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OFDM impairment mitigation techniques

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OFDM IMPAIRMENT MITIGATION TECHNIQUES

by

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B.Sc., University of Moratuwa, Sri Lanka, 2004

A thesis
presented to Ryerson University
in partial fulfillment of the
requirement for the degree of
Master of Applied Science
in the Program of
Electrical and Computer Engineering. Toronto, Ontario, Canada, 2007

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Abstract

OFDM Impairment Mitigation Techniques

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**Master of Applied Science
Electrical and Computer Engineering
Ryerson University**

In this thesis, we propose novel techniques to improve the performance of OFDM systems. We present a simple adaptive modulation technique to mitigate the nonlinear distortion effects of OFDM signals. Based on an estimation of the nonlinearity of the HPA/channel, for each OFDM symbol, a calculation is done at the transmitter side which identifies the subcarriers with high distortion and correspondingly reduces the modulation level on those subcarriers. This procedure is repeated until the nonlinear distortion is below a predetermined threshold. This technique is shown to improve the BER performance considerably while the reduction in data rate is small. The data rate is reduced by 4% for a system with 64 subcarriers and 16 QAM as primary and 4 QAM as secondary modulation levels.

The tone reservation technique used in conventional PAPR reduction is suitably modified to provide a simple solution to reduce the average power requirement in intensity modulated optical OFDM systems. With two reserved subcarriers the reduction in power is approximately 2 dB for 16 QAM modulation with 64 subcarriers and 1 dB for 256 subcarriers.

We also describe techniques to improve the BER performance of grouped linear constellation precoding (GLCP) OFDM which is used to achieve frequency diversity. We present an Adaptive Weighting (AW) technique in which during MLSE, the distance of the possible constellation points from the received symbols are weighted according to the SNR of each subcarrier before the sequence with the minimum distance is chosen. We also analyze a sub optimum iterative decoding algorithm which improves the performance of an initial zero forcing detection iteratively on a symbol by symbol basis. Both techniques improve the performance of the GLCP-OFDM system considerably at the expense of increased complexity.

Acknowledgments

Making a thesis a reality takes many dedicated people, and it is a great pleasure to acknowledge the contributions of all the people who have made this possible.

My sincere thanks to my supervisor Dr.Xavier Fernando for his expert guidance, encouragement and infinite patience. His kind advice and friendly approach made this project a memorable and enriching experience.

I would also like to acknowledge the Department of Electrical and Computer Engineering and the School of Graduate Studies at Ryerson University for their support in terms of financial aid, and work experience as a teaching assistant.

Finally, I owe a great deal to my colleagues in Wincore Lab who through their own research, comments and questions have encouraged, supported and motivated me.

Without doubt there will be errors, omissions and over-simplifications, for which I take absolute responsibility, while hoping that the rest of the material will be enough to stimulate insights into the research in OFDM systems and their potential in future wireless communications.

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List of Acronyms

ADC	analog to digital conversion
AWGN	additive white gaussian noise
BER	bit error rate
BPSK	binary phase shift keying
CDMA	code division multiple access
COFDM	coded OFDM
DAB	digital audio broadcasting
DAC	digital to analog conversion
DC	direct current
DFT	discrete fourier transform
DMT	discrete multi tone
DSL	digital subscriber loop
DSP	digital signal processing
DVB-T	digital video broadcasting-terrestrial
EDGE	enhanced data rates for GSM evolution
EVDO	evolution data optimized
FDM	frequency division multiplexing
FDMA	frequency division multiple access
FEC	forward error correction
FFT	fast fourier transform
GLCP	grouped linear constellation precoding
GLCP-ACS	GLCP adaptive subcarrier selection
GLCP-AW	GLCP adaptive weighting
GPRS	global packet radio system
GSM	global system for mobile communications
HDTV	high definition television
HPA	high power amplifier
HSDPA	high speed downlink packet access
IBO	input backoff
ICI	inter carrier interference
IEEE	institute of electrical and electronics engineers
IFFT	inverse fast fourier transform

IM/DD	intensity modulation and direct detection
IPTV	internet protocol television
IS-95	interim standard 95
ISI	inter symbol interference
LED	light emitting diode
LPF	low pass filter
LTE	long term evolution
MIMO	multiple-input multiple-output
MLSE	maximum likelihood sequence estimation
MMSE	minimum mean square error
MRC	maximum ratio combining
MZM	mach zehnder modulator
M-ZPSK	mary zero phase shift keying
OFDM	orthogonal frequency division multiplexing
OFDMA	OFDM access
OFDM-SS	OFDM spread spectrum
OW	optical wireless
PAPR	peak to average power ratio
PDL	polarization dependent loss
P/S	parallel to serial
QAM	quadrature amplitude modulation
QOS	quality of service
QPSK	quadrature phase shift keying
PRT	peak reduction tones
PSD	power spectral density
RA-OFDM	rate adaptive OFDM
RF	radio frequency
RMS	root mean square
ROF	radio over fiber
SFBC	space frequency block codes
SI	side information
SLM	selected mapping
SNR	signal to noise ratio
S/P	serial to parallel

STBC	space time block codes
TCM	trellis coded modulation
UMTS	universal mobile telecommunication system
VoIP	voice over internet protocol
W-CDMA	wideband CDMA
WiMax	wireless microwave access
WLAN	wireless local area network
2G	second generation
3G	third generation
4G	fourth generation
2-D	two dimensional
3-D	three dimensional
4-D	four dimensional

Chapter 1

Introduction

The demand for advanced information services is growing in terms of both the number of users and the services to be supported. Voice and low-rate data services are insufficient for users in a world where high-speed world wide web access is taken for granted. The trend is toward global information networks offering flexible multimedia information services to users on demand. Potential services include video on demand, interactive video, fast internet access, telemedicine, tele-education, and large file transfer. In this context the wireless handheld or the mobile phone is looked upon as the all in one device which should necessarily cater for all these requirements. This increased requirement also puts enormous demand on the network providers. A look at current mobile wireless networks would reveal that they do not carry enough capacity for such increased demand.

The 2G networks based on IS-95 and GSM were essentially designed for voice and low rate data traffic and provided a throughput of 9.6 kbps. With the 2.5G systems the GPRS allowed for the increased throughput of 171.2 kbps, another variant the EDGE allowed for throughputs of 236.4 kbps and IS-95B allowing for 115.2 kbps. The third generation of systems are targeted at data rates of 144 kbps for satellite and rural outdoor, 384 kbps for urban outdoor and 2048 kbps for indoor and low range outdoor [1]. The packet-based data service EVDO in the 3G CDMA2000 standard gives downlink data rates of 2.4-3.1 Mbps over a 1.25 MHz bandwidth, whereas in the competing W-CDMA (UMTS) standard HSDPA allows for data transmission up to 8-10 Mbps (and 20 Mbps for future MIMO systems) over a 5 MHz bandwidth in downlink.

It is expected that the future mobile users are not mere consumers of digital data but they are also active content creators. With social networking/blogging, content creation/sharing

and podcasting being a necessity in the future, there is always need for higher data rates not only in the downlink but also in the uplink. Looking beyond 3G we see an increased interest in exploiting the advantages offered by Orthogonal Frequency Division Multiplexing (OFDM) in achieving higher throughputs.

OFDM is a signalling method that uses a parallel transmission of several low rate signals each modulated into a different carrier frequency (subcarriers). The subcarriers are orthogonal to each other and therefore their individual bandwidth overlap resulting in efficient utilization of available bandwidth. Since each subchannel is narrowband, they experience almost flat channel response, which makes equalization (if necessary) very simple - the basic idea is that the low rate subcarriers experience very little inter symbol interference (ISI) when transmitted through a multipath channel. The inherent resistance to ISI which is one of the major impediments to high data rate transmission is one of the advantages of OFDM systems. OFDM can also be viewed as a multiple access scheme in that different user data or different types of data can be transmitted via different subcarriers. With proper channel coding or symbol precoding OFDM systems can overcome frequency selective fading and narrowband interference. A combined time-frequency optimization approach can be naturally extended to OFDM systems and this makes it an attractive option for future wireless systems. Another advantage of OFDM systems is that the data rates can be easily varied depending on the channel conditions, quality of service (QoS) requirements by means of adaptive modulation and space-time-frequency signal processing and dynamic scheduling. This flexibility makes OFDM ideally suited for next generation wireless systems where variable data rates is a key requirement. Due to these benefits, the UMTS Long Term Evolution (LTE) or '3.9G' which aims for 100 Mbps in the downlink and 50 Mbps is essentially based on OFDM technology. Orthogonal Frequency Division Multiple Access (OFDMA) is employed in the downlink and Single Carrier Frequency Division Multiple Access (SC-FDMA, also known as 'precoded OFDM' or 'DFT-spread OFDM') is used in the uplink [2]. It can also be noted that MIMO-OFDM has already been proposed as one of the 4G system candidates.

OFDM has been in use in broadcast networks such as DVB-T and DAB for some time with data rates upto 24 Mbps for DVB-T and 1.152 Mbps for DAB [3]. In the area of fixed access wireless communications such as WLAN also OFDM has been used. The huge popularity of the IEEE 802.11 a/g standards which provide data rates upto 54 Mbps for short ranges with high reliability and good reception has led to further research in the application of OFDM in this area. The 802.11n standard which is still under development aims to provide

typical data rates of 200 Mbps over short distances using MIMO-OFDM. There has been considerable development in the fixed access wireless communications systems (broadband wireless access) too. The IEEE 802.16 standard which is popularly known as WiMax targets throughputs of 100 Mbps over long distances and two out of the three proposed air interfaces for this system are based on OFDM [4].

The demand for more bandwidth is felt even in the seemingly infinite bandwidth optical fiber backbone networks. With internet seen as the preferred medium for future media delivery and with the convergence of VoIP, Internet Services and IPTV as the so called Triple Play there is considerable pressure on backbone networks. In addition to the voice and data traffic, the ever increasing demand for high quality multimedia such as HDTV from the internet puts increased pressure on existing networks. Recently there has been much research on the use of advanced modulation techniques such as OFDM instead of simple ON-OFF Shift Keying in existing multimode fibers as a low cost solution to this problem. OFDM offers good spectral efficiency and high immunity to chromatic dispersion [5]. The cyclic prefix for the OFDM symbols is shown to combat the dispersion of the multipath fiber channel [6], and hence well suited to the existing multimode fiber infrastructure. Also in [7] it has been shown that polarization mode dispersion provides a benefit of polarization diversity to multicarrier systems against polarization-dependent-loss (PDL) induced fading. The authors in [5] have demonstrated 100 Gbps data rates with multicarrier systems.

The distortion caused by high power amplifier (HPA)/laser nonlinearity, increased average power requirement in intensity modulated optical systems, loss of orthogonality of subcarriers due to Doppler spreading resulting in intercarrier interference (ICI), the vulnerability to frequency selective fading, increased hardware complexity and the need for more processing power are some issues which should be addressed for successful implementation of OFDM in the above mentioned areas. It is the author's opinion that when the practical issues are resolved we could see the widespread adaptation of OFDM systems in future wireless as well as wired communication systems.

There has been a number of papers in recent times which concentrate on nonlinear distortion and peak to average power ratio (PAPR) reduction of OFDM systems [8, 9, 10, 11]. We can also find a variety of channel coding [12, 13, 14] and symbol precoding [15] schemes with varying degrees of complexity in the literature which attempt to improve performance in frequency selective fading channels. The work in [16, 17, 18] describes various techniques to reduce the ICI as a result of Doppler spreading. In this thesis we present some

simple techniques to mitigate the effects of nonlinear distortion, to reduce the average power requirement and to improve performance in frequency selective fading channels.

This thesis is organized as follows. In Chapter 2, we introduce the theory behind OFDM. The key aspects of OFDM which makes it an attractive modulation scheme are discussed. We also discuss the problems associated with OFDM in detail.

Chapter 3 provides a literature survey of previous work done in mitigating OFDM impairments. We review various techniques that have been proposed to solve the nonlinear distortion problem. Also, we give some background on grouped linear constellation precoding (GLCP) OFDM and GLCP adaptive subcarrier selection (ACS) OFDM which forms the basis for our GLCP (adaptive weighting) AW OFDM. A brief summary is also given about a previous average power reduction technique in optical OFDM systems.

In Chapter 4, we describe the adaptive modulation technique in detail with appropriate system model. We describe the average power reduction technique in the context of a diffuse optical wireless system. We also present and analyze the GLCP-AW technique and the iterative decoding algorithm in the context of a wireless fading channel.

The system parameters and numerical results for the techniques proposed in Chapter 4 are presented in Chapter 5. The performance of the proposed adaptive modulation technique is compared with the symbol interleaving technique and the efficiency and complexity is also analyzed. The results for the average power reduction scheme are presented and the complexity is analyzed. The BER performance of the GLCP-AW-OFDM and the iterative decoding algorithms are compared with GLCP-OFDM and GLCP-ACS-OFDM. Chapter 6 concludes the report and points out directions for future research.

Contributions

In this thesis, we analyze the performance of OFDM in the context of future high data rate communication systems and present novel techniques to improve the OFDM impairments encountered in such systems. In particular,

- the use of OFDM in indoor optical wireless systems is analyzed in the presence of fluorescent noise interference
- a technique to reduce the average power requirement of intensity modulated optical OFDM systems is developed

- an adaptive modulation technique to reduce the nonlinear distortion of OFDM systems is presented
- the use of simplex signals as a modulation technique to counter nonlinear distortion is analyzed
- a new adaptive weighting technique to improve the BER performance of grouped linear constellation precoding (GLCP) OFDM in frequency selective fading channels is presented.
- we apply an iterative decoding technique which was presented for nonlinear distortion mitigation (in time domain), to improve the performance of GLCP-OFDM (in frequency domain).

Chapter 2

Background

2.1 Orthogonal Frequency Division Multiplexing

In single carrier systems during one symbol period (T_s) the transmitted time domain signal spans the entire available bandwidth ($B_{available}$). When the data rate is high, the symbol duration becomes small and intersymbol interference (ISI) occurs when the delay spread (σ_τ) of the channel is greater than the symbol duration. In frequency domain this is seen as the frequency selectivity of the channel where the signal bandwidth (B_s) exceeds the coherence bandwidth (B_c) of the channel and hence different frequency components of the transmitted signal undergo different channel gains resulting in distortion. The ISI as described above is one of the primary limiting factors for high data rate transmission in single carrier systems. With multicarrier systems the given data stream is split into several sub streams of lower data rate and transmitted parallelly in adjacent frequencies (subcarriers) during the same symbol duration. This effectively increases the symbol duration (or reduces the bandwidth) on each subcarrier and hence the limits of ISI can be further put back.

Central to the multicarrier transmission concept is the principle of orthogonality in time and frequency domains. That is, the signal sent over different times should be distinguishable at the receiver and also the signals should be distinguishable at different frequencies for the same time duration. But the problem is that, a signal cannot be both bandlimited in time and in frequency. However, in reality it is good enough if the signal satisfies the orthogonal property at the particular sampling time or sampling frequency under consideration. Since the receiver in a communication system considers samples only at certain time or frequency, maintaining orthogonality at only these instants would be sufficient for interference free

reception. There can be two approaches to this problem. The first approach would be to use a bandlimited pulse which is orthogonal in time as shown in Figure 2.1. A pulse with a raised cosine spectrum is an example function which satisfies the above property - it is orthogonal in frequency domain because it is band limited and orthogonal to its time shifted version if the time shift corresponds with its period T_s . If the roll-off factor (α) is set to 0 the spectrum becomes strictly rectangular and provides maximum spectral efficiency. However, then the pulse in time domain becomes a sinc signal with infinite time duration which cannot be implemented in practice. The second solution is to use a pulse which is strictly time limited and orthogonal to its frequency shifted version if the frequency shift corresponds with its period. As shown in Figure 2.2 it is preferred to have a rectangular pulse shape for practical purposes (although there has been some interest in time limited non rectangular pulses for multicarrier transmission [19, 20]). This method can be realized in practice and popularly known as (narrow sense) OFDM. It can also be noted that in this approach OFDM can be considered as a variation of frequency division multiplexing (FDM), with better spectral efficiency. As shown in the figure, the base signal of an OFDM subcarrier can be represented

Figure 2.1: Multi carrier concept. (a) Time domain signal which has a raised cosine spectrum
(b) raised cosine spectrum $B = \frac{1+\alpha}{T_s}$, (c) raised cosine with $\alpha = 0, B = \frac{1}{T_s}$

by,

$$\begin{aligned} g_k(t) &= \sqrt{\frac{1}{T_s}} e^{j2\pi \frac{k}{T_s} t} \Pi\left(\frac{t}{T_s} - \frac{1}{2}\right), & t \in [0, T_s] \\ \Pi(t) &= 1, & t \in \left[-\frac{1}{2}, \frac{1}{2}\right] \end{aligned} \quad (2.1)$$

where $g_k(t)$ is the time domain baseband representation of the k^{th} subcarrier. Each subcarrier

Figure 2.2: Multi carrier concept. (a) Strictly time limited rectangular signal (b) sinc spectrum

is modulated with complex QAM symbols S'_k s and summed up together to produce the composite OFDM time domain signal $s(t)$.

$$s(t) = \sqrt{\frac{1}{T_s}} \sum_{k=-N/2}^{N/2} S_k e^{j2\pi \frac{k}{T_s} t} \Pi\left(\frac{t}{T_s} - \frac{1}{2}\right), \quad t \in [0, T_s] \quad (2.2)$$

Looking at equation (2.2), it is evident that the OFDM signal $s(t)$ is simply a Fourier synthesis of the complex QAM symbols to be transmitted, over the symbol duration $[0, T_s]$. At the receiver each symbol S_k can be recovered using a Fourier analysis in the time window $[0, T_s]$. Considering a two subcarrier OFDM system for simplicity ($k = 1, 2$),

$$s(t) = \sqrt{\frac{1}{T_s}} S_1 e^{j2\pi \frac{1}{T_s} t} + \sqrt{\frac{1}{T_s}} S_2 e^{j2\pi \frac{2}{T_s} t} \quad (2.3)$$

then, the symbol S_1 can be extracted from the signal $s(t)$ by,

$$\begin{aligned} S_{1, \text{received}} &= \sqrt{\frac{1}{T_s}} \int_0^{T_s} s(t) e^{-j2\pi \frac{1}{T_s} t} dt \\ &= \frac{1}{T_s} \int_0^{T_s} S_1 dt + \frac{1}{T_s} \int_0^{T_s} S_2 e^{j2\pi \frac{1}{T_s} t} dt \\ &= S_1 \end{aligned} \quad (2.4)$$

with the exponential term reducing to 0 when integrated over the whole period. In general, referring to equation (2.2) the k^{th} symbol can be recovered by,

$$S_{k,received} = \sqrt{\frac{1}{T_s}} \int_0^{T_s} s(t) e^{-j2\pi \frac{k}{T_s} t} dt \quad (2.5)$$

Note that two different time signals sent at different time intervals are orthogonal to each other because the Fourier window is $[0, T_s]$ - this is the time orthogonality discussed. The vanishing second term in equation (2.4) is the result of frequency orthogonality.

In order for ISI free reception the bandwidth of the subcarrier has to be less than the coherence bandwidth of the channel ($B_{subcarrier} \ll B_c$) - the symbol duration has to be larger than the delay spread of the multiple paths ($T_s \gg \sigma_\tau$). However, T_s has to be short enough to avoid distortions due to the time variance of the channel ($T_s \ll T_c$) that is related to Doppler shift. Therefore, an appropriate symbol duration has to be taken which satisfies both constraints in designing an OFDM system. If the total available bandwidth ($B_{available}$) is fixed, then the number of subcarriers (N) can be adjusted to achieve these conditions.

The modulation of the complex symbols onto the different subcarriers can be achieved by using a pair of cosine and sine carriers for each frequency as shown in Figure 2.3 and the products are summed to produce the OFDM time domain signal $s(t)$.

FFT and IFFT

For large number of subcarriers the modulation process as described above requires a large number of precise phase-coherent oscillators. Any errors in the oscillator frequencies would translate into inter carrier interference as described in the following sections. This has been a major issue in the past for successful implementation of OFDM modulation. However, the Inverse Fast Fourier Transform (IFFT) presents a convenient method to achieve OFDM modulation for large N . The IFFT or FFT is just an efficient computational tool for implementing the Fourier synthesis as described in equation (2.2). They represent the discrete

Figure 2.3: OFDM Transmitter Using Conventional Oscillators

version of the OFDM modulation process.

$$s_n = \sqrt{\frac{1}{N_{fft}}} \sum_{k=0}^{N_{fft}-1} S_k e^{j2\pi \frac{k}{N_{fft}} n}, \quad n = 0, 1, \dots, N_{fft} - 1. \quad (2.6)$$

Comparing equations (2.2) and (2.6) it is evident that, s'_n s are just sample values of the OFDM time domain signal $s(t)$ at points $t = \frac{nT_s}{N_{fft}}$. Therefore for s'_n s to represent $s(t)$ accurately N_{fft} has to be taken high enough. Also, looking at the IFFT equation in (2.6), the k in the summation takes values from 0 to $N_{fft} - 1$, but the symbols S'_k s are not defined for $k \geq N$, this problem is solved by putting 0's for those positions of k ; i.e $S_k = 0$, $k \geq N$.

As can be seen from Figure 2.2 the theoretical bandwidth of the Sinc spectrum (raised cosine with $\alpha = 0$) is unlimited. However, for practical purposes the total bandwidth of the signal $s(t)$ with N subcarriers can be approximately given by

$$B_s \approx \frac{N}{T_s} + C \quad (2.7)$$

where C is a constant which represents the edge effects due to the non vanishing Sinc spectrum. Hence, a system having larger values for N (number of subcarriers) would have a better spectral efficiency.

Figure 2.4: OFDM Transmitter IFFT Based

From Nyquist criterion, for s_n to represent $s(t)$, the sampling frequency should be higher than twice the bandwidth of the signal. Hence $N_{fft} \geq 2N$. For efficient implementation of the IFFT/FFT algorithm (used in DSP) the N_{fft} is taken as a power of 2. In practice to ease the burden on the low pass filter after digital to analog conversion (DAC), the N_{fft} can be taken higher values to perform oversampling. At the receiver, an FFT is performed on the received signal to recover the transmitted QAM symbols S'_k s. Taking $N_{fft} = N_{ifft}$ the FFT equation can be given by,

$$S_k = \sqrt{\frac{1}{N_{fft}}} \sum_{n=0}^{N_{fft}-1} s_n e^{-j2\pi \frac{k}{N_{fft}} n}, \quad k = 0, 1, \dots, N_{fft} - 1. \quad (2.8)$$

In case of oversampling as mentioned previously, the S_k values for $k \geq N$ (which will be 0) are simply discarded.

Cyclic Prefix

As discussed above, for the transmitted QAM symbols S'_k s to be received correctly at the receiver, the orthogonal property between the base pulses of the original OFDM signal and the received OFDM signal has to be maintained. For instance if there is a time delay of τ between the transmitter and the receiver, then considering an OFDM signal with two

Figure 2.5: OFDM Receiver FFT Based

subcarriers as in (2.3), the received signal is $s(t - \tau)$. Then the received symbol $S_{1,received}$ would be,

$$\begin{aligned}
 S_{1,received} &= \sqrt{\frac{1}{T_s}} \int_0^{T_s} s(t - \tau) e^{-j2\pi \frac{1}{T_s} t} dt \\
 &= \frac{1}{T_s} \int_{\tau}^{T_s} S_1 e^{-j2\pi \frac{1}{T_s} \tau} dt + \frac{1}{T_s} \int_{\tau}^{T_s} S_2 e^{-j2\pi \frac{2}{T_s} \tau} e^{j2\pi \frac{1}{T_s} t} dt \\
 &= \frac{T_s - \tau}{T_s} S_1 e^{-j2\pi \frac{1}{T_s} \tau} + \frac{e^{j2\pi} - e^{j2\pi \frac{1}{T_s} \tau}}{j2\pi \frac{1}{T_s}} S_2 e^{-j2\pi \frac{2}{T_s} \tau}
 \end{aligned} \tag{2.9}$$

The second term in the equation represents the interference from other subcarriers and popularly known as inter-carrier-interference (ICI) in literature. The ICI can be quite significant when the number of subcarriers is large. Another potential scenario is the interference from the delayed version of an OFDM symbol sent in a previous time slot ($s(t + T_s)$) as shown in Figure 2.6(b). In this case the equation for $S_{1,received}$ would be,

$$\begin{aligned}
 S_{1,received} &= \sqrt{\frac{1}{T_s}} \int_0^{T_s} s(t + T_s - \tau) e^{-j2\pi \frac{1}{T_s} t} dt + \sqrt{\frac{1}{T_s}} \int_0^{T_s} s(t - \tau) e^{-j2\pi \frac{1}{T_s} t} dt \\
 &= \sqrt{\frac{1}{T_s}} \int_0^{\tau} s(t + T_s - \tau) e^{-j2\pi \frac{1}{T_s} t} dt + \sqrt{\frac{1}{T_s}} \int_{\tau}^{T_s} s(t - \tau) e^{-j2\pi \frac{1}{T_s} t} dt
 \end{aligned} \tag{2.10}$$

Figure 2.6: Loss of orthogonality

There will be interference from all the subcarriers in the previous symbol $s(t + T_s - \tau)$ in addition to the ICI from the current symbol $s(t - \tau)$. The interference from the previous symbol is commonly called as inter-symbol-interference (ISI) or inter-OFDM-symbol-interference.

In a typical multipath scenario, inter-carrier-interference and inter-OFDM-symbol-interference can result corresponding to loss of orthogonality in frequency domain and time domain respectively. This illustrates the need for strict timing synchronization between the transmitter and receiver which unfortunately is very difficult to maintain in a frequency selective multipath fading channel, because every echo component of the signal is a poorly synchronized signal. The idea of the cyclic prefix is to introduce a guard interval in front of the transmitted OFDM symbol as a buffer to absorb the effects of timing mismatches. As shown in Figure 2.7 a piece of length Δ from the end of the symbol is copied and pasted in front of the signal. This increases the OFDM symbol duration from $t \in [0, T_s]$ to $t \in [-\Delta, T_s]$ and equation (2.2) is now modified to ,

$$s(t) = \sqrt{\frac{1}{T'_s}} \sum_{k=-N/2}^{N/2} S_k e^{j2\pi \frac{k}{T'_s} t} \Pi\left(\frac{t + \Delta}{T'_s} - \frac{1}{2}\right), \quad t \in [-\Delta, T_s] \quad (2.11)$$

where, $T'_s = T_s + \Delta$ is the new symbol duration and $\sqrt{\frac{1}{T'_s}}$ is a scaling factor to make the energy same as in (2.2). At the receiver the Fourier analysis is performed every T'_s seconds

Figure 2.7: Cyclic Prefix Insertion

with a Fourier window of T_s . Performing a similar analysis to (2.9) and (2.10) would reveal that as long as the delay is less than the cyclic prefix ($\tau \leq \Delta$) there would not be any inter-OFDM-symbol-interference or inter-carrier-interference. In this case,

$$S_{1,received} = \sqrt{\frac{T_s}{T'_s}} S_1 e^{-j2\pi \frac{1}{T'_s} \tau} \quad (2.12)$$

Thus the transmit symbol S_1 is recovered without any ISI or ICI but only rotated by a phase factor that depends on the subcarrier frequency ($\frac{k}{T_s}$) and the delay τ . For differential demodulation the phase is cancelled out and for coherent demodulation it needs to be determined by the channel estimation.

The cyclic prefix can also be used to determine the starting point of an OFDM signal without using any dedicated timing signals. Since the start and end of a cyclically extended OFDM signal is same, by a sliding correlation analysis on the received signals one can determine the starting point of the signal [3].

The drawbacks of adding a cyclic prefix is that the received signal energy will be reduced

by a factor of $\sqrt{\frac{T_s}{T_g}}$ as seen in equation (2.12). Reduced data rates (because of the increase in symbol duration) would be another disadvantage.

Transmission of OFDM with carrier frequency

After modulation by IFFT and adding the cyclic prefix, the baseband signal is up converted to passband by the carrier frequency f_c .

$$s_{RF}(t) = \text{Re}\{\sqrt{2}e^{j2\pi f_c t}s(t)\}, \quad t \in [0, T_s] \quad (2.13)$$

For implementation purposes, the DC component is left empty, that is only subcarriers at $k = \pm 1, \pm 2, \dots, \pm N/2$ are used. The received RF signal can be given by,

$$r_{RF}(t) = s_{RF}(t) * h(t) + n(t), \quad t \in [0, T_s] \quad (2.14)$$

where $h(t)$ is the channel impulse response and $n(t)$ is the noise process. At the receiver, the

Figure 2.8: RF Transmission of OFDM signal

signal is down converted to baseband and an FFT is performed to recover the transmitted symbols on each subcarrier. If the channel time delays of multiple paths do not exceed the cyclic prefix interval and neglecting the time variance of the channel, then by a similar analysis as in (2.12), it can be shown that,

$$\begin{aligned} S_{k, \text{received}} &= \sqrt{\frac{T_s}{T_g}} H_k S_k + n_k \\ H_k &= \sum_{i=1}^L c_i e^{-j2\pi \frac{k}{T_s} \tau_i} \end{aligned} \quad (2.15)$$

where, c_i and τ_i are the complex channel gain and path delay of the i^{th} echo respectively and n_k is the projection of the AWGN into the k^{th} subcarrier frequency. As can be noted from the equation the transmitted symbols $S'_k s$ in different subcarriers undergo a flat gain which is different across different frequencies for one OFDM symbol. The equalization for the received symbols can be performed in frequency domain by sending periodic training sequences. Therefore, the SNR at the receiver for the k^{th} subcarrier can be given by [3],

$$SNR = \frac{E[|\sqrt{\frac{T_s}{T_s}} S_k H_k|^2]}{E[|n_k|^2]} \quad (2.16)$$

2.2 Advantages of OFDM

- Since each subcarrier is narrowband a highly frequency selective fading channel can be considered as a flat fading channel with respect to each subcarrier as long as the multipath delay spread is less than the cyclic prefix length as seen in (2.15). Hence, the transmit symbol can be recovered without ISI, but only rotated by a phase factor. This phase factor is cancelled out for differential demodulation and for coherent demodulation it has to be determined by channel estimation, and a simple frequency domain equalization can be performed.
- OFDM systems are inherently resistant to narrowband interference. Interference in certain frequencies can be easily overcome by avoiding transmission on those subcarriers. Techniques such as adaptive modulation or forward error correction (FEC) are efficient methods to overcome this problem.
- OFDM systems are more bandwidth efficient than traditional FDM systems since the subcarrier bands overlap with each other. With large number of subcarriers, the edge effects due to the non vanishing Sinc spectrum becomes less significant as seen in (2.7).
- In addition to being a transmission scheme, it can be considered as a multiple access scheme in that different user data or different types of data can be transmitted via different subcarriers. At the receiver one user only needs to demodulate the subcarriers allocated to him. The channel behaviour on different frequencies as seen by each user is different and hence adaptive subcarrier allocation algorithms can be designed to maximize the total throughput.

- Flexibility is perhaps the most attractive feature of OFDM systems as this gives the ability to comply with worldwide regulations and the ability to coexist with current and future systems; the subcarriers can be dynamically turned on/off to comply with changing regulations and with different types of devices and services. OFDM data rates can be easily varied depending on the channel conditions, QOS requirements by means of adaptive modulation (adaptive bit, power loading). Complex MIMO systems which make use of Space-Time Block Codes (STBC) and Space-Frequency Block Codes (SFBC) and multi-carrier CDMA systems can provide very high data rates making use of the best of each such system. That is, With OFDM systems we have the flexibility to perform space-time-frequency signal processing.

2.3 Issues in OFDM

- As seen in Figure 2.3 the inphase and quadrature components of $s(t)$ are superpositions of many sinusoids with random phases and hence they show a Gaussian distribution. Hence there can be a large peak to average ratio, and there is a requirement for power amplifiers to have a large input backoff (IBO) to operate in the linear region. This reduces the power efficiency of the RF amplifier. Smaller IBO could cause nonlinear amplification to occur at peak values resulting in mixing products between all the subcarriers. This leads to loss of orthogonality of the subcarriers resulting in distortion. In addition to distortion there will be severe out of band radiation as a result of the product mixes between subcarriers. In modern standards there is stringent requirements on the out of band radiation and this could be a major hindrance in satisfying the standards requirements. With intensity modulated optical OFDM systems which uses a fixed bias signal to modulate the Laser/LED, there is a requirement to have a high bias signal in order to achieve the correct biasing points at all times due to the high variation of the OFDM signal. This translates into increased average power requirement.
- Lack of frequency diversity is another issue in OFDM systems. Frequency selective impairments exist in many different environments and systems which use or propose to use OFDM. WiMax and 4G systems are commonly affected by frequency selective fading, frequency dependent attenuations are seen in discrete-multi-tone (DMT) digital

Figure 2.9: OFDM time domain signal

subscriber loop (DSL) systems, frequency dependent interference (eg: fluorescent noise interference) affects indoor optical wireless systems and narrowband jamming is also a common problem in many of the current wired/wireless systems. The effects of non linear distortion also affects different subcarriers differently (for one OFDM symbol). For instance, channel nulls ($H_k = 0$ in (2.15)) in certain frequencies would render symbols sent over those subcarriers undetectable.

- Sensitivity to timing and frequency synchronization is another issue. The frequency and time dispersiveness of the channel can cause loss of orthogonality between the subcarriers. Timing synchronization is important to avoid ISI due to multipath propagation. If the delays due to multipath propagation exceed the cyclic prefix delay then again there will be inter-OFDM-symbol-interference and inter-carrier-interference as discussed in Section 2.1. On the other hand, having an excessive guard interval would decrease the power and bandwidth efficiency.

Since each subcarrier is narrowband ($B_{subcarrier} = \frac{1}{T_s}$) tight frequency synchronization is needed to avoid ICI arising because of energy spill overs from adjacent bands. For instance if there is a small frequency offset due to channel variations and errors in the frequency tracking mechanism, then after down conversion the complex base band

Figure 2.10: Histogram of $s(t)$

received signal would be,

$$r(t) = e^{j2\pi\delta ft} s(t) \quad (2.17)$$

Considering the two subcarrier system as in (2.3) the received symbol $S_{1,received}$ is,

$$\begin{aligned} S_{1,received} &= \sqrt{\frac{1}{T_s}} \int_0^{T_s} r(t) e^{-j2\pi\frac{1}{T_s}t} dt = \frac{1}{T_s} \int_0^{T_s} s(t) e^{j2\pi\delta ft} e^{-j2\pi\frac{1}{T_s}t} dt \\ &= \frac{1}{T_s} \int_0^{T_s} S_1 e^{j2\pi\delta ft} dt + \frac{1}{T_s} \int_0^{T_s} S_2 e^{j2\pi(\frac{1}{T_s} + \delta f)t} dt \\ &= \frac{j(1 - e^{j2\pi\delta f T_s})}{2\pi\delta f T_s} S_1 + \frac{j(1 - e^{j2\pi(1 + \delta f T_s)})}{2\pi(1 + \delta f T_s)} S_2 \\ &= \frac{j(1 - e^{j2\pi\delta f T_s})}{2\pi\delta f T_s} S_1 + \frac{j(1 - e^{j2\pi\delta f T_s})}{2\pi(1 + \delta f T_s)} S_2 \end{aligned} \quad (2.18)$$

There is a reduction in amplitude in the received symbol $S_{1,received}$ and a phase rotation in addition to the interference term from S_2 . The phase rotation is time dependent with a change in phase of $e^{j2\pi\delta f T_s}$ between two adjacent OFDM symbols (i.e symbols between $[0, T_s]$ and $[T_s, 2T_s]$ etc). For systems with large number of subcarriers there will be more interference terms from the other subcarriers (ICI). For small frequency offsets

($\delta f \cdot T_s \ll 1$), the first term in (2.18) is dominant and the other terms can be regarded as additive disturbance [3]. Figure 2.11 illustrates this result viewed in frequency domain. In a typical mobile radio channel, the frequency synchronization problem is

Figure 2.11: Conceptual diagram of reduction in amplitude plus ICI due to energy spill overs from adjacent bands caused by frequency offset.

made worse with Doppler shifts which also introduce a frequency modulation. Hence, the received signal ($r(t)$) is a superposition of several signals with different frequency offsets (and associated spectrum shifts) and the received symbols undergo significant ICI.

- Significant out of band radiation in OFDM systems due to the Sinc shape of the spectrum is another issue. To satisfy the spectrum masks of different standards, the OFDM signal after DAC is sent through a LPF which removes the edge effects of the OFDM spectrum (side lobes). Filtering the side lobes degrades the quality of the signal as the side lobes are also part of the useful OFDM signal. In time domain, the rectangular signal will be smoothened and broadened by the filter impulse response. Typically the ISI due to filtering can be absorbed by the cyclic prefix, but this will reduce the effective length of the guard interval to counter ISI due to the channel impulse response. In some systems, no useful information is sent over subcarriers at

the edges which are the ones mostly affected by filtering. However, this reduces the bandwidth efficiency.

The low pass filtering is also required to filter out the periodic aliasing spectra occurring at the sampling frequency (Figure 2.12). Due to the slow decaying OFDM spectrum and to ease the requirement on the steepness of the LPF filter slope the we need to have the aliasing spectra at a far enough frequency position. This can be accomplished by oversampling; by using a higher IFFT length N_{ifft} . Oversampling is essentially transferring complexity from the analog part to the digital part of the system.

Figure 2.12: Digital to analog conversion and low pass filtering.

- Increased hardware complexity at the transmitter and receiver to perform IFFT and FFT on thousands of subcarriers is also a concern. As discussed above, many practical systems use a large number of subcarriers to increase the spectral efficiency and when oversampling is also used, the length of IFFT to be performed is significantly increased. This results in increased hardware complexity and power consumption.

Chapter 3

Literature Survey

As mentioned in the previous chapters, the nonlinear distortion of signal as a result of the high PAPR, the increased average power requirement in intensity modulated optical systems and lack of frequency diversity are some of the critical issues that affect the widespread adoption of OFDM in future wireless systems and there have been a number of solutions proposed in literature. While it is beyond the scope of this thesis to review all the literature available with regard to these issues, we attempt to review some recent research which may relate directly with the work presented in this thesis.

Section 3.1 is a literature survey on the various peak to average power reduction techniques proposed for wireless OFDM systems and average power reduction techniques for optical OFDM systems. In section 3.2, several methods to improve the resilience of OFDM systems to frequency selective fading are evaluated.

3.1 Peak to average power reduction techniques

In [8] interleaving of data in an OFDM symbol is proposed. This method is based on the fact that an OFDM symbol with a highly correlated data pattern will have a large PAPR. By breaking the long correlation patterns using pseudo-random bit or symbol interleaving the PAPR can be reduced. The symbol or bit interleaving is achieved by using a number of interleavers (K) at the transmitter side and selecting the symbol (or bit) sequence which produces the lowest PAPR after IFFT. The information ($\log_2 K$) on which interleaver is used is sent after employing suitable error correction. The authors in [8] have also proposed an adaptive interleaving scheme which uses an early terminating threshold. In this

method the PAPR of the time domain signal after IFFT corresponding to the interleaved sequence is calculated starting from the first interleaver. If it is below a certain threshold the search is stopped. The search continues until an interleaved sequence which satisfies the threshold is found. If all the K interleaved sequences do not satisfy the threshold then the sequence with the minimum PAPR is chosen and transmitted with the corresponding side information. It has also been shown that bit interleaving gives a better performance than symbol interleaving although there may be an increase in complexity. The computational cost of implementing such a solution would be K IFFT operations (a constant) per one OFDM symbol. The concept of this technique is simple and intuitive and can be easily implemented in practical systems. However, with this method significant PAPR may not be realizable without increasing the number of interleavers used considerably. Even then, the PAPR reductions would be limited by the natural limit that can be achieved by an optimum interleaving pattern.

The authors in [21] have proposed an M-ZPSK modulation scheme, which includes a signal point of zero amplitude. In this method one out of M bit patterns of $\log_2 M$ bits will be mapped to a constellation of zero amplitude. This results in the generation of M independent OFDM symbols, out of which the one with the lowest PAPR is selected and transmitted with side information containing the mapping used. The cost of implementing this solution is M IFFT operations per one OFDM symbol. It has also been shown that this adaptive M-ZPSK modulation scheme reduces the sensitivity to frequency offset errors. The underlying idea in the methods proposed in [21] and [8] is same in that both methods attempt to decorrelate the data (or introduce 'randomness') and thereby reducing the PAPR of the OFDM symbol. Hence, in this method also the achievable PAPR reductions are limited by the modulation scheme M , as it determines the number of independent OFDM symbols generated for a given data.

In [22], [9] the authors have introduced a novel method for reducing the PAPR of OFDM systems known popularly as the Tone Reservation Technique. In this method the PAPR is reduced by adding time domain signals to the original OFDM signal (after IFFT). These 'anti peak signals' are created depending on the OFDM time domain signal so that the PAPR of the composite signal after addition is minimized. This artificial time domain signal can be generated by reserving certain subcarriers out of the pool of all available subcarriers. Since each subcarrier of an OFDM system occupies a disjoint frequency space this addition is distortionless and leads to simple recovery of the data symbols at the receiver by discarding

the data symbols received in those subcarriers which are 'reserved'. Since the amplitudes of the tones which are reserved for PAPR reduction need not be selected according to the signal constellation, there is freedom to select the amplitudes of these tones in order to reduce the fluctuations of the OFDM transmit signal from the average level. In this work it is derived that, optimizing the time domain 'anti peak' signal leads to a convex optimization problem that can be transformed into a linear program. The complexity of the optimization problem can be reduced when the symmetrical nature of the IFFT matrix is taken into account. The drawbacks of these methods are the computational cost and the decrease in bandwidth efficiency. The complexity of the linear programming problem can be further reduced using simple gradient algorithms without significant sacrifice in performance. There have been numerous variations reported in literature which employ the same basic concept as described above. This technique is especially attractive in systems with feedback loops as subcarriers with low SNRs can be used as the peak reduction tones, resulting in improved bandwidth efficiency.

An interesting approach to contain the level of OFDM signal is reported in [23], where the authors propose the use of simplex modulation to achieve a constant OFDM signal level after IFFT. Simplex signals are based on the principle that in an n -dimensional Euclidean space we can have $n + 1$ signal points that are equally spaced. Therefore a 4-dimensional constellation which contains 4 signal points can be reduced to a 3-dimensional constellation with 4 signal points each of which are separated by the same distance as they had originally in the 4-d case, Figure 3.1. Therefore in the case of a QPSK constellation, it can be first represented in 4-d and then the dimensions can be further reduced to 3 using simplex principle. But, now the signal level required for each constellation point is reduced as illustrated in the diagram. However, this results in a bandwidth expansion factor of $\frac{3}{2}$. In OFDM systems, the 3 dimensions can be represented by three orthogonal carriers-a cosine and a sine of the same frequency and another cosine waveform of a different orthogonal frequency. In our work we have evaluated the performance of this approach. The modulation of the three orthogonal signals (cosine, sine and cosine of another orthogonal frequency) can be easily implemented in IFFT as described in Chapter 4 and the performance is analyzed in Section 5.2. The simplex symbol approach is especially attractive in optical OFDM systems where nonlinearity takes precedence over bandwidth. However, this approach is not effective at higher modulation schemes (eg:16-QAM,64-QAM) as the reduction in dimension due to the simplex approach is insignificant when n is high. Moreover, increased n will result in the

need for more orthogonal axis to represent one symbol and the loss in bandwidth efficiency outweighs the benefit due to a constant signal level.

Figure 3.1: Signal spaces of simplex signals

Hence, it is evident that simplex signals cannot be used to replace higher modulation schemes. For example, in order to use it instead of 16-QAM we need to generate 15 orthogonal axis (or equivalently 15 orthogonal subcarriers) to represent one symbol and hence it is not feasible in practical systems. However, this is ideally suited for a QPSK system in which to represent one symbol using simplex signals we need only three orthogonal axis as shown in Figure 3.1. If S_1, S_2, S_3, S_4 are the 4 possible symbols of the QPSK constellation, they are represented in a 2-d constellation by,

$$S_1 = [-1, -1], S_2 = [-1, 1], S_3 = [1, -1], S_4 = [1, 1]$$

transforming this into a 4-d constellation while maintaining the distance between adjacent symbols constant,

$$S_1 = [\sqrt{2}, 0, 0, 0], S_2 = [0, \sqrt{2}, 0, 0], S_3 = [0, 0, \sqrt{2}, 0], S_4 = [0, 0, 0, \sqrt{2}]$$

using the simplex transformation,

$$\hat{S}_m = S_m - \frac{1}{4} \sum_{m=1}^4 S_m, \quad m = 1, 2, 3, 4$$

Now the symbols $\hat{S}_1, \hat{S}_2, \hat{S}_3, \hat{S}_4$ can be expressed in a 3-d constellation where the three orthogonal base axes can be derived by Graham-Schmidt orthogonalization procedure [24]. Then the new symbol constellation on 3-d would be $[x_1, y_1, z_1], [x_2, y_2, z_2], [x_3, y_3, z_3], [x_4, y_4, z_4]$ where $[x_i, y_i, z_i]$ is the projection of $\hat{S}_i$ onto the orthogonal axis x, y, z found by Graham-Schmidt procedure.

In traditional OFDM systems to modulate a complex symbol we need a sine carrier and cosine carrier for the inphase and quadrature phase components as shown in Figure 2.3. For the 3-d constellation derived above, three carriers (a cosine and sine of the same frequency and a cosine carrier of another frequency) are used for one symbol - considering a symbol sequence of

$$[S_1, S_2] = [x_1 \ y_1 \ z_1, x_2 \ y_2 \ z_2] \quad (3.1)$$

then the OFDM time domain signal $s(t)$ is,

$$s(t) = x_1 \cos(2\pi \frac{1}{T_s} t) + y_1 \sin(2\pi \frac{1}{T_s} t) + z_1 \cos(2\pi \frac{2}{T_s} t) + \\ x_2 \sin(2\pi \frac{2}{T_s} t) + y_2 \cos(2\pi \frac{3}{T_s} t) + z_2 \sin(2\pi \frac{3}{T_s} t) \quad (3.2)$$

To implement the above equation using IFFT the amplitudes are taken two at a time and combined as, $[x_1 + jy_1, z_1 + jy_2, y_2 + jy_3]$ and then an IFFT is performed on the sequence. In the receiver after FFT, the amplitude values are reorganized to produce the original transmitted symbols. This concept can be extended for any number of symbols. From (3.2) it is clear that for QPSK the bandwidth expansion factor due to simplex signals would be $\frac{3}{2}$.

A slight improvement of the technique would be to assign the mapping of the axis adaptively so that after IFFT the constellation with the lowest PAPR is chosen. For instance, corresponding to the sequence $[S_1, S_2]$, the modified simplex sequence would be $[z_1 \ y_1 \ x_1, z_2 \ y_2 \ x_2]$. In Section 5.2.2 the results of a normal QAM, the simplex technique and the modified simplex technique are presented. In addition to the reduction in PAPR, the simplex technique also results in increased diversity as seen by (3.2).

In the paper [25] the authors illustrate that if information about the nonlinearity of the channel or the HPA/Laser is available at the transmitter side, then a better performance can be achieved by directly minimizing the distortion as a result of the nonlinearity rather than minimizing the OFDM signal's PAPR. In this work, the goal is to minimize the difference

between the original OFDM time domain signal $x(t)$ and the estimated distorted time domain signal $x_{\text{distorted}}(t)$ after nonlinearity-i.e a minimum mean square error (MMSE) criterion is applied as the performance measure instead of minimum PAPR criterion. The nonlinearity is estimated at the transmitter side using an adaptive nonlinear system identification process. In this scheme since only the criterion to measure the effect of nonlinearity is changed, all existing PAPR reduction methods based on minimum PAPR criterion can be easily extended to incorporate the new scheme. The authors have done a case study using selected mapping (SLM) method with the new scheme and with the conventional scheme and demonstrated the superiority in performance of the proposed scheme. However, the drawbacks of this scheme are, increased complexity associated with nonlinearity estimation at the transmitter side, errors in the estimation of nonlinearity may affect performance and the requirement to perform multiple IFFTs as part of the distortion mitigation process. This approach is especially useful if the nonlinearity is relatively stable and can be estimated with sufficient accuracy as in the case of MZM/Laser nonlinearity in radio-over-fiber systems. In the work presented in Section 4.1 we describe a simple adaptive modulation technique for nonlinear distortion mitigation in optical OFDM systems, with suitable modifications to the above described scheme.

In recent times there has been considerable research in the use of OFDM in optical fiber communication systems as well, as described in Chapter 1. OFDM Radio-over-fiber (ROF) and coherent OFDM optical systems are considered viable solutions to the next generation high speed backbone networks. Also in [26] extensive analysis is done on the performance of OFDM system in indoor wireless optical channels (IEEE 802.11 infrared option) and an adaptive system, which uses channel equalization based on sending initial training sequences to estimate the channel response is proposed. It has also been shown that OFDM systems provide immunity against fluorescent-light noise near dc, which is the main source of interference in indoor optical wireless systems [27]. The main drawback of OFDM with Intensity Modulation and Direct Detection (IM/DD) is poor optical average power efficiency. The OFDM electrical signal is a sum of modulated sinusoids, and thus can take negative and positive values. However, the optical intensity (instantaneous optical power) must be non-negative, and hence a dc bias must be added to the OFDM electrical signal. As the number of subcarriers increases, the minimum value of the OFDM electrical signal decreases (becomes more negative), and the required dc bias increases. Since the average optical power is proportional to this dc bias, the optical average-power efficiency worsens as the number

of subcarriers increases. The authors in [27] have proposed two classes of techniques for optical wireless systems, a block coding and a variable bias technique. In the first technique, block coding is applied between the information bits to be transmitted and the symbol amplitudes to be modulated onto the subcarriers, in order to increase the minimum value of the OFDM electrical signal. In the second technique, the fixed dc bias is replaced by a bias signal that varies on a symbol-by-symbol basis. However, the efficient implementation of the proposed techniques becomes impractical when the number of subcarriers becomes large. In Section 4.2 we suitably modify the Tone Reservation technique reported in [22], in order to increase the minimum value of the OFDM electrical signal, thereby reducing the average power required; i.e dummy symbols are sent in certain reserved subcarriers (distributed over the frequency band) in order to minimize the required dc bias.

3.2 Frequency selective fading resistant techniques

As mentioned earlier, one of the disadvantages of OFDM is the lack of frequency diversity—since each symbol is transmitted over a single subchannel in a narrow frequency band independently. As a result, the symbols carried by the subchannels experiencing a high channel attenuation (e.g channel nulls) are usually not accurately decoded in the receiver. In order to combat the bad subchannels in OFDM some notion of frequency diversity should be introduced, which effectively distributes the individual symbol information across a number of subcarriers rather than a single subcarrier.

OFDM spread spectrum (OFDM-SS) systems are proposed to increase the resilience to frequency selectivity, narrowband interference and jamming [28, 29]. In OFDM-SS systems spectral spreading is accomplished by putting the same data on all the carriers, producing a spreading factor equal to the number of carriers. At the receiver, the energy from all the carriers is coherently combined to produce the decision variable [28]. The use of spreading tends to limit the available throughput due to the bandwidth expansion. Rate Adaptive OFDM (RA-OFDM) [30] is a compromise between the conventional OFDM and spread spectrum OFDM, where the amount of spreading (processing gain) is adjusted depending on the signal quality—i.e a large processing gain will be used when the SNR is low and a small processing gain will be used when the SNR is high. Hence, with RA-OFDM the desired system performance can be maintained across a large dynamic range of signal strengths (channel conditions) by controlling the spreading.

The use of error-correcting codes (FEC) across OFDM subcarriers is also commonly used for achieving frequency diversity in OFDM systems, and popularly known as Coded-OFDM (COFDM) and used in DVB-T and WLAN standards [15, 31]. In COFDM systems the codewords are made to span across several subcarriers. The FEC techniques include convolutional codes and trellis-coded modulation. Trellis-coded modulation together with interleaving enables a better tradeoff between performance and bandwidth efficiency, while enjoying low-complexity Viterbi decoding. However, standard design paradigms for TCM make it difficult to design systems which achieve diversity gain equal to the code length [32].

In UMTS LTE, the symbols to be transmitted are precoded (in frequency domain) before being mapped onto subcarriers. The symbol precoding results in averaging out the power over all subcarriers and hence achieves a low crest factor; it also improves the diversity of the OFDM system. The precoding is achieved by a Discrete Fourier Transform (DFT) before subcarrier mapping and such systems are known as 'DFT spread OFDM' systems [2]. A novel grouped linear constellation precoded OFDM (GLCP-OFDM) system is proposed in [32]. In this system, first the available subcarriers are grouped into different subsets depending upon their correlation. Subchannels with independent channel gains are grouped together in one subset. Within each subset a linear constellation precoding is designed to maximize both diversity and coding gain. The proposed system is shown capable of achieving the maximum multipath diversity, large coding gains, and guaranteeing symbol detectability with low decoding complexity. Since, in GLCP-OFDM, symbol precoding is performed on subgroups of OFDM subcarriers rather than on the total number of OFDM subcarriers, the complexity of decoding is decreased. However, the error performance of this system is worse than non-precoded systems at low-to-medium SNRs due to one or more poor subcarriers affecting the entire decoding process. The authors in [15] have proposed precoding with adaptive subcarrier selection (GLCP-ACS-OFDM) as a solution to improve the performance of GLCP-OFDM. In GLCP-ACS-OFDM, at the receiver the SNR of each subcarrier is estimated and subcarriers with poor SNRs are discarded before maximum likelihood estimation or zero forcing estimation is done. This technique improves the error performance in the low-to-medium SNR change and also further reduces decoding complexity, since not all the subcarriers are used for detection. However, the disadvantage of this approach is that in the high SNR region the performance goes down as information on certain subcarriers are 'discarded'. Further extending this approach, in Section 4.3 we describe an adaptive weighting technique, which shows good error performance both in low and high SNR regimes for

MLSE. However, MLSE has exponential complexity and cannot be realized in practice.

In [33] the authors present a simplified maximum likelihood detection algorithm to eliminate the clipping distortion in the OFDM symbols due to the nonlinearity in the power amplifier. In this algorithm first a hard decision is made on the received symbols and a decision vector is formed; then all the possible values of the first symbol are considered (while keeping the other symbols unchanged) and different symbol sequences (vectors) are created; these vectors are then converted into a time domain signal using an IFFT and sent through the same nonlinear function (assuming it is known at the receiver) and the sequence which minimizes the distance from the original received (distorted) signal is selected; the value for the first symbol is thus found and fixed. Similarly, the value for the other successive symbols are found by fixing all the other symbols. This sequential mean square error reduction is suboptimum but it reduces the complexity from exponential to linear. This algorithm is able to progressively eliminate the nonlinear distortion in the received symbols because uncorrelated symbols during transmission become correlated at the receiver due to the nonlinearity. Hence minimizing the error in the time domain will result in reduced errors in the frequency domain as well. In GLCP-OFDM since the symbols mapped onto the subcarriers at the transmitter side are already correlated due to precoding, we can apply the same technique directly in frequency domain to improve the receiver performance; after the initial hard decision is obtained by zero forcing equalization, sequential mean square error reduction is performed subsequently on the recovered symbols. This methodology is described in Section 4.3.3 and the performance is evaluated in Section 5.4.3.

Chapter 4

Techniques

4.1 Adaptive Modulation for Nonlinear Distortion Mitigation

As mentioned in the previous sections, the nonlinear distortion noise caused by HPA/Laser and optical channel could be a major limiting factor in future high speed wired/wireless OFDM systems employing thousands of subcarriers (eg:1024, 2048, 4096, 8192). In the paper [25] the authors illustrate that if information about the nonlinearity of the channel or the HPA/Laser is available at the transmitter side, then a better performance can be achieved by directly minimizing the distortion as a result of the nonlinearity rather than minimizing the OFDM signal's PAPR. This approach is especially useful if the nonlinearity is relatively stable and can be estimated with sufficient accuracy as in the case of laser nonlinearity in optical communications systems. In this section we describe a simple algorithm that follows this concept assuming that information about the nonlinear behaviour is available at the transmitter side. The authors in [25] use an adaptive nonlinear system identification process which can be based on either look-up-table modeling or polynomial modeling approach. Then, for each OFDM symbol, its corresponding time domain signal $x(t)$ is sent through the estimated nonlinearity and an FFT is performed and the distortion on the symbols of each subcarrier is measured at the transmitter side as shown in Figure 4.1. Then the subcarriers which experience high distortion are estimated and the modulation level on those subcarriers is reduced and the procedure is repeated. When the distortion on all the subcarriers are below a certain threshold level, the OFDM symbol is sent along with side information containing

the modulation information of the subcarriers. This approach differs slightly from the criteria developed in [25] where the goal is to minimize the difference between the original OFDM time domain signal $x(t)$ and the estimated distorted time domain signal $x_{\text{distorted}}(t)$ after nonlinearity. However, in our approach we attempt to minimize the estimated distortion on each subcarrier (by taking an FFT after the estimated nonlinearity at the transmitter side as in Figure 4.1). This methodology attempts to give a better performance, but at the expense of increased complexity.

This adaptive modulation approach is made possible since the serial bits to be transmitted are converted parallelly before OFDM modulation-that is we have the freedom to choose the modulation level of each subcarrier without having to change the modulation level of other subcarriers from the pool of available parallel bits. Another observation is that since the modulation levels on the subcarriers are independent, a change in one subcarrier would not considerably affect the time domain signal $x(t)$ according to the IFFT equation - this means the distortion effects after nonlinearity also would be relatively same except for the subcarrier position whose modulation is changed. Therefore by identifying the subcarriers (frequency positions) that are more likely to have distortion and reducing their modulation level we can improve the BER performance for the symbols on those subcarriers.

Figure 4.1: Block schematic of the adaptive modulation system

With this algorithm, $x(t)$ is passed through the estimated nonlinearity and an FFT is taken and the distortion on all subcarriers is estimated. The subcarrier indexes whose

distortion level exceeds a predetermined threshold (d_{thr}) are determined and sorted according to descending order of distortion. Out of these high distortion subcarrier indexes, only the first *selection_length* number of indexes are chosen as the selected subcarrier indexes and added to the *solution_set* as described in Section 4.1.1. The modulation level of the entries in the *solution_set* are reduced to a lower order. We need not reduce the modulation level on all the subcarriers whose distortion exceed the threshold. This is because when the modulation level of the highest distortion subcarriers are reduced it would also bring down the distortion level of other subcarriers as well. If the improvement on distortion of the other subcarriers still falls below the threshold, this procedure is repeated and more subcarriers are added to the *solution_set* in the same manner. When all the subcarriers have acceptable distortion levels ($< d_{thr}$) the algorithm exits and the subcarrier indexes collected in *solution_set* are sent along with the OFDM symbol as side information (SI) with suitable error protection. It can be observed that this approach is essentially trading off bandwidth (or power) in order to improve the BER performance on the face of nonlinearity. However, by the suitable selection of d_{thr} (according to the external noise processes) and *fraction* (according to the granularity desired) we can ensure that the desired BER performance is achieved with minimum compromise in bandwidth and complexity.

4.1.1 Algorithm

Initial Settings

initial modulation level for all subcarriers

adaptive_subcarrier_info = $\{M_0, M_1, \dots, M_{N-1}\}$; $M_k = \text{higher_modulation_order}, \forall_k$

threshold_distortion_level = d_{thr} , a predetermined constant

subcarrier_indexes_with_high_distortion = $\{\}$, index of subcarriers with distortion exceeding d_{thr}

fraction of the distorted subcarriers to be considered = *fraction*, a predetermined constant

selected_subcarrier_indexes = $\{\}$, the subcarrier indexes selected to be added to the solution set

solution_set = $\{\}$, solution set for each iteration

Step 1: Calculate the distortion of symbols in each of the subcarriers. The distortion vector

can be given as,

$$distortion = [d_0 \ d_2 \ \dots \ d_{N-1}]$$

If the distortion in all the subcarriers is below the threshold level,

$$d_k < d_{thr}, \text{ for } k \in \{0, 1, \dots, N-1\}$$

then exit the algorithm. Otherwise proceed to **Step 2**.

Step 2: Find the indexes of subcarriers which have distortion above the threshold level and sort them according to descending order of distortion.

$$subcarrier_indexes_with_high_distortion = \{i_0, i_1, \dots, \dots\}$$

$$\text{where } d_{i_k} \geq d_{i_{k+1}} \text{ and } d_{i_k} \geq d_{thr} \text{ for } k \in \{0, 1, \dots, N-1\}$$

Step 3: Divide the number of subcarriers in the sorted set by the predetermined length to find the number of subcarriers to be added to the solution set. Select the first *selection_length* subcarriers from *subcarrier_indexes_with_high_distortion* and make the *selected_subcarrier_indexes* list.

$$selection_length = \left\lfloor \frac{length(subcarrier_indexes_with_high_distortion)}{fraction} \right\rfloor$$

$$selected_subcarrier_indexes = \{i_0, i_1, \dots, i_{selection_length-1}\}$$

Step 4: Update the solution set with *selected_subcarrier_indexes*. Determine the adaptive subcarrier info for the next iteration based on the entries in the solution set.

Go to **Step 1**.

$$solution_set = solution_set \cup selected_subcarrier_indexes$$

$$adaptive_subcarrier_info = \{M_0, M_1, \dots, M_{N-1}\},$$

$$\text{where } M_k = \text{higher_modulation_order if } k \notin solution_set,$$

$$M_k = \text{lower_modulation_order if } k \in solution_set$$

4.2 Average Power Reduction Technique

4.2.1 Optical Wireless OFDM Systems

Indoor optical wireless (OW) channels with Intensity Modulation and Direct Detection (IM/DD) can be described by [27],

$$y(t) = rx(t) \otimes h(t) + n(t) \quad (4.1)$$

where $x(t)$ is the transmitted optical signal, $h(t)$ is the channel impulse response, $n(t)$ is the noise process, r is the photodetector responsivity and $y(t)$ is the received photocurrent. The symbol $\otimes$ represents the convolution operation. The channel input is the transmitted optical intensity ($x(t)$), which must be nonnegative.

$$x(t) \geq 0 \quad (4.2)$$

The channel output $y(t)$ is the received photocurrent. The channel is considered as a fixed, linear system having impulse response $h(t)$ and frequency response $H(jw)$.

The average optical power P is given by the mean value of $x(t)$ as,

$$P = E[x(t)] \quad (4.3)$$

in contrast to electrical channels, where the average power is the mean value of $x^2(t)$. In order to ensure that $x(t)$ is nonnegative, a bias signal needs to be added to the time domain electrical signal $s(t)$, i.e

$$x(t) = A[s(t) + b(t)] \quad (4.4)$$

where, A is a scale factor. Then average optical power (emitted from the LED) will be,

$$P = E[A(s(t) + b(t))] \quad (4.5)$$

If the input data $s(t)$ is random,

$$P = AE[b(t)] \quad (4.6)$$

Therefore, in order to minimize the average power we need to minimize the dc bias required. The block diagram of the wireless optical OFDM system is shown in Figure 4.2. First, $M \cdot P$

Figure 4.2: Optical OFDM System

data bits each of duration T_b is generated and modulated by $P 2^M - QAM$ modulators. Then, the DC offset value and pilots are added (L symbols) and $(N - L)$ 0s are added to the symbol sequence. The resulting sequence (N points) is also appended with its complex conjugate in the reverse order before computing the IFFT. The outgoing IFFT symbol is a real sequence of $2N$ points and it can drive directly the LED after adding an appropriate dc bias. A cyclic prefix is added before transmitting to combat ISI. The transmitted signal is distorted by channel response and corrupted by noise. The high frequency fluorescent lamp interference, preamplifier thermal noise and photodiode shot noise are the main noise processes in indoor optical wireless systems. The overall noise process is modeled as zero-mean additive white Gaussian noise (AWGN) with two sided power spectral density $\frac{N_0}{2}$. At the OFDM receiver, the corrupted cyclic prefix is removed; the FFT of $2N$ points is computed by the receiver. Only the first L points are considered and after removing the DC offset and pilots, the remaining P symbols demodulated and the bits are received.

4.2.2 Complex Conjugate Insertion

Unlike in radio wireless systems in optical systems we need real signals to modulate the LED. However, in baseband models the output after the IFFT (eg: QAM) will be a sequence of complex numbers. This problem is solved by augmenting the original sequence (D_n) by appending its complex conjugate to it [34] as shown in Figure 4.3. The $2N$ point IFFT of

this augmented sequence is then a sequence of $2N$ real numbers. The augmented sequence

Figure 4.3: System with real transmission

is formed from the original sequence D_n [24] as,

$$\begin{aligned} D'_n &= D_n, & n &= 1, 2, \dots, N-1 \\ &= D_{2N-n}^*, & n &= N+1, N+2, \dots, 2N-1 \end{aligned} \quad (4.7)$$

In order to contain conjugate symmetry, it is essential that D'_0 and D'_N be real. If the original D_0 is 0 (as in most baseband models), then D'_0 and D'_N are set to 0. Otherwise D'_0 may be set to $Re(D_0)$ and D'_N to $Im(D_0)$. For the case of $D_0 = 0$ the output of the IFFT is

$$\begin{aligned} d_m &= \sum_{n=0}^{2N-1} D'_n \exp\left(\frac{j2\pi mn}{N}\right) \\ &= 2Re\left\{ \sum_{n=0}^{2N-1} D_n \exp\left(\frac{j2\pi mn}{N}\right) \right\} \\ &= 2 \sum_{n=0}^{2N-1} \left[A_n \cos\left(\frac{2\pi mn}{N}\right) - B_n \sin\left(\frac{2\pi mn}{N}\right) \right], \quad m = 0, 1, \dots, 2N-1 \end{aligned} \quad (4.8)$$

where $D_n = A_n + jB_n$. The scaling factor of two is immaterial and can be dropped. This real orthogonal transformation is fully equivalent to the complex one, and all subsequent analyses are applicable.

4.2.3 Channel Model

There are several models used to estimate the impulse response of indoor optical wireless channels. In this work we use a statistical model developed by R.Perez-Jimenez [35] to characterize the channel. The statistical model closely approximates the channel response, which can be obtained by iterative solutions (Monte Carlo etc) but does not require computational

complexity. Here, the channel response is approximated by a Rayleigh function where the RMS delay spread (τ_{rms}) and mean excess delay (τ_m) are given by,

$$\begin{aligned}\tau_m(ns) &= -0.46n^{0.28} + 0.33n^{-0.17}d_{err} + K(n, d_{err}) \cos(0.018\theta - 0.012) \\ \tau_{rms}(ns) &= -0.82n^{0.03} + 0.58n^{-0.11}d_{err} + (-0.54 + 0.19d_{err}) \cos(0.019\theta - 0.32) \\ K(n, d_{err}) &= (-0.17n^{0.41} + 0.18n^{0.26}d_{err})\end{aligned}\quad (4.9)$$

where, n is the mode number and θ is the angle between the transmitter and receiver as shown in Figure 4.4.

Figure 4.4: Channel Parameters

4.2.4 Fluorescent Light Interference

The fluorescent light interference is the dominant noise source for diffuse optical systems. In our analysis the fluorescent light model developed by [36] is used. The interfering signal $i(t)$ is modeled with two sinusoids and their harmonics. The first sinusoid exists at 50 Hz, which represents the frequency of the power supply, and the second one at approximately 37.5 kHz and is due to the oscillation frequency of electronic lamp ballasts in fluorescent lighting. The high frequency interference harmonics exist into the MHz range. The fluorescent noise can be modeled as follows.

$$\begin{aligned}
i_{low}(t) &= \frac{I_B}{F_1 \cdot A_1} \sum_{i=1}^{20} [b_i \cos(2\pi(100i - 50)t + \varphi_i) + c_i \cos(2\pi 100it + \phi_i)] \\
i_{high}(t) &= \frac{I_B}{F_1 \cdot A_2} \sum_{j=1}^{n3} d_j \cos(2\pi f_{high}jt + \vartheta_j)
\end{aligned} \tag{4.10}$$

where $F_1 \approx 3$, $A_1 \approx 5.9$, $A_2 \approx 2.1$ and $f_{high} = 37.5kHz$. The values for b_i , c_i , φ_i , ϕ_i , and ϑ_j can be found in [36].

4.2.5 Tone Reservation for Average Power Reduction

In optical wireless OFDM systems there are typically a number of subchannels (tones) that cannot carry data reliably due to low SNRs. For example, typically the tones near dc are not utilized because of fluorescent noise interference. If these tones are not used for data transmission, they can be used to design a signal to offset any large peaks from the data-bearing subchannels. This principle was introduced to reduce the peak to average power ratio (PAPR) in radio frequency OFDM systems in [22] and [9] and was given the name tone reservation by the authors of [22]. Since the amplitudes of the tones which are reserved for PAPR reduction need not be selected according to the signal constellation, we have freedom to select the amplitudes of these tones in order to reduce the fluctuations of the OFDM transmit signal from the average level. It should be noted that since each subcarrier in the OFDM signal is orthogonal to each other, the reserved subcarriers which carry dummy symbols can be simply disregarded at the receiver. However, in this technique obviously the bandwidth efficiency (or the data rate) will be reduced as some subcarriers do not carry useful information and hence it is important to keep the number of reserved subcarriers as small as possible. In this work we employ this technique in the optical wireless (IM/DD) case where the requirement is to reduce the average optical power as opposed to reducing PAPR like in traditional wireless systems (i.e we have to minimize the fixed dc bias required to ensure the modulating signal of the LED is positive).

Let us denote the complex valued QAM vector input (frequency domain vector) to the IFFT block by,

$$\mathbf{X} = [X_0 \ X_1 \ X_2 \ \dots \ X_{N-1}]^T \tag{4.11}$$

Then the output of the IFFT (the OFDM transmit signal) will be the sum of N independent QAM sub signals each with equal bandwidth and frequency separation $\frac{1}{T_s}$, where T_s is the time duration of 1 OFDM symbol. Thus, the OFDM transmit signal (the time domain vector),

$$\mathbf{x} = [x_0 \ x_1 \ x_2 \ \dots \ x_{N-1}]^T \quad (4.12)$$

is generated with ,

$$x_n = \frac{1}{N} \sum_{k=0}^{N-1} X_k \exp\left(\frac{j2\pi kn}{N}\right), \quad n = 0, 1, \dots, N-1 \quad (4.13)$$

This can be written down in matrix form as, $\mathbf{x} = \mathbf{Q}\mathbf{X}$ where $\mathbf{Q}$ is the IFFT matrix with elements $q_{n,k} = \frac{1}{N} \exp\left(\frac{j2\pi kn}{N}\right)$. When N is large x_n can be modeled as a Gaussian random process with 0 mean. By adding structured time domain vectors to x_n , we can reduce the maximum (negative) value of the time domain signal $\mathbf{x}$ (and hence the fixed dc bias that is required to make $x(t)$ nonnegative will be reduced).

If we add a frequency domain vector,

$$\mathbf{C} = [C_0 \ C_1 \ C_2 \ \dots \ C_{N-1}]^T \quad (4.14)$$

to $\mathbf{X}$, the new time domain vector (after IFFT) will be

$$\begin{aligned} \mathbf{x} + \mathbf{c} &= IFFT(\mathbf{X} + \mathbf{C}) \\ &= \mathbf{Q}(\mathbf{X} + \mathbf{C}) \end{aligned} \quad (4.15)$$

where $\mathbf{c} = \mathbf{Q}\mathbf{C}$. Obviously the vectors $\mathbf{X}$ and $\mathbf{C}$ should have the same dimension for this to

Figure 4.5: Average Power Reduction Technique

be possible. Now, the peak reduction vector and the data-bearing vector are made to lie in

disjoint frequency spaces. If there are L unused subchannels whose indexes form the set I_u then,

$$\begin{aligned} \mathbf{X} &= [X_0 \ X_1 \ X_2 \ \dots \ X_{N-1}], \text{ where } X_k = 0, \text{ when } k \in I_u \\ \mathbf{C} &= [C_0 \ C_1 \ C_2 \ \dots \ C_{N-1}], \text{ where } C_k = 0, \text{ when } k \notin I_u \end{aligned} \quad (4.16)$$

This formulation is distortionless and leads to very simple decoding of the data symbols which are extracted from the received sequence by choosing the set of values at the output of the FFT at the receiver. Also it allows simple optimization techniques for the computation of the peak reduction vector $\mathbf{C}$. The L nonzero values in $\mathbf{C}$ are called the peak reduction tones (PRT).

Now the problem is to find the value of the time domain vector $\mathbf{c}$ (or the equivalent frequency domain vector $\mathbf{C}$) in order to maximize the minimum value of the composite signal $\mathbf{x} + \mathbf{c}$ (ideally bring it very close to 0). Expressed mathematically,

$$\begin{aligned} \mathbf{c}^* &= \arg \max_{\mathbf{c}} (\min(\mathbf{x} + \mathbf{c})) \\ &= \arg \max_{\mathbf{C}} (\min(\mathbf{x} + \mathbf{Q}\mathbf{C})) \end{aligned} \quad (4.17)$$

where $\mathbf{c}^*$ is the optimum time domain vector. The problem can be further simplified by considering $\hat{\mathbf{C}} = [C_{i_1} \ C_{i_2} \ C_{i_3} \ \dots \ C_{i_L}]^T$, where the $C_{i_1}, C_{i_2}, \dots, C_{i_L}$, are the non zero values of the vector $\mathbf{C}$ and $\hat{\mathbf{Q}} = [q_{i_1} | \dots | q_{i_L}]$ be the submatrix of $\mathbf{Q}$ constructed by choosing columns $\{i_1, i_2, \dots, i_L\}$. It can be easily shown that (since $C_k = 0, k \notin I_u$),

$$\mathbf{c} = \mathbf{Q}\mathbf{C} = \hat{\mathbf{Q}}\hat{\mathbf{C}} \quad (4.18)$$

This can be further simplified (making use the complex symmetric property of vector $\mathbf{C}$ and the structured nature of the IFFT matrix) to produce,

$$\begin{aligned} \hat{\mathbf{Q}}\hat{\mathbf{C}} &= \overline{\mathbf{Q}} \overline{\mathbf{C}} \\ \overline{\mathbf{Q}} &= [\bar{q}_1 \ \bar{q}_2 \ \dots \ \bar{q}_N]^T \\ \bar{q}_k &= [\cos(\frac{2\pi j_1 k}{N}) + \dots + \cos(\frac{2\pi j_{n_u/2} k}{N}) - \sin(\frac{2\pi j_1 k}{N}) - \dots - \sin(\frac{2\pi j_{n_u/2} k}{N})]^T \\ \overline{\mathbf{C}} &= [C_{rj_1} \ C_{rj_2} \ \dots \ C_{rj_{n_u/2}}, \ C_{ij_1} \ C_{ij_2} \ \dots \ C_{ij_{n_u/2}}]^T; \ C_{rk} = \text{real}(C_k), C_{ik} = \text{imag}(C_k) \end{aligned} \quad (4.19)$$

Now the problem in (4.17) is reduced to,

$$\mathbf{c}^* = \arg \max_{\overline{\mathbf{C}}} (\min(\mathbf{x} + \overline{\mathbf{Q}} \overline{\mathbf{C}})) \quad (4.20)$$

The significance of (4.20) is that in here all the variables are real and hence this can be expressed as a classic optimization problem and can be solved by Linear Programming techniques. The Matlab optimization toolbox is used for this purpose. A Matlab function which takes in $\mathbf{X}$ (the frequency domain data vector) as the input, and gives out $\mathbf{C}$ (the power reduction tones) as the output using the algorithms provided by the optimization toolbox is developed. This function is incorporated into the simulink model using the S-Function block as shown in Figure 4.5. At the receiver a selector block is used to separate $\mathbf{X}$ and $\mathbf{C}$ after FFT.

4.3 GLCP-OFDM

In this section the details of the proposed adaptive weighting technique is presented. A block diagram of the GLCP-AW-OFDM scheme is given in Figure 4.6. The $N_0 \times 1$ QAM modulated symbol vector to be transmitted in one OFDM symbol is partitioned into P sub-vectors of size $K_0 \times 1$, where $N_0 = PK_0$. Thus, the i^{th} symbol sub-vector can be given by,

$$\mathbf{x}_i = [x_{iK_0} \ x_{iK_0+1} \ \dots \ x_{(i+1)K_0-1}] \quad (4.21)$$

where, $0 \leq i \leq P - 1$. Each symbol sub-vector is then precoded using the precoding matrix Φ of size $K \times K_0$, where $K \geq K_0$. The resulting precoded symbol sub-vectors $\mathbf{s}_i$ of size $K \times 1$ are given by,

$$\mathbf{s}_i = \Phi \mathbf{x}_i, \ 0 \leq i \leq P - 1 \quad (4.22)$$

With precoding the information in one symbol is spread over all the subcarriers. When $K = K_0$ there is no redundancy introduced in precoding, but when $K > K_0$ redundancy is introduced and the resulting precoded symbol sub-vector dimension is greater than the dimension of the original symbol sub-vector.

Figure 4.6: Block schematic of the proposed GLCP-AW-OFDM system

4.3.1 Subcarrier Mapping

In order to achieve maximum frequency diversity the precoded symbol sub-vector $\mathbf{s}_i$ in (4.22) should be modulated using subcarriers which undergo uncorrelated fading. An obvious way to achieve this is to use subcarriers which are further apart to modulate each adjacent precoded symbol in $\mathbf{s}_i$. Since there are P precoded symbol sub-vectors $\mathbf{s}_i$'s of length K , the optimal assignment would be to use the subcarriers which are separated by a distance of K from the pool of available subcarriers as shown in Figure 4.7.

If the channel matrix for the N subcarrier OFDM system is,

$$\mathbf{H} = \text{diag}[H_0, H_1, \dots, H_{N-1}] \quad (4.23)$$

then the channel matrix corresponding to the precoded sub-vector set $\mathbf{s}_i$ would be,

$$\mathbf{H}_i = \text{diag}[H_i, H_{i+K}, \dots, H_{i+(P-1)K}] \quad (4.24)$$

Figure 4.7: Subcarrier Mapping Scheme to maximize frequency diversity

Therefore the received post-FFT signal $\mathbf{y}_i$ corresponding to the sub-vector $\mathbf{s}_i$ becomes,

$$\begin{aligned} \mathbf{y}_i &= \mathbf{H}_i \mathbf{s}_i + \mathbf{n}_i \\ &= \mathbf{H}_i \Phi_i \mathbf{x}_i + \mathbf{n}_i \end{aligned} \quad (4.25)$$

where, $\mathbf{n}_i$ is a complex Gaussian noise vector of size $K \times 1$. In the following sections only the demodulation of the precoded symbol sub-vector $\mathbf{s}_i$ or $(\mathbf{x}_i)$ would be considered and hence the following simplification in notation is made with respect to the channel matrix coefficients and $\mathbf{x}_i$.

$$\begin{aligned} \mathbf{y}_i &= [y_{i,0} \ y_{i,1} \ \dots \ y_{i,K-1}]^T, & K \times 1 \\ \mathbf{H}_i &= [H_{i,0} \ H_{i,1} \ \dots \ H_{i,K-1}]^T, & K \times 1 \\ \mathbf{s}_i &= [s_{i,0} \ s_{i,1} \ \dots \ s_{i,K-1}]^T, & K \times 1 \\ \mathbf{n}_i &= [n_{i,0} \ n_{i,1} \ \dots \ n_{i,K-1}]^T, & K \times 1 \\ \mathbf{x}_i &= [x_{i,0} \ x_{i,1} \ \dots \ x_{i,K_0-1}]^T, & K_0 \times 1 \end{aligned} \quad (4.26)$$

$$\Phi = \begin{pmatrix} \phi_{0,0} & \phi_{0,1} & \dots & \phi_{0,K_0-1} \\ \phi_{1,0} & \phi_{1,1} & \dots & \phi_{1,K_0-1} \\ \dots & \dots & \dots & \dots \\ \phi_{K-1,0} & \phi_{K-1,1} & \dots & \phi_{K-1,K_0-1} \end{pmatrix}, \quad K \times K_0$$

4.3.2 Symbol Detection with Adaptive Subcarrier Weighting

For each symbol sub-vector $\mathbf{x}_i$ a total of K post-FFT signal values can be obtained using (4.25). The SNR of the post-FFT signal $y_{i,r}$ can be calculated by,

$$\begin{aligned} SNR_{i,r} &= \frac{|H_{i,r}|^2 E[|\phi_{i,r} \mathbf{x}_i|^2]}{E[|n_{i,r}|^2]}, \quad 0 \leq r \leq K-1 \\ &= |H_{i,r}|^2 |\phi_{i,r}|^2 \frac{\sigma_x^2}{\sigma_n^2} \end{aligned} \quad (4.27)$$

where, σ_x^2 and σ_n^2 are the average symbol and AWGN noise powers respectively. In the receiver, with maximum likelihood sequence detection, the estimated symbol sub-vector $\mathbf{x}_{i,\text{estimated}}$ corresponding to the transmitted symbol sub-vector $\mathbf{x}_i$ is given by,

$$\mathbf{x}_{i,\text{estimated}} = \arg \min_{\hat{\mathbf{x}}_i \in C} \|\mathbf{y}_i - \mathbf{H}_i \Phi_i \hat{\mathbf{x}}_i\|^2 \quad (4.28)$$

where, C is the space defined by having an entry each of the K_0 dimensional vector $\hat{\mathbf{x}}_i$, taken from an M-ary constellation. In terms of individual elements equation (4.28) can be expanded to give,

$$\begin{aligned} \mathbf{x}_{i,\text{estimated}} &= \arg \min_{\hat{\mathbf{x}}_i \in C} \{|y_{i,0} - H_{i,0} \hat{s}_{i,0}|^2 + |y_{i,1} - H_{i,1} \hat{s}_{i,1}|^2 + \dots + |y_{i,K-1} - H_{i,K-1} \hat{s}_{i,K-1}|^2\} \\ \hat{\mathbf{s}}_i &= \Phi_i \hat{\mathbf{x}}_i = [\hat{s}_{i,0} \ \hat{s}_{i,1} \ \dots \ \hat{s}_{i,K-1}]^T \end{aligned} \quad (4.29)$$

According to maximum likelihood sequence detection the deviation term ($d_r = |y_{i,r} - H_{i,r} \hat{s}_{i,r}|^2$) of the considered signal ($\hat{s}_{i,r}$) from the received signal ($y_{i,r}$) at each subcarrier is calculated and summed together and the sequence ($\hat{\mathbf{x}}_i$), which minimizes the total deviation ($\sum_{r=0}^{K-1} d_r$), is selected as the transmitted sequence. However, different subcarriers have different SNRs and hence the deviation term from a bad subcarrier could contribute adversely on the decision. We can observe that better performance can be achieved if more weight is given to the deviation of the considered symbol from the received symbol in a subcarrier with high SNR than in a subcarrier with low SNR -i.e the deviation terms (d_r) should be compensated according to the 'trustworthiness' of the subcarrier. This can be achieved by including weights that are proportional to the SNR of each subcarrier in the

following manner.

$$\mathbf{x}_{i,\text{estimated}} = \arg \min_{\hat{\mathbf{x}}_i \in C} \{w_0 \cdot |y_{i,0} - H_{i,0} \hat{s}_{i,0}|^2 + w_1 \cdot |y_{i,1} - H_{i,1} \hat{s}_{i,1}|^2 + \dots + w_{K-1} \cdot |y_{i,K-1} - H_{i,K-1} \hat{s}_{i,K-1}|^2\}$$

$$w_r \propto SNR_r \quad (4.30)$$

Notice this approach resembles the maximal ratio combining method popular in rake receivers. The weights for each subcarrier in (4.30) are selected adaptively for each OFDM symbol.

4.3.3 Symbol Detection with sequential mean square error reduction

The MLSE has exponential complexity; for example, for an M-ary modulation, the number of sequences that needs to be considered is M^K for the system described above and is often practically not possible. The authors in [33] have proposed a sequential mean square error reduction algorithm to mitigate the nonlinear clipping distortion of OFDM signals in time domain. In this section, we employ this algorithm in frequency domain to improve the receiver performance of the precoded symbols. Referring to equations (4.25) and (4.26),

Step 1: First an initial estimate of the received symbols is obtained using zero forcing detection.

$$\mathbf{x}_{i,\text{zf estimate}} = (\mathbf{H}_i \Phi_i)^{-1} \mathbf{y}_i \quad (4.31)$$

Based on this estimate hard decisions are made on the received symbols.

$$\mathbf{x}_{i,\text{zf hard decision}} = [\hat{x}_{i,0} \ \hat{x}_{i,1} \ \hat{x}_{i,2} \ \dots \ \hat{x}_{i,K_0-1}]^T \quad (4.32)$$

where $\hat{x} \in C$, C is the possible constellation space.

Step 2: The first element ($\hat{x}_{i,0}$) of $\mathbf{x}_{i,\text{zf hard decision}}$ is considered, and M new vectors are created by inserting all possible symbols from C as the first element, while keeping the other

elements in $\mathbf{x}_{i,\text{zf hard decision}}$ constant.

$$\mathbf{v}_{i,m} = [S_{i,m} \hat{x}_{i,1} \hat{x}_{i,2} \dots \hat{x}_{i,K_0-1}]^T, \quad 1 \leq m \leq M \quad (4.33)$$

Step 3: For $1 \leq m \leq M$ compute the distances d_m ; choose the vector which results in the minimum distance as the refined new symbol vector.

$$\begin{aligned} d_m &= \|\mathbf{y}_i - \mathbf{H}_i \Phi_i \mathbf{v}_{i,m}\|^2 \\ \mathbf{x}_{i,\text{zf hard decision}} &= \arg \min_{\mathbf{v}_{i,m}} (d_m) \end{aligned} \quad (4.34)$$

Step 4: Repeat Steps 2 and 3 for all K_0 elements in the vector; this progressively refines the symbol decisions.

When the algorithm is run once we need to search a total of $M \times K_0$ possibilities. The whole algorithm can be repeated many iterations to achieve a better performance; in this case the complexity would be number of iterations $\times M \times K_0$. A theoretical analysis of iterative decoding algorithms are always difficult and in Section 5.4.3 we present the simulation results for the performance of GLCP-OFDM with iterative decoding. However, it is evident that the performance of the above algorithm would depend on the initial zero forcing solution. In our analysis, we compare the performance when the initial solution is zero forcing and zero forcing with adaptive subcarrier selection (ACS) as described in [15].

Chapter 5

Results

In this chapter, a Monte Carlo analysis is utilized to evaluate the performance of the proposed OFDM impairment mitigation techniques. This chapter is organized as follows. Section 5.1 presents the results for the nonlinear distortion mitigation techniques. In Section 5.3 the results for the Average Power Reduction techniques are presented and the performance is analyzed. Performance of the GLCP-OFDM system with adaptive weighting and sequential mean square error reduction algorithm is given in Section 5.4.

5.1 Adaptive Modulation for Nonlinear Distortion Mitigation

5.1.1 System Parameters

For the simulation of the proposed adaptive modulation technique to mitigate the nonlinear distortion the following parameters are chosen. An OFDM system with $N = 64$ subcarriers with 16 QAM as primary (higher rate) modulation and 4 QAM as secondary (lower rate) modulation (when the nonlinear distortion is above the threshold) is chosen. The adaptive algorithm is run with the number of maximum number of iterations permitted taken to vary from 1, 2 and unlimited. The performance is compared with the Symbol Interleaving technique presented in [8] with the number of interleavers taken to be 20 and 40. A memoryless nonlinearity characterized by a Bessel function of the first kind as given in [11] is used to model the nonlinear system behaviour with the value π/V_π taken to be 1. To perform adap-

tive modulation it is assumed that information about the nonlinearity is available (estimated with sufficient accuracy by the nonlinear system identification process) at the transmitter.

5.1.2 Results for Nonlinear Distortion Mitigation using Adaptive Modulation

Figure 5.1 compares the BER performance of the adaptive modulation technique with the symbol interleaving technique. As seen from the results there is negligible performance

Figure 5.1: BER vs. E_b/N_o performance of OFDM without nonlinear distortion mitigation, with Symbol Interleaving, and Adaptive Modulation techniques.

improvement when the number of interleavers is increased. When the adaptive algorithm is constrained with 1 maximum number of iterations the performance is better in low SNR and is worse in the high SNR region than the symbol interleaving case. The performance progressively improves when the maximum number of iterations is increased as expected. In a practical transmission scenario, a cap on the maximum number of iterations can be

placed depending upon the transmission conditions, the computing power available at the transmitter side and the required quality of service.

The analysis is not complete without a comparison between the data rate reduction of the proposed technique and its complexity. In the following analysis, adaptive modulation technique with no limits on the maximum number of iterations (unconstrained) is compared with the symbol interleaving technique with 20 interleavers. The transmission efficiency is found by,

$$\text{transmission efficiency} = \frac{\text{number of bits transmitted}}{\text{number of bits that could be transmitted}} \quad (5.1)$$

Therefore, in our simulations where 16-QAM and 4-QAM are the modulation levels used, the transmission efficiency for one OFDM symbol (after the adaptive modulation process has converged) can be calculated by,

$$\begin{aligned} \text{transmission efficiency} &= \frac{N_4 \times 2 + N_{16} \times 4}{N \times 4} \\ N &= N_4 + N_{16} \end{aligned} \quad (5.2)$$

where, N_4 and N_{16} are the number of subcarriers having 4-QAM modulation and 16-QAM modulation respectively. The average transmission efficiency is then found by averaging the efficiency over many OFDM symbols. The average transmission efficiency of this technique for the given simulation parameters is found to be around 96.77%, whereas, the transmission efficiency of the symbol interleaving technique is 100% as it involves no reduction in the data rate. The adaptive modulation algorithm takes on average 2.4305 iterations to converge per OFDM symbol. Each iteration requires 2 FFT operations (1 IFFT and 1 FFT) and also it consists of the demodulation and distortion estimation parts. However, the complexity could be further reduced if one considers the symmetrical nature of the IFFT matrix and the fact that only the modulation of a fraction of the subcarriers is reduced in each iteration. The complexity of the symbol interleaving solution (in the considered case of 20 interleavers) is 20 IFFT operations per OFDM symbol and in this case the complexity cannot be further reduced by considering the symmetrical nature of the IFFT operation as each interleaver permutes the data symbols in a pseudo-random order. However, this technique does not estimate the nonlinear distortion and hence it has a lesser complexity on that aspect.

Figure 5.2, Figure 5.3 and Figure 5.4 illustrate the received symbol scatter diagrams.

Figure 5.2: Received 16 QAM constellation without nonlinear distortion mitigation.

Figure 5.3: Received 16 QAM constellation with Symbol Interleaving.

Figure 5.4: Received 16 QAM constellation with Adaptive Modulation.

The scatter diagrams are obtained without considering the effect of AWGN.

5.2 Using Simplex Signals

5.2.1 System Parameters

For the simulation of nonlinear distortion mitigation using simplex signals the following parameters are chosen. A 4-QAM modulation is used with 512 QAM symbols transmitted per one OFDM symbol. This translates into $N_1 = 512$ subcarriers when normal QAM modulation is used, and $N_2 = 768$ subcarriers when simplex signal modulation is used. The performance is compared with the Symbol Interleaving technique presented in [8] with the number of interleavers taken to be 20. The nonlinear function considered is same as in Section 5.1.1.

5.2.2 Results for Nonlinear Distortion Mitigation using Simplex Signals

Figure 5.7 compares the performance of Simplex Signal modulation with the Symbol Interleaving technique. As seen from the figure the performance improvement achieved by the Symbol Interleaving technique over the normal 4 QAM case is not significant. This is expected since the number of subcarriers N_1 is quite large and hence, we cannot realistically achieve significant improvements by a random permutation of input symbols. However, the Simplex Signal technique gives good performance, but here the downside is the bandwidth expansion required to transmit a 3-D symbol. It expands the bandwidth by a factor of 1.5 ($\frac{N_2}{N_1}$). It can be noted from Figure 5.7 (simplex signal and selected simplex curves) that the performance improvement achieved by permuting the simplex signal axis is also negligible.

Figure 5.8, Figure 5.9, Figure 5.10 and Figure 5.11 give the respective scatter diagrams. The scatter diagrams are obtained without considering the effect of AWGN.

Figure 5.5: PSD before nonlinearity.

Figure 5.6: PSD after nonlinearity.

Figure 5.7: BER vs. E_b/N_o performance of 4 QAM OFDM without nonlinear distortion mitigation, with Symbol Interleaving, with Simplex Signals, and with Selected Modified Simplex Signals.

5.3 Average Power Reduction Technique

5.3.1 System Parameters

In this simulation a $10 \times 10 \times 10$ m empty square room is considered with the transmitter and receiver placed at the corners with 45° separation between them as shown in Figure 4.4. The ceiling reflector is at a height of 1.5 m above the transmitter and receiver level. The optical source is assumed to have a Lambertian radiation pattern with mode number $n = 1$. Only the first four significant multipath components are considered for the simulations. The simulation is done for a data rate of 100 Mbits/s with the number of subcarriers (N) is taken to vary from 8, 16, 32, 64, 128, 256. The modulation scheme used is 16 QAM.

Figure 5.8: Received 4 QAM constellation without nonlinear noise mitigation.

Figure 5.9: Received 4 QAM constellation with Symbol Interleaving.

Figure 5.10: Received 3d constellation with Simplex Signals.

Figure 5.11: Received 3d constellation with Selected Modified Simplex Signals.

5.3.2 Results for Average Power Reduction

Figure 5.12 illustrates the average power reductions achieved using the described power reduction technique. When two subcarriers are used as peak reduction tones, good power reductions as high as 3 dB can be achieved even with higher values of N . However, when $N = 256$, the achievable power reductions are about 1.5 dB. By using more subcarriers for peak reduction this can be further improved (mathematically, when the number of reserved subcarriers are increased, we have more degree of freedom to find the optimal solution to the Linear Program). In this work the subcarriers (the 1st and 2nd subcarriers closer to dc), which are affected by the fluorescent noise are used for this purpose. This serves two things: (a) achieving average power reduction (b) improving the BER performance. The obvious tradeoff would be the reduced effective data rate. However, when the number of subcarriers is large ($N = 64, 128, 256$) the reduction in effective data rate is not significant (1 – 3%).

In the method presented by Khan et.al [27], with 8 subcarriers an average power reduction of about 3.6 dB is achieved using the block coding and/or variable bias schemes. With block coding scheme there will be bandwidth expansion for a given bit rate, but the bandwidth expansion ratio decreases as the number of subcarriers increases. With the variable bias schemes the fixed dc bias is replaced by a time varying bias signal and does not entail a bandwidth expansion. However, the implementation would be more complex. When the number of subcarriers is small, the computation required to implement the proposed

Figure 5.12: Average Power reductions achieved for different number of subcarriers.

techniques can be performed off-line and implemented via table lookup, and the simulations are done for carriers upto 8. However, in most practical systems a large number of subcarriers are used-for IEEE 802.11a/g standards (using radio frequencies) 64 point IFFT needs to be performed. In this work we are primarily concerned with using infrared as the physical layer alternative (as specified in IEEE 802.11 standard) for radio frequencies and naturally it is required to use a large number of subcarriers (32,64,128). With large number of subcarriers the table lookup becomes impractical and the efficient implementation of the above method proposed by Kahn is difficult.

However, by using the Tone Reservation technique, we can achieve power reductions similar to that obtained by Kahn's method. This method can be implemented for large number of subcarriers without considerable computational cost. The structured nature of the IFFT matrix is a significant factor that reduces the computational cost of calculating the solution to the linear program. In the simulations performed the matlab optimization function is able to converge within 1 – 3 iterations to find the optimal solution for smaller

values of N and within 5–6 iterations for larger values of N . Theoretically it had been shown that, for a linear program with L unknowns and $2N$ inequalities the complexity is shown to be $O(LN^2)$ [9]. But with a very structured Q (the IFFT matrix) the linear programming problem can be solved with complexity $O(N \log(N))$ [9].

In real implementation, finding the optimal solution could still be inefficient and cannot be practical for larger values of N ($N = 128, 256$) (we should note that the optimization has to be done in real time for each transmitted OFDM symbol). However, the complexity of the optimization problem can be further reduced (to $O(N)$) by using a gradient algorithm [9]. This sub optimal algorithm can give results close to the optimal solution after a few steps. It is to be noted that in our case a fixed bias will be used regardless of the OFDM transmit signal after IFFT. Therefore as long as the (negative value of) signal does not exceed this fixed bias (the composite signal after dc biasing remains non-negative) it is acceptable and hence we do not require the optimal solution each time. Here, we must note that in traditional radio wireless OFDM systems the goal is to minimize the peak to average power ratio and hence the smaller the ratio the better. But in optical wireless OFDM systems (with fixed bias) the goal is to minimize the average transmitted optical power, which is directly related to the bias used. (In optical wireless systems also we prefer the peak to average ratio to be as little as possible but this is only a secondary consideration.)

The BER curves in Figure 5.13, Figure 5.14 are plotted for SNR varying from -10 dB to 25 dB. (SNR of -10 dB is quite possible in diffuse optical wireless systems due to ambient light noise). As seen from Figure 5.13 when no carrier suppression is used, there is no significant improvement in the BER when the SNR goes beyond 10 dB—we can see an irreducible BER floor beyond 10 dB. This is the region where the fluorescent noise interference becomes the dominant noise source. When the subcarriers near dc are suppressed significant improvement in the BER curve can be observed when the SNR goes beyond 10 dB. When the first two subcarriers are suppressed the BER curves approach the desired BER values. It should also be noted that when the SNR is below 10 dB, there is not much improvement even with fluorescent noise suppression - this is the region where the white noise becomes the dominant noise source.

Figure 5.14 illustrates the performance when multipath dispersion is considered. The BER for a SNR of 25 dB (when 2 carriers are suppressed) is around 10^{-4} , which is below the desired BER for practical systems. In our simulations we considered a $10 \times 10 \times 10$ m square room for channel modeling and the simulations are done for a bit rate of 100 Mbits/s.

Figure 5.13: Performance under the effect of fluorescent noise and AWGN without considering multipath dispersion.

Figure 5.14: Performance under the effect of multipath dispersion, fluorescent noise and AWGN.

This could be the probable reason for high multipath propagation effects. We cannot make a sacrifice on these parameters to improve the BER performance. It should be noted that one of the real applications of diffuse optical wireless systems is WLANs (IEEE 802.11 standard) and to cater for real time multimedia speeds we need to have a high data rate over such range. Hence there is a need to use equalization schemes in order to mitigate the effects of multipath dispersion. With OFDM a simple frequency domain scheme can be applied for this purpose. Since each subcarrier can be considered as narrowband the frequency response of the channel can be considered flat for each subcarrier, but each subcarrier will experience different gains. The indoor optical wireless channels can be considered as slowly varying [26] and hence by sending a training sequence the channel response (gain) of each sub-band can be estimated from the received symbols and this value can be used to equalize the gains for the data symbols ($H_p = Y_p/X_p$). As mentioned above, in any scheme that uses the reserved

Figure 5.15: Analysis of bandwidth efficiency.

subcarrier technique the price that we have to pay is the data rate; the effective data rate goes down because some of the subcarriers do not carry useful information (note, however, the

total data rate is kept constant and hence the simulation results in Figure 5.13, Figure 5.14 are valid). Hence, the analysis is not complete without considering the tradeoff in terms of the effective data rate. It can be seen from Figure 5.15 that when the number of subcarriers are large ($N = 64, 128, 256$) the reduction in bandwidth efficiency is not significant and hence in practical systems (which usually use large number of subcarriers), this tradeoff is acceptable.

5.4 GLCP-OFDM

5.4.1 System Parameters

To simulate the performance of the GLCP-OFDM system the following typical parameters are chosen. A subcarrier subgroup size of $K = 8, 16$ is chosen. The modulation scheme used is BPSK. An 8×8 , 16×16 Hadamard matrix is used for precoding depending on the subcarrier subgroup size. The wireless channel is assumed to have a Rayleigh distribution with an exponential power delay profile. Phase shift on each path is assumed to be uniformly distributed over $[0, 2\pi]$. A SNR range of 0-25 dB at steps of 1 dB is considered with $SNR = \frac{\sigma_h^2 \sigma_x^2}{\sigma_w^2}$. The per-bit SNR is given by $E_b/N_0 = SNR/\log_2 M$, where M is the constellation size [15]. In the case of redundant precoding ($K_0 < K$), an 8×6 , 16×14 Hadamard matrix is used.

5.4.2 Simulation results for GLCP-OFDM with Adaptive Subcarrier Weighting

As can be seen from Figure 5.16 the GLCP-ACS-OFDM provides significant improvement over the GLCP-OFDM (Precoding only) system in low to medium SNR region. However, the performance difference begins to fade out when the SNR is high as the effect of neglecting information from a particular subcarrier begins to affect the BER performance. This is not the case with GLCP-AW-OFDM where the subcarriers are weighted according to their respective SNR and hence no useful information is lost even in the high SNR region and hence we see a superior error performance in low as well as high SNR regions.

Figure 5.16: BER vs. E_b/N_o performance of GLCP-OFDM, GLCP-ACS-OFDM and GLCP-AW-OFDM with $K = K_0 = 8$.

5.4.3 Simulation results for GLCP-OFDM with sequential mean square error reduction

MLSE is often too complex to be implemented in practical scenarios and hence GLCP-OFDM with sequential mean square error reduction could be an attractive alternative, with sub optimum decoding. We have compared the performance of precoding with zero forcing detection, precoding with ACS detection and sequential mean square error detection methods. The sequential mean square error reduction is performed with different initial hard decided solutions (either a solution by simple zero forcing or a ACS and zero forcing). The whole algorithm is also run more than one iteration to compare the performance improvement (for example: the curve corresponding to 'iterative decoding-zf-1' stands for the error performance when the algorithm is started with an initial zero forcing solution and run for one iteration).

Figure 5.17 illustrates the performance of the system with a subcarrier subgroup size of 8. Looking at the figure it is evident that Precoding with ACS performs better in low

SNR region and worse than the Precoding only case in high SNR region as expected. It is also evident, that Iterative decoding improves the performance but the performance is dependent upon the initial solution fed to the algorithm. We can also note that running the whole sequential mean square error reduction algorithm twice results in a slightly improved performance. Running the algorithm higher number of times (not shown in the results) did not result in any significant improvements. Figure 5.18 shows the performance when redundant precoding is used (number of subcarriers is greater than the number of symbols), and we see that the performance is improved compared to the previous figure as expected, with a data rate reduction of $\frac{K_0}{K}$. It can be noted that in this case, Precoding with ACS is better than the Precoding only case in the high SNR region too, since redundancy is introduced; hence the error introduced in neglecting information on one subcarrier does not affect performance considerably. The results for the iterative decoding algorithm is similar to the previous figure with the performance closely dependent upon the quality of the initial hard decision fed to the algorithm.

Figure 5.19 and Figure 5.20 also analyze the same detection schemes when the subcarrier subgroup size is 16, with and without redundant precoding. In Figure 5.19 we can note that the performance of Precoding with ACS is better than the Precoding only case over all SNR regions (even without any redundancy); this is because the information on only one worst subcarrier among 16 subcarriers is neglected, and hence the performance degradation due to that is negligible. Also looking at the curves we can conclude that running the iterative decoding algorithm more than once does not result in a performance improvement that would justify the increased complexity.

Figure 5.17: BER vs. E_b/N_o performance of GLCP-OFDM with sequential mean square error reduction with $K = K_0 = 8$.

Figure 5.18: BER vs. E_b/N_o performance of GLCP-OFDM with sequential mean square error reduction with $K = 8, K_0 = 6$.

Figure 5.19: BER vs. E_b/N_o performance of GLCP-OFDM with sequential mean square error reduction with $K = K_0 = 16$.

Figure 5.20: BER vs. E_b/N_o performance of GLCP-OFDM with sequential mean square error reduction with $K = 16, K_0 = 14$.

Chapter 6

Conclusions and Future work

6.1 Conclusions

Beyond 3G networks we see an increased demand for bandwidth not only in the downlink but also in the uplink. There is also the need for speed in backbone networks and broadcast networks to accommodate 4G services, high quality multimedia and converged digital services. The inherent resistance of OFDM systems to ISI and its flexibility to accommodate space-time-frequency processing make it a viable solution for the challenges of the future.

Although the concept of OFDM had been around for some time, it is the introduction of FFT/IFFT technique to implement the multicarrier modulation instead of the filter bank technique, that made OFDM practically implementable. The nonlinear distortion of signal as a result of the high PAPR, the increased average power requirement in intensity modulated optical systems and the signal's vulnerability to frequency selective fading are some of the issues that still need to be alleviated for widespread adoption of OFDM systems in future communication systems. In this thesis we present novel techniques to improve the above mentioned problems.

In the adaptive modulation scheme to mitigate the nonlinear distortion, a calculation is done at the transmitter side which identifies the subcarriers with high distortion and correspondingly reduces the modulation level on those subcarriers. This procedure is repeated recursively until the nonlinear distortion is below a predetermined threshold. This method outperforms OFDM without distortion mitigation by 6 dB and the symbol interleaving technique in [8] by 2 dB for a target BER of 10^{-2} . The data rate reduction in this method is 3.5%.

In the modified tone reservation technique certain subcarriers are reserved for power reduction purposes and modulated in order to increase the minimum value of the OFDM (time domain) electrical signal after IFFT and hence reducing the fixed dc bias required and thereby effectively reducing the average optical power. When two subcarriers are used for peak reduction, reasonable power reductions (3 dB) are achieved and the results comparable to that of the block coding and variable bias techniques used by Kahn [27]. This technique can be implemented for large number of subcarriers and can be efficiently implemented in real systems. By selecting the subcarriers, which are affected by fluorescent noise for average power reduction, we have shown that significant BER improvement also can be achieved.

In the GLCP-AW method during MLSE, the distance of the possible constellation points from the received symbols are weighted according to the SNR of each subcarrier before the sequence with the minimum distance is chosen. This technique is shown to outperform GLCP-OFDM and GLCP-ACS-OFDM over all SNR ranges at the expense of slightly increased complexity. The performance improvement is realized due to the fact that no information is discarded in this approach as in the GLCP-ACS method. However, a more realistic solution would be to use the sequential mean square error reduction algorithm which reduces the exponential search to linear search. From our analysis we conclude that precoding with ACS combined with the sequential mean square error reduction algorithm would be a good solution to the lack of frequency diversity issue of OFDM systems, if the receiver could handle the linearly complex iterative decoding algorithm. A suitable application of this technique would be in UMTS LTE systems where the uplink transmission is based on 'DFT-spread OFDM'.

6.2 Future Research

- In the adaptive modulation technique presented above, the distortion after nonlinearity is estimated only by running a simulation. It would be a useful contribution to estimate the distortion analytically. For example, for a polynomial nonlinearity the distortion of a given signal can be estimated in frequency domain using a convolution according to the polynomial coefficients.
- As mentioned in [25], a more effective criteria to evaluate the performance of various nonlinear distortion mitigation schemes is to measure the reduction in the actual

nonlinear distortion as opposed to the peak to average power ratio reduction. This is especially useful if the nonlinearity is relatively stable and can be estimated with reasonable accuracy as in the case of laser or optical fiber nonlinearity. An analysis of various existing techniques using this criteria would be useful especially for future very high bit rate optical OFDM systems.

- In the optical wireless case the main source of interference is from fluorescent noise and since we can approximately model the noise, we have priori knowledge of which subcarriers are 'bad'. However, in other OFDM systems with intensity modulation, the channel frequency response may not be known apriori; therefore in this case an adaptive approach should be used in order to maximize performance. Also in our work we took two adjacent subcarriers as power reduction tones. Further work should be done to investigate if there is any advantage in selecting subcarriers which are far apart - i.e the effect of the placement of reserved tones on the performance should be analyzed.
- The use of simplex signals has been shown as an effective means of containing the signal amplitudes within a particular level. Further investigation needs to be done on the other characteristics of simplex signals. To transmit a simplex symbol more than one subcarrier is needed, thus this system inherently involves some form of diversity. This property can be further exploited when the adaptive weighting technique discussed in this work is applied in the demodulation of one simplex symbol which spans across multiple subcarriers.
- The results for the techniques to improve the performance of GLCP-OFDM systems are obtained assuming perfect channel knowledge. It would be useful to analyze their performance with imperfect channel estimation; looking in terms of the future '3.9G' (UMTS LTE) and 4G systems, there is a need to do more research on a joint strategy to cover both nonlinear distortion and frequency diversity using precoding.

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