{j2\pi kn/N}, \quad n = 0, 1, \ldots, N - 1 \quad \text{(B.7)} \]

However, for the CE-OFDM signal a real-valued time domain OFDM signal is obtained by enforcing Hermitian symmetry in the symbols prior to the IDFT such that

\[ X = [0, X_1, \ldots, X_{N/2-1}, X_{N/2-1}^*, \ldots, X_1^*, X_0^*] \quad \text{(B.8)} \]

The zero at index \( k = 0 \) is used to maintain conjugate symmetry. This sequence is then passed through the \( N \)-point IDFT to yield a real-valued sequence, \( (x_n^{rv}) \), which can be expressed as

\[ x_n^{rv} = 2\Re \left\{ \sum_{k=1}^{\tilde{N}-1} X_k e^{j2\pi kn/\tilde{N}} \right\}, \quad n = 0, 1, \ldots, \tilde{N} - 1 \quad \text{(B.9)} \]

where \( \tilde{N} = 2N \). The PAPR of the transmitted signal is
Figure B.2: CE-OFDM system model with blanking at the receiver.
PAPR = \frac{\max(|x_n|^2)}{\mathbb{E}[|x_n|^2]} \quad (B.10)

After that the real-valued OFDM signal is fed to a phase modulator to produce \( s(t) = A e^{j\phi(t)} \), where \( A \) is the signal amplitude. In this paper we adopt the two component mixture-Gaussian noise model to characterize the total noise in which IN is a result of two random sequences, a real Bernoulli and complex Gaussian process \([44,144]\)

\[ n_n = w_n + i_n \quad n = 0, 1, 2, \ldots, \tilde{N} - 1 \quad (B.11) \]

where

\[ i_n = b_n g_n, \quad n = 0, 1, 2, \ldots, \tilde{N} - 1 \quad (B.12) \]

while \( n_n \) is the total noise component, \( w_n \) is the additive white Gaussian noise (AWGN), \( i_n \) is the IN, \( g_n \) is complex white Gaussian noise with mean zero and \( b_n \) is the Bernoulli process with probability mass function

\[
P(b_n) = \begin{cases} 
  p, & b_n = 1 \\
  1 - p, & b_n = 0 
\end{cases} \quad n = 0, 1, \ldots, \tilde{N} - 1 \quad (B.13) \]

where \( p \) denotes the IN probability of occurrence. The probability density function (PDF) of the total noise can therefore be expressed as

\[
P_{n_n}(n_n) = p_0 \mathcal{G}(n_n, 0, \sigma^2_0) + p_1 \mathcal{G}(n_n, 0, \sigma^2_1) \quad (B.14)\]

\( \mathcal{G}(.) \) is the Gaussian PDF given by \( \mathcal{G}(x, \mu, \sigma^2) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left(-\frac{(x-\mu)^2}{2\sigma^2}\right) \), \( p_0 = 1 - p, \) \( p_1 = p, \) \( \sigma^2_0 = \sigma^2_w \) and \( \sigma^2_1 = \sigma^2_w + \sigma^2_i \). The variances \( \sigma^2_w \) and \( \sigma^2_i \) denote the AWGN and IN power which define the input SNR and signal-to-impulsive noise ratio (SINR) as \( \text{SNR} = 10 \log_{10}(1/\sigma^2_w) \) and \( \text{SINR} = 10 \log_{10}(1/\sigma^2_i) \), respectively. The
received signal is then given as

$$r_n = \begin{cases} s_n + w_n, & H_0 \\ s_n + w_n + i_n, & H_1 \end{cases} n = 0, 1, \ldots, \tilde{N} - 1$$  \hspace{1cm} (B.15)$$

where the null hypothesis $H_0$ implies the absence of IN, $P(H_0) = 1 - p$, whereas the alternative hypothesis $H_1$ implies the presence of IN, $P(H_1) = p$. At the receiver’s front-end blanking is applied as follows

$$y_n = \begin{cases} r_n, & |r_n| \leq T \\ 0, & |r_n| > T \end{cases} n = 0, 1, \ldots, \tilde{N} - 1$$  \hspace{1cm} (B.16)$$

where $T$ is the blanking threshold, $r_k$ and $y_k$ are the input and the output of the blanking device, respectively. It should be noted that the blanking device only processes the amplitude of the received signal while preserving its phase. Next, $y_n$ is fed to the S-to-P convertor and then passed through the phase demodulator where $\text{arg}(.)$ calculates the phase of the received samples while the phase unwrapper is used to reduce the effect of phase ambiguities by adding multiples of $2\pi$ radians where appropriate to eliminate jumps greater than $\pi$ radians.

### B.4 CCDF and Probability of Blanking Error

The PAPR value is a random quantity depending on the input data symbols $X_n$ and therefore, its distribution is usually viewed statistically in terms of CCDF \[108\]. The CCDF of PAPR is defined as the probability that the PAPR of a data block exceeds a given threshold ($\text{PAPR}_o$) and is expressed mathematically as $\text{CCDF} = \Pr \{\text{PAPR} > \text{PAPR}_o\}$. For accurate estimates of the PAPR, oversampling of 4 is implemented in our investigations. Fig. B.3 compares results of the CCDF for the CE-OFDM and conventional OFDM systems with $N = 64$. The analytical results of the latter system are found as

$$\text{CCDF}_{\text{ofdm}} = 1 - (1 - \exp(-\text{Peak}_o))^N$$  \hspace{1cm} (B.17)$$

From this figure it is evident that the analytical results of the conventional OFDM scheme correlate well with the simulated ones. It is also clearly visible
that the CE-OFDM system can drastically reduce the signal peak. For instance, at CCDF $= 10^{-3}$ the proposed system is able to reduce the peak value by 3.5 units compared to OFDM. This PAPR enhancement implies minimizing the probability of blanking error ($P_b$). $P_b$ is defined as the probability that the received signal, $A_r = |r_n|$, exceeds the blanking threshold when it is unaffected by IN and is given mathematically by the following joint probability

$$P_b = P(B, H_0) = Pr(A_r > T | H_0) P(H_0) \tag{B.18}$$

where $B$ is the event of blanking. It was shown in [76] that the major cause responsible for worsening this probability is the high PAPR of the transmitted signal which is an inherent problem associated with the OFDM waveform. On the contrary, CE-OFDM, as shown above, has the lowest possible PAPR value of 0 dB allowing to achieve the minimum probability of blanking error performance. To illustrate this, we plot in Fig. B.4 some results for $P_b$ performance versus $T$ for both the CE-OFDM and conventional OFDM systems.
Figure B.4: Probability of blanking error versus blanking threshold for the CE-OFDM and OFDM systems.

It should be pointed out that the analytical results for the latter scheme are found using [76]

\[ P_b = \exp \left( -\frac{T^2}{2(\sigma_x^2 + \sigma_w^2)} \right) (1 - p) \]  

(B.19)

The first observation one can see from these results is that the analytical and simulated results of the conventional OFDM scheme are in a good agreement. It should be stated here that our simulations from this point onward, unless it is clearly stated otherwise, are based on CE-OFDM/OFDM with \( N = 64 \) sub-carriers, oversampling rate of 4, input SNR = 25 dB and SINR = −15 dB. It can clearly be seen from Fig. B.4 that the CE-OFDM scheme is very robust against blanking errors providing zero probability of blanking error at around 1 whereas in the case of OFDM \( P_b = 0.3 \) at the same blanking threshold value and it improves as \( T \) increases. It is also observed that when \( T < 1 \), the proposed system has \( P_b = 1 \). This is due to the fact that in this region all the signal samples have amplitudes of 1, as shown in
Fig. B.1, and will consequently all be blanked. Furthermore, it is worth mentioning that in this region the conventional OFDM system has slightly lower probability of blanking error in comparison to CE-OFDM.

### B.5 Output SNR as Versus Blanking Threshold

This section investigates the impact of blanking on the SNR at the output of the blanking device. Fig. B.5 illustrates the output SNR versus $T$ for the CE-OFDM and conventional OFDM systems for various IN probabilities when input SNR = 25 dB and SINR = −15 dB. It is clear that the analytical results of the conventional OFDM system, which are obtained using [65, (2) and (26)], and the simulated ones are matching. The simulated output SNR is simply calculated as follows $\text{SNR}_{\text{out}} = \mathbb{E} \left[ |x_n|^2 \right] / \mathbb{E} \left[ |y_n - x_n|^2 \right]$. A common observation one can see for both systems is that the output SNR performance degrades as $p$ becomes higher. However, irrespective of the IN probability, CE-OFDM with blanking can generally outperform the OFDM-based scheme when the blanking threshold value is moderate providing a SNR gain of up to 4 dB and 3 dB when $p = 0.01$ and 0.05, respectively; whereas when $T$ is very large both systems perform similarly due to the fact that IN will impact both systems in a similar manner in the absence of the blanking device. On the other hand, when $T$ is very small both systems suffer greatly from severe performance degradation since most of the useful signal energy is lost. Furthermore, it is evident that for each system, and for each IN probability, there exists an optimal blanking threshold that maximizes the output SNR. Interestingly enough, however, and unlike the OFDM system in which the maximum achievable output SNR is reached gradually, in the CE-OFDM system the output SNR hits its peak sharply at around $T = 1$. This phenomena is of great importance as discussed below.

### B.6 Performance Optimization

In this section extensive simulations are conducted to find the optimal blanking threshold of the proposed scheme as well as the corresponding maximum achievable output SNR in various noise scenarios. To start with, we plot the
optimal threshold versus SINR in Fig. B.6 for both the CE-OFDM and conventional OFDM systems with $p = 0.01$ and $0.1$. Comparing the CE-OFDM and OFDM results, it can be seen that the optimal threshold of the former scheme is always lower compared to that of the latter. In addition, unlike conventional OFDM in which the optimal threshold decreases as $p$ becomes higher, in the CE-OFDM approach it remains constant regardless of the value of $p$. Another interesting feature for the proposed scheme is the fact that the optimal threshold is independent of SINR. These advantageous properties of the proposed system imply that optimal performance can always be achieved with no need for prior knowledge about the noise characteristics. This will, consequently, eliminate the requirement for noise estimations and hence estimation errors can be avoided and receiver complexity can also be reduced considerably.

The maximum achievable output SNR for the optimized systems is plotted in Fig. B.7 versus SINR for various pulse probabilities. In order to provide comparative results, the output SNR of the conventional OFDM approach is also included on this plot. As a first remark on these results, it is evident that
for both modulation schemes the output SNR is reduced as $p$ increases. It is also obvious that the proposed technique always outperforms the conventional OFDM system throughout the SINR spectrum while offering highest gains in the intermediate SINR region ($-3 \, \text{dB} > \text{SINR} > -10 \, \text{dB}$) of up to 6 dB and 5 dB when $p = 0.01$ and 0.1, respectively. However, in the low SINR region ($\text{SINR} < -30 \, \text{dB}$), both systems behave similarly which is intuitively justified by the fact that when noise pulses are extremely high both schemes are able to perfectly detect and cancel the noise pulses. To summarize, CE-OFDM with blanking has the following advantageous properties: a) it has a better performance than conventional OFDM and does not require any noise estimations to combat IN b) As shown in Fig. B.2, CE-OFDM can be obtained from the conventional OFDM with only minor adjustments.

Figure B.6: Optimal blanking threshold versus SINR for the CE-OFDM and OFDM systems with various values of $p$. 
**Figure B.7:** Maximum achievable output SNR as a function of SINR for the CE-OFDM and OFDM systems with different $p$ values.

### B.7 Conclusion

In this paper we investigated the performance of CE-OFDM with a blanking device over IN power-line channels under various noise conditions. The performance of the proposed system is assessed in terms of the probability of blanking error and output SNR. In addition, the problem of blanking threshold optimization is addressed and the corresponding maximum achievable output SNR is presented. For comparison sake, results for the conventional OFDM based system are demonstrated throughout the paper. It is found that significant improvements can be attained with the proposed technique in the form of considerably minimized probability of blanking error and increased output SNR. Furthermore, it is demonstrated that output SNR gains of up to 6 dB can be achieved relative to the OFDM system in the intermediate SINR region.
Production Notes

The main body of this thesis was typeset using the LyX 2.1.1 document preparation software [145], and the bibliography was done using BibTeX. The two main numerical tools used in this thesis are MATLAB R2012a, [146], and Mathematica 9.0.1.0 [147]. The block diagrams and system models were drawn using CorelDRAW X3 [148]. The LyX output was converted to Portable Document Format (PDF), the size of which is about 7.3 megabytes. This thesis takes approximately 15 seconds to compile.

The work was done at the University of Manchester on a Dell Optiplex 980 workstation running 64-bit Windows 7 Enterprise operating system with a 2.93 gigahertz microprocessor and 8 Gigabyte of memory. This thesis was printed on a Hewlett Packard LaserJet 1300n printer.