OFDM Based WLAN Systems

Muhammad Imadur Rahman, Suvra Sekhar Das, Frank H.P. Fitzek

Center for TeleInFrastruktur (CTiF), Aalborg University
Neils Jernes Vej 12, 9220 Aalborg Øst, Denmark
phone: +45 9635 8688; e-mail: {*imr,ssd,ff*}*@kom.aau.dk*

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Abstract

This report is intended to provide an overview of the present state of WLANs with respect to physical layer issues and identify some key research issues for future generations of WLANs. The current standards (IEEE 802.11a in USA, HiperLAN/2 in Europe and MMAC in Japan) are all based on OFDM in their PHY layer. Thus, detailed attention on basics of OFDM related WLAN systems are given in the report. This is to introduce the existing techniques that are proposed or implemented in different OFDM based WLANs. A special attention is given on synchronization and channel estimation issues in WLAN due to the fact that virtually it is impossible to obtain a reasonable quality of service without perfect (or near perfect or efficient) synchronization and channel estimation. At the end of the report, some interesting topics that need to be studied in the development of future generation wireless systems based on OFDM are presented.

It is worth mentioning here that the explanation on various topics presented here are in no way a complete description. For details, it is suggested to study the references that are mentioned in this report.

Preface

Science is a wonderful thing if one does not have to earn ones living at it.
-Albert Einstein

This report is an overview of Orthogonal Frequency Division Multiplexing(OFDM). It is an outcome of the literature survey in the initial phase of our PhD studies under the Department of Communications Technology of Aalborg University. The main focus of our PhD program is OFDM for broadband mobile wireless communications.

The scope of the report is limited to a survey of the OFDM field, establishment of its analytical description and to list the challenges and areas of research. It does not intend to provide any solution to the problems listed. This document is targeted for use as a first hand guide to OFDM fundamentals, to explore the activities around OFDM till to date, and to survey the potential research areas involving OFDM implementation towards the next generation communication systems.

We owe great thanks to Prof. Ramjee Prasad and Assoc. Prof. Ole Olsen for their caring guidance and kind co-operation in helping us prepare this report. This work was partially supported by Samsung, Korea under the 'JADE' project framework in our research group. We highly appreciate their support in this regard.

As we progress with the PhD studies in Aalborg, we intend to improve upon this report with our increasing knowledge and competencies in OFDM based wireless systems.

Should this report have any factual error or typo mistakes, or should you have any suggestion on how to improve the contents or writing style, you are always welcome to contact the authors.

Muhammad Imadur Rahman & Suvra Sekhar Das
Aalborg University, Denmark
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Abbreviations

2G	2^{nd} Generation
3G	3^{rd} Generation
4G	4^{th} Generation
ADC	Analog-to-Digital Converter
ADSL	Asymmetric Digital Subscriber Lines
ASC	Antenna Selection Combining
BPSK	Binary Phase Shift Keying
CCK	Complementary Code Keying
CDD	Cyclic Delay Diversity
COFDM	Coded Orthogonal Frequency Division Multiplexing
CP	Cyclic Prefix
CSI	Channel State Information
DA	Data Aided Algorithms
DAB	Digital Audio Broadcasting
DAC	Digital-to-Analog Converter
DBLAST	Diagonal Bell Labs LAyered Space-Time
DFT	Discrete Fourier Tranform
DVB	Digital Video Broadcasting
DVB-T	Digital Video Broadcasting - Terrestrial
DSSS	Direct Sequence Spread Spectrum
EGC	Equal Gain Combining
FEC	Forward Error Correction
FFT	Fast Fourier Transform
FHSS	Frequency Hopping Spread Spectrum
FM	Frequency Modulation
HDSL	High-bit-rate Digital Subscriber Lines
HDTV	High Definition TeleVision
HiperLAN	High Performance Radio Local Area Networks
ICI	Inter Carrier Interference
IDFT	Inverse Discrete Fourier Transform
IFFT	Inverse Fast Fourier Transform
IR	Infra Red
ISI	Inter Symbol Interference
LOS	Line-Of-Sight
LS	Least Squared
LMMSE	Linear Minimum Mean Squared Error

MAN	Metropolitan Area Networks
MARC	Maximum Average (Signal-to-Noise) Ratio Combining
MIMO	Multiple Input Multiple Output
MMAC	Multimedia Mobile Access Communications Systems
MMSE	Minimum Mean Squared Error
MRC	Maximum Ratio Combining
MRRC	Maximum Ratio Receiver Combining
MSE	Mean Squared Error
MW	Millimeter Wave
NDA	Non Data-Aided Algorithms
NLOS	Non Line-Of-Sight
OFDM	Orthogonal Frequency Division Multiplexing
PAPR	Peak-to-Average Power Ratio
PBCC	Packet Binary Convolutional Coding
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
RMS	Root-Mean-Square
SC	Selection Combining
SM	Spatial Multiplexing
SNR	Signal-to-Noise Ratio
SSC	Subcarrier Selection Combining
STC	Space-Time Coding
VBLAST	Vertical Bell Labs LAyered Space-Time
VDSL	Very-high-speed Digital Subscriber Lines
VoWIP	Voice over Wireless Internet Protocol
WLAN	Wireless Local Area Networks
W-OFDM	Wideband Orthogonal Frequency Division Multiplexing

Chapter 1

Introduction

Wireless communication has gained a momentum in the last decade of 20^{th} century with the success of 2^{nd} *Generations* (2G) of digital cellular mobile services. Worldwide successes of GSM, IS-95, PDC, IS-54/137 etc. systems have shown new way of life for the new information and communication technology era. These systems were derived from a voice legacy, thus primary services were all voice transmission. 2G systems provided better quality of services at lower cost and a better connectivity compared to previous analog cellular systems. Numerous market researches show that there is a huge demand for high-speed mobile multimedia services all over the world. With the advent of 3^{rd} *Generation* (3G) wireless systems, it is expected that higher mobility with reasonable data rate (up to 2Mbps) can be provided to meet the current user needs. But, 3G is not the end of the tunnel, ever increasing user demands have drawn the industry to search for better solutions to support data rates of the range of tens of Mbps. Naturally dealing with ever unpredictable wireless channel at high data rate communications is not an easy task. Hostile wireless channel has always been proved as a bottleneck for high speed wireless systems. This motivated the researchers towards finding a better solution for combating all the odds of wireless channels; thus, the idea of multi-carrier transmission has surfaced recently to be used for future generations of wireless systems.

3G promises a wire line quality of services via a wireless channel. For wide area coverage, further expansions of 3G systems are already a question of research in all over the world. Certainly the bit rate will be much higher than 2Mbps for such a system, up to tens of Mbps. For local area coverage, Wireless Local Area Networks (WLANs), such as IEEE 802.11a, HiperLAN/2 or MMAC[1] standards are capable of providing data rates up to 54 Mbps. Along with these three, there are few other emerging short-range wireless applications available, such as Bluetooth, HomeRF, etc.

WLANs can potentially be a promising tool in different user environments, namely home, corporate and public environment etc. WLANs are used to connect wireless users to a fixed LAN in corporate environments. A major WLAN application will be in public sectors, where WLAN can be used to connect a user to the backbone network. Airports, hotels, high-rising offices, city centers will be target area for such public WLAN usage. It is becoming more and more evident that WLANs will play a greater role in future. A popular vision of future generations of telecommunications systems suggests that it will be an amalgamation of high data-rate wireless wide area networks (such as UMTS) and newly standardized WLANs. However systems of the

[1] IEEE802.11a is an USA-standard, HiperLAN/2 is a European standard and MMAC is developed in Japan. All three of the standards are almost similar in their PHY layer.

near future will require WLANs with data rate of greater than 100 Mbps, and so there is a need to further improve the capacity of existing WLAN systems. Although the term 4G is not yet clear to the industry, it is likely that they will enhance the 3G networks in capacity, allowing greater range of applications and better universal access. Some of the visionaries term the system as *Mobile Broadband Services* (MBS). A seamless and uninterrupted service quality for a user regardless of the system he/she is using will be one of the main goals of future systems. The expected systems will require an extensive amount of bandwidth per user.

Several technologies are considered to be candidates for future applications. According to many, *Wireless Personal Area Networks* (WPAN) will be a major application of future wireless communications [1]. WPAN will enable a kind of ubiquitous communications, that is, WPAN will provide a continuous network connection to the user. This will revolutionize the future home, where wireless communications appliances will be an integral part of home life. A user can communicate through his networked WPAN that includes various communication-enabled devices and can seamlessly move or address another user or WPAN nearby, through WPAN-WPAN connections or through some backbone networks or WLAN. WPANs will be very similar to WLANs in terms of operation, application and implementation, thus OFDM will be vividly present in all future wireless devices as it appears now.

Orthogonal Frequency Division Multiplexing (OFDM) is a special form of multi-carrier transmission where all the subcarriers are orthogonal to each other. OFDM promises a higher user data rate and great resilience to severe signal fading effects of the wireless channel at a reasonable level of implementation complexity. It has been taken as the primary physical layer technology in high data rate Wireless LAN/MAN standards. IEEE 802.11a and HiperLAN/2 have the capability to operate in a range of a few tens of meters in typical office space environment. IEEE 802.16a uses *Wideband OFDM* (W-OFDM) a patented technology of Wi-LAN to server up to 1 km radius of high data rate fixed wireless connectivity. In the upcoming standard IEEE 802.20, which is targeted at achieving data rate of greater than 1 Mbps at 250 kmph, OFDM is one of the potential candidate. Thus we see that there is a strong possibility that next generation wireless era belongs to OFDM technology.

1.1 Scope of the Report

This report attempts to connect the different strings of OFDM and create a comprehensive reference. We attempt to provide an understanding of OFDM system along with a description of the challenges and research areas for enhancing system capacity and improving link quality.

1.2 Organization of the Report

The rest of this report is organized as follows.

In *Chapter 2*, we talk about the wireless channel impairments that are encountered in designing a communication system. It covers multipath interference, doppler effect, shadow fading, propagation path loss, time and frequency selective fading and finally it includes the benefit of using Multi-carrier modulation technique to overcome the channel impairments.

In *Chapter 3*, we first introduce the reader to the history and evolution of OFDM. It then covers the components of an OFDM transceiver one by one. The chapter ends with a discussion of the the advantages and disadvantages of OFDM systems. It is explained how OFDM deals with

ISI/ICI and how OFDM provides in spectral efficiency. Synchronization, PAPR and co-channel interference issues are also introduced in this chapter.

In *Chapter 4*, we build an analytical model of OFDM system based on the basics obtained in the previous chapter. Mainly we have explained the system flow in analytical expressions to have a better understanding of the scheme. At the end of the chapter, we have also included a simple matrix model of the system in co-operation with analytical model. The matrix model will give a better understanding, in case there is a need to prepare a simulation model in certain programming environment, such as MATLAB.

Chapter 5 deals with diversity issues. We introduced how MIMO techniques can be incorporated with OFDM systems to significantly improve the capacity and link quality of the system.

In *Chapter 6*, we discuss the synchronization issues in detail. It covers carrier frequency synchronization, symbol timing and sampling clock synchronization.

In *Chapter 7*, we deal with channel estimation techniques, and explain three methods of estimation, which are pilot based, training symbol based and blind channel estimation.

Chapter 8 briefly identifies a number of key research issues related to the development of OFDM wireless system aimed at high data rate and vehicular mobility.

Finally *Chapter 9* concludes the report.

Chapter 2

Channel Impairments and MCM

Wireless channel is always very unpredictable with harsh and challenging propagation situations. Wireless channel is very different from wire line channel in a lot of ways. Multipath reception is the unique characteristic of wireless channels. Together with multipath, there are other serious impairments present at the channel, namely propagation path loss, shadow fading, Doppler spread, time dispersion or delay spread, etc.

2.1 Multipath Scenario

Multipath is the result of reflection of wireless signals by objects in the environment between the transmitter and receiver. The objects can be anything present on the signal travelling path, i.e. buildings, trees, vehicles, hills or even human beings. Thus, multipath scenario includes random number of received signal from the same transmission source; depending on the location of transmitter and receiver, a direct transmission path referred to as the *Line-Of-Sight* (LOS) path may be present or may not be present. When LOS component is present (or when one of the components is much stronger than others), then the environment is modelled as Ricean channel, and when no LOS signal is present, the environment is described as Rayleigh channel.

Multipaths arrive at the receiver with random phase offsets, because each reflected wave follows a different path from transmitter to reach the receiver. The reflected waves interfere with direct LOS wave, which causes a severe degradation of network performance. The resultant is random signal fades as the reflections destructively (and/or constructively) superimpose one another, which effectively cancels part of signal energy for a brief period of time. The severity of fading will depend on delay spread of the reflected signal, as embodied by their relative phases and their relative power [2].

A common approach to represent the multipath channel is channel impulse response which gives us the delay spread of the channel. Delay spread is the time spread between the arrival of the first and last multipath signal seen by receiver. In a digital system, delay spread can lead to ISI. In Figure 2.1, delay spread amounts to τ_{max}. It is noted that delay spread is always measured with respect to the first arriving component.

Let's assume a system transmitting in the time intervals T_{sym}. The longest path with respect to the earliest path arrives at the receiver with a delay of τ_{max}; in other words, the last path arrives τ_{max} seconds after the first path arrives. This means that a received symbol can theoretically be influenced by previous symbols, which is termed as ISI. With high data rate, T_{sym} can be very small; thus the number of symbols that are affected by ISI can be in multiple of tens or more.

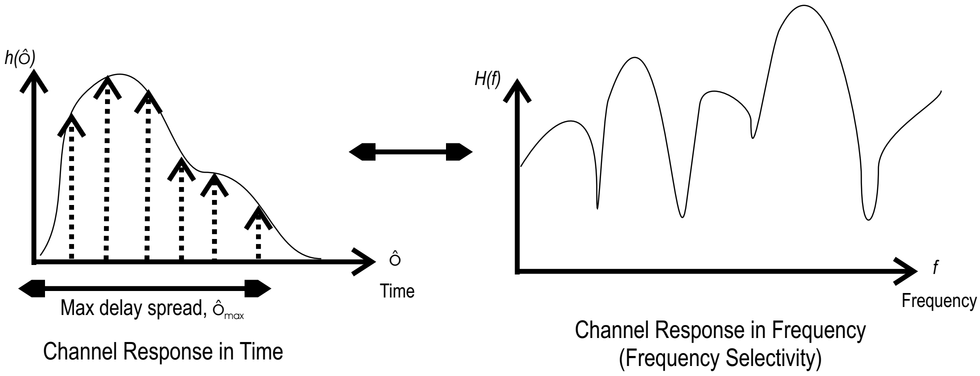

Figure 2.1: Channel Impulse Responses and Corresponding Frequency Response

Combating the influence of such large ISI at the receiver is very challenging and sometimes may become unattainable at very severe channel conditions [3].

2.2 Doppler Effect

Doppler spread is caused by the relative motion of transmitter and receiver. For example, in an urban environment in the city center, the vehicles are always moving; the walking pedestrians are also changing their locations continuously, thus their movements affect the transmission medium. A high Doppler can be experienced when a user is located in a fast moving car or in a speedy train, because the relative motion will be higher when either transmitter or receiver is moving very fast. This relative motion of transmitter and receiver changes the received signal from the original transmitted signal. When they are moving towards each other, the frequency of the received signal is higher than the source and when they moving away from each other, the received frequency decreases. When the relative speed is higher, then Doppler shift can be very high, and thus the receiver may become unable to detect the transmitted signal frequency. Even at lower relative motion when the Doppler shift is usually very little, if the transmission and reception technique is very sensitive to carrier frequency offset, then the system may fail.

2.3 Shadow Fading or Shadowing

Shadow fading is another troublesome effect of wireless channel. Wireless signals are obstructed by large obstacles, like huge buildings, high hills, etc. These large objects cause reflections off their surface and attenuation of signals passing through them, resulting in shadowing, or shadow fading. These shadows can result in large areas with high path loss, causing problems with communications. The amount of shadowing depends on the size of the object, the structure of the material, and the frequency of the RF signal. Large attenuations by huge obstacles can result in deep fading behind them. Under this condition, most of the received signal energy comes from reflected and diffracted paths of the original signal, because LOS is absent due to large object between the transmitter and the receiver.

2.4 Propagation Path Loss

Together with multipath effect, shadowing and Doppler spread, the propagation loss is also very significant at some specific situation. The propagation loss increases by fourth power of distance [4]. So, for higher distances, propagation loss becomes significant. Well defined situation specific path loss models are available to estimate the propagation path loss.

2.5 Time Dispersion and Frequency Selective Fading

Time dispersion represents distortion of the signal that is manifested by the spreading of the modulation symbols in time domain. We all know that channel is mostly band-limited in case of broadband multimedia communications, i.e. the coherence bandwidth of the channel is always smaller than modulation bandwidth. So ISI is unavoidable in wireless channels. In many instances, fading by the multipath will be frequency selective. This implies that signals will be affected only at part of the available frequency band. The effect has a random pattern for any given time. At certain frequencies, it will be enhanced (constructive interference) and will be completely (or partially) suppressed at other frequencies. Frequency selective fading occurs when channel introduces time dispersion and the delay spread is larger than the symbol period. Frequency selective fading is difficult to compensate because the fading characteristics is random and sometimes may not be easily predictable. When there is no dispersion and the delay spread is less than the symbol period, the fading will be flat, thereby affecting all frequencies in the signal equally. Practically flat fading is easily estimated and compensated with a simple equalization [2].

2.6 The Benefit of Using Multicarrier Transmission

A single carrier system suffers from trivial ISI problem when data rate is extremely high. We need high data rate to support wireless broadband applications, thus these applications always suffer from ISI. According to previous discussions, we have seen that with a bandwidth B and symbol duration T_{sym}, when $\tau_{max} > T_{sym}$, then ISI occurs. Multichannel transmission has been surfaced to solve this problem. The idea is to increase the symbol period of subchannels by reducing the data rate and thus reducing the effect of ISI. Reducing the effect of ISI yields an easier equalization, which in turn means simpler reception techniques.

Wireless multimedia solutions require up to tens of Mbps for a reasonable quality of service. If we consider single carrier high speed wireless data transmission, we see that the delay spread at such high data rate will definitely be greater than symbol duration even considering the best case outdoor scenario. Now, if we divide the high data rate channel over number of subcarriers, then we have larger symbol duration in the subcarriers and maximum delay spread is much smaller than the symbol duration.

Figure 2.2 describes this very efficiently [5]. Let's assume that we have available bandwidth B of 1MHz. Now in a single carrier approach, we transmit the data at symbol duration of 1μs. Consider a typical outdoor scenario where maximum delay spread can be 10μs, so at the worst case scenario, at least 10 symbols will be affected by each and every symbol. Thus, ISI effect of every symbol will be spread to 10 successive symbols.

In a single carrier system, this situation is compensated by using adaptive equalization technique. Adaptive equalization estimates the channel impulse response and multiplies complex

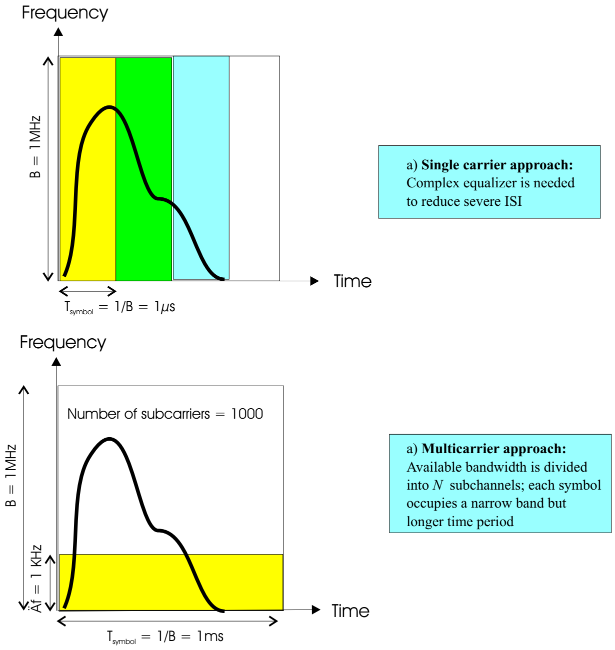

Figure 2.2: Single Carrier Vs Multicarrier Approach

Design Parameters for outdoor channel	Required data rate	1Mbps
	RMS delay spread, σ	10μs
	Channel coherence bandwidth, $B_c = \frac{1}{5\sigma}$	20KHz
	Frequency selectivity condition	$\sigma > \frac{T_{sym}}{10}$
Single carrier approach	Symbol duration, T_{sym}	1μs
	Frequency selectivity	$10\mu s > \frac{1\mu s}{10} \Longrightarrow$ **YES**
	ISI occurs as the channel is frequency selective	
Multicarrier approach	Total number of subcarriers	128
	Data rate per subcarrier	7.8125kbps
	Symbol duration per subcarrier	$T_{carr} = 128\mu s$
	Frequency selectivity	$10\mu s > \frac{128\mu s}{10} \Longrightarrow$ **NO**
	ISI is reduced as flat fading occurs. CP completely removes the remaining ISI	

Table 2.1: Comparison of Single Carrier and Multicarrier Approach in terms of Channel Frequency Selectivity

conjugate of the estimated impulse response with the received data signal at the receiver. However, there are some practical computational difficulties in performing these equalization techniques at tens of Mbps with compact and low cost hardware. It is worth mentioning here that compact and low cost hardware devices do not necessarily function at very high data speed. In fact, equalization procedures take bulk of receiver resources, costing high computation power and thus overall service and hardware cost becomes higher. Complex receivers are very efficient in performance, but not cost efficient.

One way to achieve reasonable quality and solve the problems described above for broadband mobile communication is to use parallel transmission. In a crude sense, someone can say in principle that parallel transmission is just the summation of a number of single carrier transmissions at the adjacent frequencies. The difference is that the channels have lower data transmission rate than the original single carrier system and the low rate streams are orthogonal to each other. If we consider a multi-carrier approach where we have N number of subcarriers, we can see that we can have $\frac{B}{N}$ Hz of bandwidth per subcarrier. If $N = 1000$ and $B = 1$MHz, then we have subcarrier bandwidth B_{carr} of 1kHz. Thus, symbol duration in each subcarrier will be increased to 1ms ($= \frac{1}{1kHz}$). Here each symbol occupies a narrow band but longer time period. This clearly shows that maximum delay spread of 1 msec will not have any ISI effects on received symbols in the outdoor scenario mentioned above. In another thought, multi-carrier approach turns the channel to a flat fading channel and thus can easily be estimated.

Theoretically increasing the number of subcarriers should be able to give better performance in a sense that we will able to handle larger delay spreads. But several typical implementation problems arise with large number of subcarriers. When we have large numbers of subcarriers, then we will have to assign the subcarriers frequencies very close to each other. We know that receiver needs to synchronize itself to every subcarrier frequency in order to recover data related to that particular subcarrier. When spacing is very little, then the receiver synchronization components need to be very accurate, which is still not possible with low-cost RF hardware. Thus, a reasonable trade-off between carrier spacing and number of subcarriers must be achieved.

Table 2.1 describes how multicarrier approach can convert the channel to flat fading channel from frequency selective fading channel. We have considered a multicarrier system with a single carrier system, where the system data rate requirement is 1Mbps. When we use 128 subcarriers for multicarrier system, we can see that the ISI problem is clearly solved. It is obvious that if we increase the number of subcarriers, the system will provide even better performance.

Chapter 3

OFDM Fundamentals

The nature of WLAN applications demands high data rates. Naturally dealing with ever-unpredictable wireless channel at high data rate communications is not an easy task. The idea of multi-carrier transmission has surfaced recently to be used for combating the hostility of wireless channel as high data rate communications. OFDM is a special form of multi-carrier transmission where all the subcarriers are orthogonal to each other. OFDM promises a higher user data rate transmission capability at a reasonable complexity and precision.

At high data rates, the channel distortion to the data is very significant, and it is somewhat impossible to recover the transmitted data with a simple receiver. A very complex receiver structure is needed which makes use of computationally extensive equalization and channel estimation algorithms to correctly estimate the channel, so that the estimations can be used with the received data to recover the originally transmitted data. OFDM can drastically simplify the equalization problem by turning the frequency selective channel to a flat channel. A simple one-tap equalizer is needed to estimate the channel and recover the data.

Future telecommunication systems must be spectrally efficient to support a number of high data rate users. OFDM uses the available spectrum very efficiently which is very useful for multimedia communications. Thus, OFDM stands a good chance to become the prime technology for 4G. Pure OFDM or hybrid OFDM will be most likely the choice for physical layer multiple access technique in the future generation of telecommunications systems.

3.1 History and Development of OFDM

Although OFDM has only recently been gaining interest from telecommunications industry, it had a long history of existence. It is reported that OFDM based systems were in existence during the Second World War. OFDM had been used by US military in several high frequency military systems such as KINEPLEX, ANDEFT and KATHRYN [6]. KATHRYN used AN/GSC-10 variable rate data modem built for high frequency radio. Up to 34 parallel low rate channels using PSK modulation were generated by a frequency multiplexed set of subchannels. Orthogonal frequency assignment was used with channel spacing of 82Hz to provide guard time between successive signaling elements [7].

In December 1966, Robert W. Chang[1] outlined a theoretical way to transmit simultaneous data stream trough linear band limited channel without *Inter Symbol Interference* (ISI) and

[1]Robert W. Chang, *Synthesis of Band-limited Orthogonal Signals for Multichannel Data Transmission*, The Bell Systems Technical Journal, December 1966.

Inter Carrier Interference (ICI). Subsequently, he obtained the first US patent on OFDM in 1970 [8]. Around the same time, Saltzberg[2] performed an analysis of the performance of the OFDM system. Until this time, we needed a large number of subcarrier oscillators to perform parallel modulations and demodulations.

A major breakthrough in the history of OFDM came in 1971 when Weinstein and Ebert[3] used *Discrete Fourier Transform* (DFT) to perform baseband modulation and demodulation focusing on efficient processing. This eliminated the need for bank of subcarrier oscillators, thus paving the way for easier, more useful and efficient implementation of the system.

All the proposals until this time used guard spaces in frequency domain and a raised cosine windowing in time domain to combat ISI and ICI. Another milestone for OFDM history was when Peled and Ruiz[4] introduced *Cyclic Prefix* (CP) or cyclic extension in 1980. This solved the problem of maintaining orthogonal characteristics of the transmitted signals at severe transmission conditions. The generic idea that they placed was to use cyclic extension of OFDM symbols instead of using empty guard spaces in frequency domain. This effectively turns the channel as performing cyclic convolution, which provides orthogonality over dispersive channels when CP is longer than the channel impulse response [6]. It is obvious that introducing CP causes loss of signal energy proportional to length of CP compared to symbol length, but, on the other hand, it facilitates a zero ICI advantage which pays off.

By this time, inclusion of FFT and CP in OFDM system and substantial advancements in *Digital Signal Processing* (DSP) technology made it an important part of telecommunications landscape. In the 1990s, OFDM was exploited for wideband data communications over mobile radio FM channels, *High-bit-rate Digital Subscriber Lines* (HDSL at 1.6Mbps), *Asymmetric Digital Subscriber Lines* (ADSL up to 6Mbps) and *Very-high-speed Digital Subscriber Lines* (VDSL at 100Mbps).

Digital Audio Broadcasting (DAB) was the first commercial use of OFDM technology. Development of DAB started in 1987. By 1992, DAB was proposed and the standard was formulated in 1994. DAB services came to reality in 1995 in UK and Sweden. The development of *Digital Video Broadcasting* (DVB) was started in 1993. DVB along with *High-Definition TeleVision* (HDTV) terrestrial broadcasting standard was published in 1995. At the dawn of the 20th century, several *Wireless Local Area Network* (WLAN) standards adopted OFDM on their physical layers. Development of European WLAN standard HiperLAN started in 1995. HiperLAN/2 was defined in June 1999 which adopts OFDM in physical layer. Recently IEEE 802.11a in USA has also adopted OFDM in their PHY layer.

Perhaps of even greater importance is the emergence of this technology as a competitor for future *4th Generations* (4G) wireless systems. These systems, expected to emerge by the year 2010, promise to at last deliver on the wireless Nirvana of anywhere, anytime, anything communications. Should OFDM gain prominence in this arena, and telecom giants are banking on just this scenario, then OFDM will become the technology of choice in most wireless links worldwide [9].

[2] B. R. Saltzberg, *Performance of an Efficient Parallel Data Transmission System*, IEEE Transactions on Communications, COM-15 (6), pp. 805-811, December 1967.

[3] S. B. Weinstein, P. M. Ebert, *Data Transmission of Frequency Division Multiplexing Using The Discrete Frequency Transform*, IEEE Transactions on Communications, COM-19(5), pp. 623-634, October 1971.

[4] R. Peled & A. Ruiz, *Frequency Domain Data Transmission Using Reduced Computational Complexity Algorithms*, in Proceeding of the IEEE International Conference on Acoustics, Speech, and Signal Processing, ICASSP '80, pp. 964-967, Denver, USA, 1980.

3.2 OFDM Transceiver Systems

A complete OFDM transceiver system is described in Figure 3.1. In this model, *Forward Error Control/Correction* (FEC) coding and interleaving are added in the system to obtain the robustness needed to protect against burst errors (see Section 3.3 for details). An OFDM system with addition of channel coding and interleaving is referred to as *Coded OFDM* (COFDM).

In a digital domain, binary input data is collected and FEC coded with schemes such as convolutional codes. The coded bit stream is interleaved to obtain diversity gain. Afterwards, a group of channel coded bits are gathered together (1 for BPSK, 2 for QPSK, 4 for QPSK, etc.) and mapped to corresponding constellation points. At this point, the data is represented in complex numbers and they are in serial. Known pilot symbols mapped with known mapping schemes can be inserted at this moment. A serial to parallel converter is applied and the IFFT operation is performed on the parallel complex data. The transformed data is grouped together again, as per the number of required transmission subcarriers. Cyclic prefix is inserted in every block of data according to the system specification and the data is multiplexed to a serial fashion. At this point of time, the data is OFDM modulated and ready to be transmitted. A *Digital-to-Analog Converter* (DAC) is used to transform the time domain digital data to time domain analog data. RF modulation is performed and the signal is up-converted to transmission frequency.

After the transmission of OFDM signal from the transmitter antenna, the signals go through all the anomaly and hostility of wireless channel. After the receiving the signal, the receiver down-converts the signal; and converts to digital domain using *Analog-to-Digital Converter* (ADC). At the time of down-conversion of received signal, carrier frequency synchronization is performed. After ADC conversion, symbol timing synchronization is achieved. An FFT block is used to demodulate the OFDM signal. After that, channel estimation is performed using the demodulated pilots. Using the estimations, the complex received data is obtained which are demapped according to the transmission constellation diagram. At this moment, FEC decoding and de-interleaving are used to recover the originally transmitted bit stream.

3.3 Channel Coding and Interleaving

Since OFDM carriers are spread over a frequency range, there still may be some frequency selective attenuation on a time varying basis. A deep fade on a particular frequency may cause the loss of data on that frequency for that given time, thus some of the subcarriers can be strongly attenuated and that will cause burst errors. In these situations, FEC in COFDM can fix the errors [10]. An efficient FEC coding in flat fading situations leads to a very high coding gain, especially if soft decision decoding is applied. In a single carrier modulation, if such a deep fade occurs, too many consecutive symbols may be lost and FEC may not be too effective in recovering the lost data [11].

Experiences show that basic OFDM system is not able to obtain a BER of 10^{-5} or 10^{-6} without channel coding. Thus, all OFDM systems now-a-days are converted to COFDM. The benefits of COFDM are two-fold in terms of performance improvement. First, the benefit that the channel coding brings in, that is the robustness to burst error. Secondly, interleaving brings frequency diversity. The interleaver ensures that adjacent outputs from channel encoder are placed far apart in frequency domain. Specifically for a rate encoder, the channel encoder provides two output bits for one source bit. When they are placed far apart from each other (i.e. placed on subcarriers that are far from each other in frequency domain), then they experience

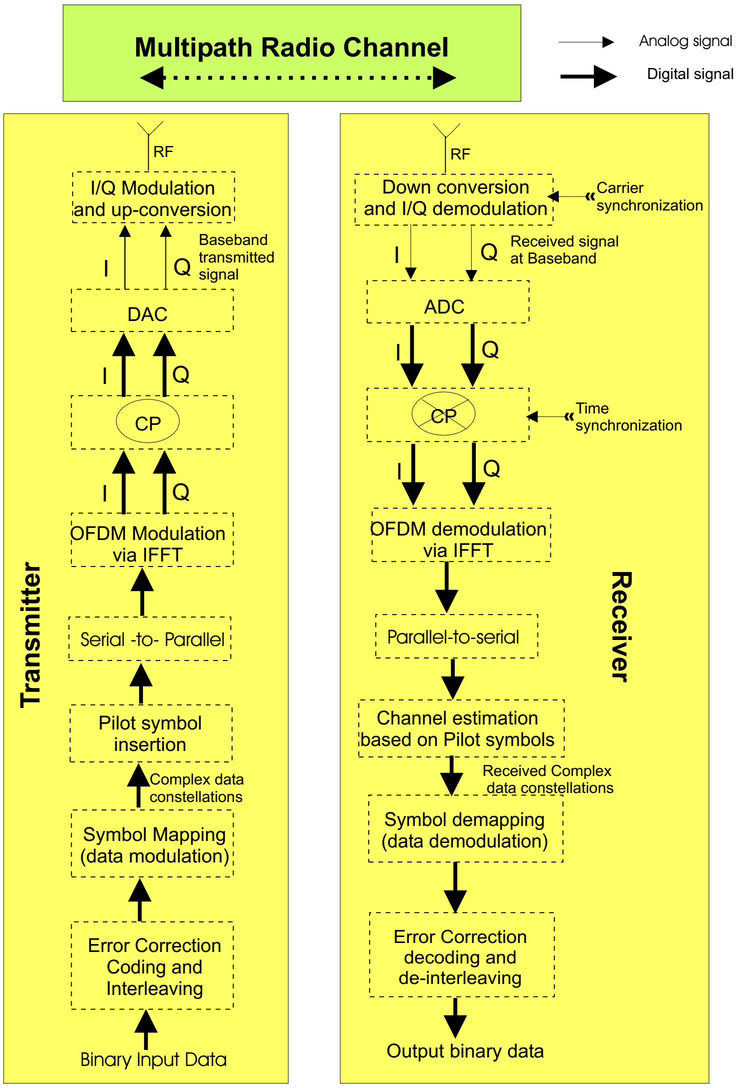

Figure 3.1: OFDM Transceiver Model

Data rate (Mbps)	Modulation scheme	Coding rate	Coded bits per subcarrier	Code bits per OFDM symbol	Data bits per OFDM symbol
6	BPSK	$\frac{1}{2}$	1	48	24
9	BPSK	$\frac{3}{4}$	1	48	36
12	QPSK	$\frac{1}{2}$	2	96	48
18	QPSK	$\frac{3}{4}$	2	96	72
24	16-QAM	$\frac{1}{2}$	4	192	96
36	16-QAM	$\frac{3}{4}$	4	192	144
48	64-QAM	$\frac{2}{3}$	6	288	192
54	64-QAM	$\frac{3}{4}$	6	288	216

Table 3.1: IEEE 802.11a OFDM PHY Modulation Techniques

unique gain (and/or unique fade). It is very unlikely that both of the bits will face a deep fade, and thus at least one of the bits will be received intact on the receiver side, and as a result, overall BER performance will improve [9].

According to Table 3.1, IEEE 802.11a standard offers wide variety of choices for coding and modulation, this allows a chance of making trade-offs for lot of considerations. The standard enables several data rates by making use of different combinations of modulation and channel coding scheme. It is worth mentioning here that the standard demands all 802.11a complaint products to support all the data rates. Table 3.1 presents the different arrangement of modulation and coding scheme that are used to obtain the data rates [12].

3.4 Advantages of OFDM System

3.4.1 Combating ISI and Reducing ICI

When signal passes through a time-dispersive channel, the orthogonality of the signal can be jeopardized. CP helps to maintain orthogonality between the sub carriers. Before CP was invented, guard interval was proposed as the solution. Guard interval was defined by an empty space between two OFDM symbols, which serves as a buffer for the multipath reflection. The interval must be chosen as larger than the expected maximum delay spread, such that multi path reflection from one symbol would not interfere with another. In practice, the empty guard time introduces ICI. ICI is crosstalk between different subcarriers, which means they are no longer orthogonal to each other [6]. A better solution was later found, that is cyclic extension of OFDM symbol or CP. CP is a copy of the last part of OFDM symbol which is appended to front the transmitted OFDM symbol [13].

CP still occupies the same time interval as guard period, but it ensures that the delayed replicas of the OFDM symbols will always have a complete symbol within the FFT interval (often referred as FFT window); this makes the transmitted signal periodic. This periodicity plays a very significant role as this helps maintaining the orthogonality. The concept of being able to do this, and what it means, comes from the nature of IFFT/FFT process. When the IFFT is taken for a symbol period during OFDM modulation, the resulting time sample process

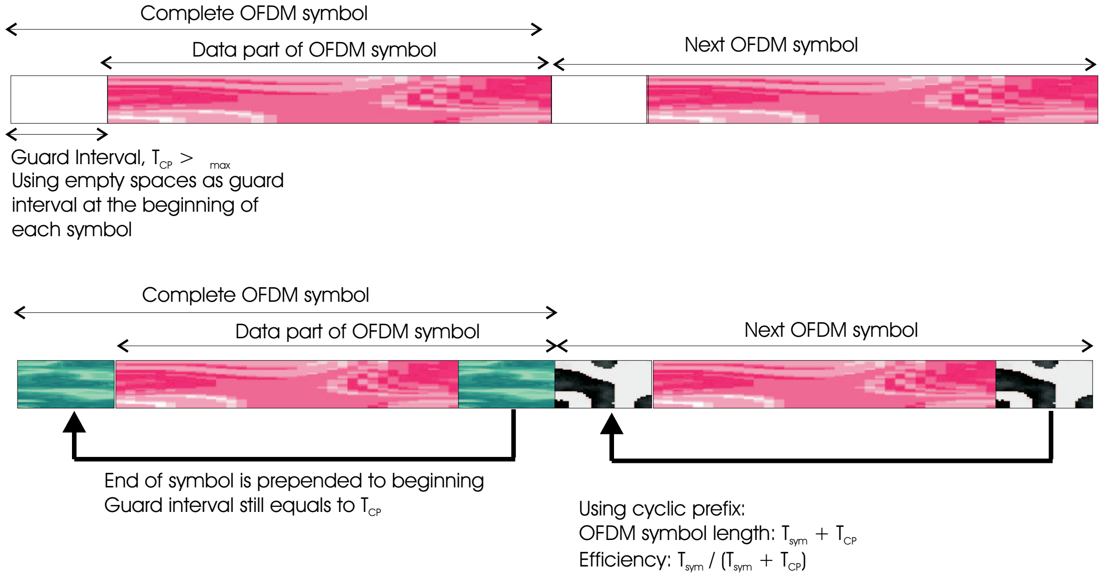

Figure 3.2: Role of Guard Intervals and Cyclic Prefix in Combatting ISI and ICI

is technically periodic. In a Fourier transform, all the resultant components of the original signal are orthogonal to each other. So, in short, by providing periodicity to the OFDM source signal, CP makes sure that subsequent subcarriers are orthogonal to each other.

At the receiver side, CP is removed before any processing starts. As long as the length of CP interval is larger than maximum expected delay spread τ_{max}, all reflections of previous symbols are removed and orthogonality is restored. The orthogonality is lost when the delay spread is larger than length of CP interval. Inserting CP has its own cost, we loose a part of signal energy since it carries no information. The loss is measured as

$$SNR_{loss\_CP} = -10 \log_{10} \left(1 - \frac{T_{CP}}{T_{sym}} \right) \tag{3.1}$$

Here, T_{CP} is the interval length of CP and T_{sym} is the OFDM symbol duration. It is understood that although we loose part of signal energy, the fact that zero ICI and ISI situation pay off the loss.

To conclude, CP gives two fold advantages, first occupying the guard interval, it removes the effect of ISI and by maintaining orthogonality it completely removes the ICI. The cost in terms signal energy loss is not too significant.

3.4.2 Spectral Efficiency

Figure 3.3 illustrates the different between conventional FDM and OFDM systems. In the case of OFDM, a better spectral efficiency is achieved by maintaining orthogonality between the subcarriers. When orthogonality is maintained between different subchannels during transmission, then it is possible to separate the signals very easily at the receiver side. Classical FDM ensures this by inserting guard bands between sub channels. These guard bands keep the subchannels far enough so that separation of different subchannels are possible. Naturally inserting guard bands results to inefficient use of spectral resources.

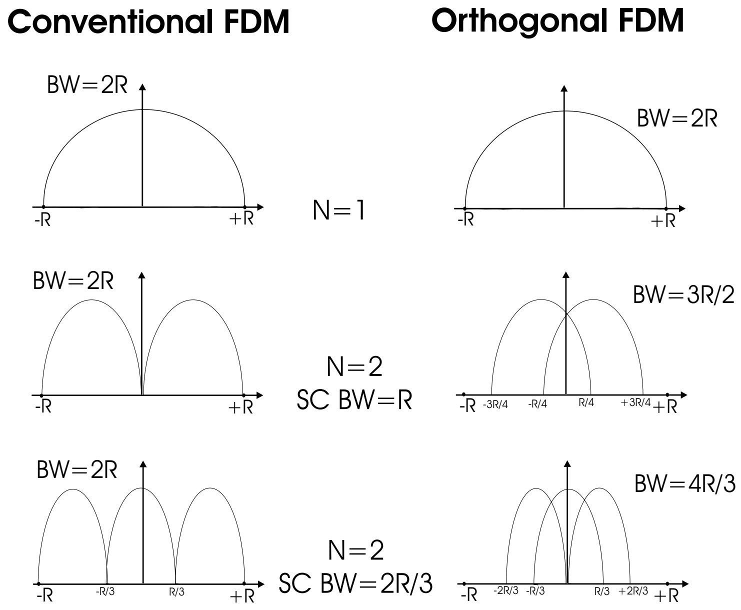

Figure 3.3: Spectrum Efficiency of OFDM Compared to Conventional FDM

Orthogonality makes it possible in OFDM to arrange the subcarriers in such a way that the sidebands of the individual carriers overlap and still the signals are received at the receiver without being interfered by ICI. The receiver acts as a bank of demodulator, translating each subcarrier down to DC, with the resulting signal integrated over a symbol period to recover raw data. If the other subcarriers all down converted to the frequencies that, in the time domain, have a whole number of cycles in a symbol period T_{sym}, then the integration process results in zero contribution from all other carriers. Thus, the subcarriers are linearly independent (i.e., orthogonal) if the carrier spacing is a multiple of $\frac{1}{T_{sym}}$ [14].

3.4.3 Some Other Benefits of OFDM System

1. The beauty of OFDM lies in its simplicity. One trick of the trade that makes OFDM transmitters low cost is the ability to implement the mapping of bits to unique carriers via the use of IFFT [9].

2. Unlike CDMA, OFDM receiver collects signal energy in frequency domain, thus it is able to protect energy loss at frequency domain.

3. In a relatively slow time-varying channel, it is possible to significantly enhance the capacity by adapting the data rate per subcarrier according to SNR of that particular subcarrier [6].

4. OFDM is more resistant to frequency selective fading than single carrier systems.

5. The OFDM transmitter simplifies the channel effect, thus a simpler receiver structure is enough for recovering transmitted data. If we use coherent modulation schemes, then very simple channel estimation (and/or equalization) is needed, on the other hand, we need no channel estimator if differential modulation schemes are used.

6. The orthogonality preservation procedures in OFDM are much simpler compared to CDMA or TDMA techniques even in very severe multipath conditions.

7. It is possible to use maximum likelihood detection with reasonable complexity [10].

8. OFDM can be used for high-speed multimedia applications with lower service cost.

9. OFDM can support dynamic packet access.

10. Single frequency networks are possible in OFDM, which is especially attractive for broadcast applications.

11. Smart antennas can be integrated with OFDM. MIMO systems and space-time coding can be realized on OFDM and all the benefits of MIMO systems can be obtained easily. Adaptive modulation and tone/power allocation are also realizable on OFDM.

3.5 Disadvantages of OFDM System

3.5.1 Strict Synchronization Requirement

OFDM is highly sensitive to time and frequency synchronization errors, especially at frequency synchronization errors, everything can go wrong [15]. Demodulation of an OFDM signal with an offset in the frequency can lead to a high bit error rate.

The source of synchronization errors are two; first one being the difference between local oscillator frequencies in transmitter and receiver, secondly relative motion between the transmitter and receiver that gives Doppler spread. Local oscillator frequencies at both points must match as closely as they can. For higher number of subchannels, the matching should be even more perfect. Motion of transmitter and receiver causes the other frequency error. So, OFDM may show significant performance degradation at high-speed moving vehicles [8].

To optimize the performance of an OFDM link, accurate synchronization is a prime importance. Synchronization needs to be done in three factors: symbol, carrier frequency and sampling frequency synchronization. A good description of synchronization procedures is given in [16]. We have discussed the synchronization issues in detail in Section **??**.

3.5.2 Peak-to-Average Power Ratio(PAPR)

Peak to Average Power Ratio (PAPR) is proportional to the number of sub-carriers used for OFDM systems. An OFDM system with large number of sub-carriers will thus have a very large PAPR when the sub-carriers add up coherently. Large PAPR of a system makes the implementation of Digital-to-Analog Converter (DAC) and Analog-to-Digital Converter(ADC) to be extremely difficult. The design of RF amplifier also becomes increasingly difficult as the PAPR increases.

There are basically three techniques that are used at present to reduce PAPR, they are *Signal Distortion Techniques*, *Coding Techniques* and finally the *Scrambling Technique*. Since OFDM is characterized by

$$x(t) = \frac{1}{\sqrt{N}} \sum_{n=1}^{N} a_n e^{jw_n t}. \tag{3.2}$$

Here a_n is the modulating signal. For Large number of a_n both the real and imaginary parts tend to be Gaussian distributed, thus the amplitude of the OFDM symbol has a Rayleigh distribution, while the power distribution is central chi squared.

The clipping and windowing technique reduces PAPR by non-linear distortion of the OFDM signal. It thus introduces self interference as the maximum amplitude level is limited to a fixed level. It also increases the out of band radiation, but this is the simplest method to reduce the PAPR. To reduce the error rate, additional Forward error correcting codes can be used in conjunction with the the clipping and windowing method.

Another technique called *Linear Peak Cancellation* can also be used to reduce the PAPR. In this method, time shifted and scaled reference function is subtracted from the signal, such that each subtracted reference function reduces the peak power of at least one signal sample. By selecting an appropriate reference function with approximately the same bandwidth as the transmitted function, it can be assured that the peak power reduction does not cause out of band interference. One example of a suitable reference function is a *raised cosine window*. Detailed discussion about coding methods to reduce PAPR can be found in [6].

3.5.3 Co-Channel Interference in Cellular OFDM

In cellular communications systems, CCI is combated by combining adaptive antenna techniques, such as sectorization, directive antenna, antenna arrays, etc. Using OFDM in cellular systems will give rise to CCI. Similarly with the traditional techniques, with the aid of beam steering, it is possible to focus the receiver's antenna beam on the served user, while attenuating the co-channel interferers. This is significant since OFDM is sensitive to CCI.

3.6 OFDM System Design Issues

System design always needs a complete and comprehensive understanding and consideration of critical parameters. OFDM system design is of no exception, it deals with some critical, and often conflicting parameters. Basic OFDM philosophy is to decrease data rate at the subcarriers, so that the symbol duration increases, thus the multipaths are effectively removed. This poses a challenging problem, as higher value for CP interval will give better result, but it will increase the loss of energy due to insertion of CP. Thus, a tradeoff between these two must be obtained for a reasonable design.

3.6.1 OFDM System Design Requirements

OFDM systems depend on four system requirement:

- **Available bandwidth:** Bandwidth is always the scarce resource, so the mother of the system design should be the available for bandwidth for operation. The amount of bandwidth will play a significant role in determining number of subcarriers, because with a large bandwidth, we can easily fit in large number of subcarriers with reasonable guard space.

- **Required bit rate:** The overall system should be able to support the data rate required by the users. For example, to support broadband wireless multimedia communication, the system should operate at more than 10 Mbps at least.

- **Tolerable delay spread:** Tolerable delay spread will depend on the user environment. Measurements show that indoor environment experiences maximum delay spread of few hundreds of $n$sec at most, whereas outdoor environment can experience up to 10μs. So the length of CP should be determined according to the tolerable delay spread.

- **Doppler values:** Users on a high speed vehicle will experience higher Doppler shift where as pedestrians will experience smaller Doppler shift. These considerations must be taken into account.

3.6.2 OFDM System Design Parameters

The design parameters are derived according to the system requirements. The requirement of the system design must be fulfilled by the system parameters. Following are the design parameters for an OFDM system [6]:

- **Number of subcarriers:** Increasing number of subcarriers will reduce the data rate via each subcarrier, which will make sure that the relative amount of dispersion in time caused by multipath delay will be decreased. But when there are large numbers of subcarriers, the synchronization at the receiver side will be extremely difficult.

- **Guard time (CP interval) and symbol duration:** A good ratio between the CP interval and symbol duration should be found, so that all multipaths are resolved and not significant amount of energy is lost due to CP. As a thumb rule, the CP interval must be two to four times larger than the *Root-Mean-Square* (RMS) delay spread. Symbol duration should be much larger than the guard time to minimize the loss of SNR, but within reasonable amount. It cannot be arbitrarily large, because larger symbol time means that more subcarriers can fit within the symbol time. More subcarriers increase the signal processing load at both the transmitter and receiver, increasing the cost and complexity of the resulting device [17].

- **Subcarrier spacing:** Subcarrier spacing must be kept at a level so that synchronization is achievable. This parameter will largely depend on available bandwidth and the required number of subchannels.

- **Modulation type per subcarrier:** This is trivial, because different modulation scheme will give different performance. Adaptive modulation and bit loading may be needed depending on the performance requirement. It is interesting to note that the performance of OFDM systems with differential modulation compares quite well with systems using non-differential and coherent demodulation [18]. Furthermore, the computation complexity in the demodulation process is quite low for differential modulations.

- **FEC coding:** Choice of FEC code will play a vital role also. A suitable FEC coding will make sure that the channel is robust to all the random errors.

Chapter 4

OFDM System Model

In this chapter, an analytical model for OFDM scheme is presented. The first section provides the scope of the chapter, and presents an overview of a generic OFDM system. The subsystems of interest to the modelling are then found. The second section contains an analytical time-domain model for an OFDM system, disregarding the different multiple-access techniques, as we are here focussing on the OFDM system as a basic transmission scheme. It should be fairly simple to extend this model to multi-user scenario.

The analytical model that is explained in this chapter is taken from these author's another work in [19].

4.1 System

In this deliverable, several OFDM-based multiple-access schemes are described in a downlink (DL) scenario. In the DL scenario, the base station transmits a signal (containing data from several users) to the different mobile stations. At the mobile station, only one user signal is extracted. This is illustrated in Figure 4.1, where data streams from several users are transmitted over the same OFDM carrier signal. The source bits from each user are source encoded, and then digitally modulated (i.e. M bits are grouped and used to determine the constellation point). The resource mapping combines the different user signals in a manner dependent on the multiple-access scheme used, be it assigning different users to different subcarriers (as in OFDMA) or using CDMA techniques (such as in MC-CDMA). Following the resource mapping, the resulting subcarrier sequence is processed in the OFDM block, where the signal is converted from the subcarrier domain to the time-domain and the cyclic prefix (CP) is added. The signal is then converted to a carrier frequency and transmitted over the channel.

In the receiver, the signal is down-converted and each OFDM symbol is processed, i.e. the CP is removed and the signal is converted into the subcarrier domain. The OFDM block also estimates the channel transfer function in order to compensate for fading. Following OFDM processing, the user signals are separated and the signal for the u^{th} user is extracted.

The functional block that are of interest in this chapter are boxed in Figure 4.1. The analytical model presented in Section 4.2 only concerns the OFDM modulation/demodulation and the channel, as this processing is common to any multiple-access schemes on top of basic OFDM system.

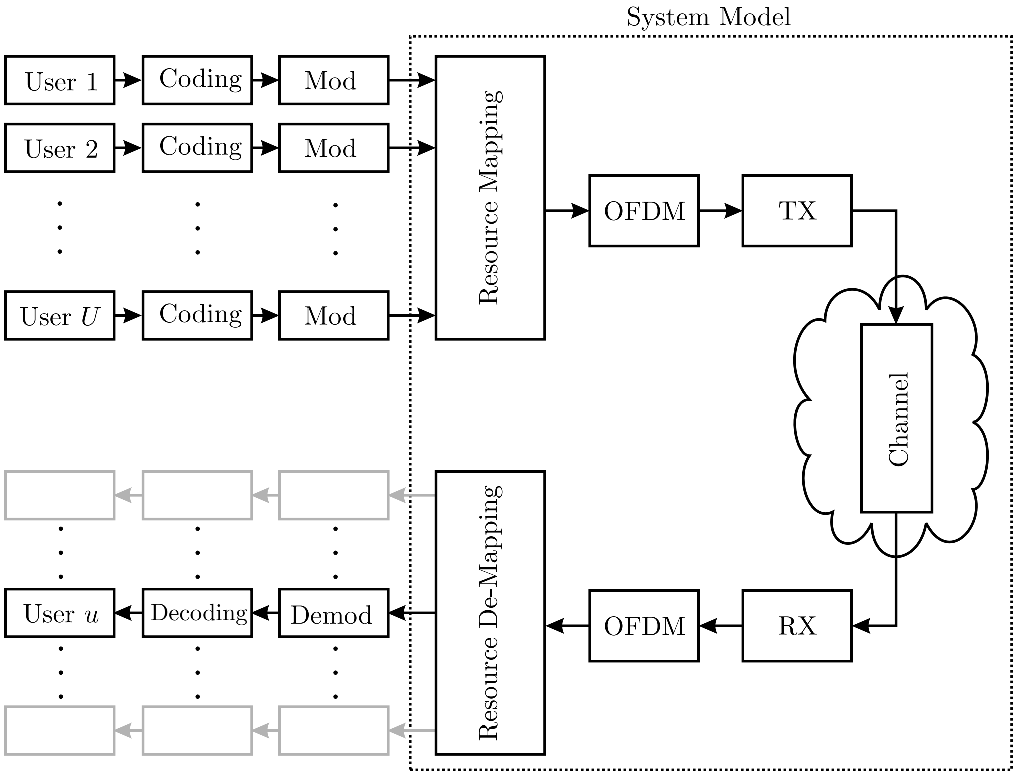

Figure 4.1: Generic OFDM system downlink diagram. Several users are multiplexed onto the same OFDM carrier signal. The OFDM signal is transmitted via the channel to the receiver, where only one user signal is extracted. The blocks inside the box are the focus of this chapter.

4.2 Analytical model of OFDM System

In this section, an analytical time-domain model of an OFDM transmitter and receiver, as well as a channel model, are derived.

4.2.1 Transmitter

The s^{th} OFDM symbol is found using the s^{th} subcarrier block, $\mathbf{X}_s[k]$. In practice, the OFDM signal is generated using an inverse DFT. In the following model, the transmitter is assumed ideal, i.e. sampling or filtering do not affect the signal on the transmitter side. Therefore, a continuous transmitter output signal may be constructed directly using a Fourier series representation within each OFDM symbol interval.

Each OFDM symbol contains N subcarriers, where N is an even number (frequently a power of two). The OFDM symbol duration is T_u seconds, which must be a whole number of periods for each subcarrier. Defining the subcarrier spacing as $\Delta\omega$, the shortest duration that meets this requirement is written as:

$$T_u = \frac{2\pi}{\Delta\omega} \Leftrightarrow \Delta\omega = \frac{2\pi}{T_u} = 2\pi\Delta f \tag{4.1}$$

Using this relation, the spectrum of the Fourier series for the duration of the s^{th} OFDM symbol is written as:

$$\mathbf{X}_s(\omega) = \sum_{k=-N/2}^{N/2-1} \mathbf{X}_s[k]\delta_c(\omega - k\Delta\omega) \tag{4.2}$$

In order to provide the OFDM symbol in the time-domain, the spectrum in (4.2) is inverse Fourier transformed and limited to a time interval of T_u. The time-domain signal, $\tilde{x}_s(t)$, is therefore written as:

$$\begin{aligned}\tilde{x}_s(t) &= \mathcal{F}\left\{\mathbf{X}_s(\omega)\right\}\Xi_{T_u}(t) \\ &= \begin{cases} \frac{1}{\sqrt{T_u}}\sum_{k=-N/2}^{N/2-1}\mathbf{X}_s[k]e^{j\Delta\omega kt} & 0 \le t < T_u \\ 0 & \text{otherwise}\end{cases} \quad (4.3)\end{aligned}$$

$$(4.4)$$

where Ξ_{T_u} is a unity amplitude rectangular gate pulse of duration T_u. Following the frequency-to time-domain conversion, the signal is extended, and the cyclic prefix is added:

$$\tilde{x}'_s(t) = \begin{cases} \tilde{x}_s(t + T_u - T_g) & 0 \le t < T_g \\ \tilde{x}_s(t - T_g) & T_g < t < T_s \\ 0 & \text{otherwise}\end{cases} \tag{4.5}$$

where T_g is the cyclic prefix duration and $T_s = T_u + T_g$ is the total OFDM symbol duration. It should be noted, that (4.5) has the following property:

$$\tilde{x}'_s(t) = \tilde{x}'_s(t + T_u) \Leftrightarrow 0 \le t < T_g \tag{4.6}$$

that is, a periodicity property within the interval $[0, T_g]$. The transmitted complex baseband signal, $\tilde{s}(t)$, is formed by concatenating all OFDM symbols in the time-domain:

$$\tilde{s}(t) = \sum_{s=0}^{S-1} \tilde{x}'_s(t - sT_s) \tag{4.7}$$

This signal is finally upconverted to a carrier frequency and transmitted:

$$s(t) = \Re e \left\{ \tilde{s}(t) e^{j2\pi f_c t} \right\} \tag{4.8}$$

where $s(t)$ denotes the transmitted RF signal and f_c is the RF carrier frequency. For frequency hopping systems, the carrier frequency is changed at certain intervals. This is written as:

$$f_c[s] = f_{c,0} + f_h[s] \tag{4.9}$$

where $f_c[s]$ is the carrier frequency for the s^{th} OFDM symbol, $f_{c,0}$ is the center frequency of the band and $f_h[s]$ is the frequency deviation from the band center when transmitting the s^{th} OFDM symbol. The period of $f_h[s]$ is χ, where χ is the hopping sequence period measured in whole OFDM symbols.

The transmitter model described in this section is illustrated in Figure 4.2.

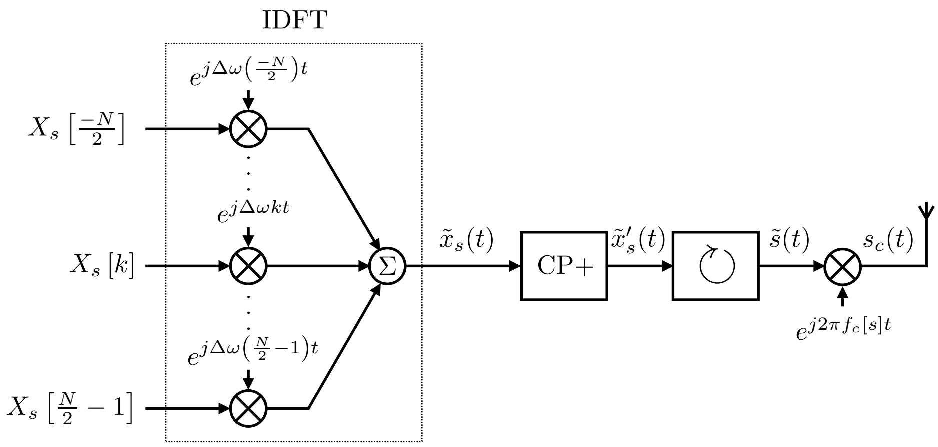

Figure 4.2: Transmitter diagram for the OFDM analytical model, given by (4.1)–(4.9). The subcarriers for the s^{th} OFDM symbol each modulate a carrier, which are separated by $\Delta\omega$. The resulting waveforms are then summed, and the CP is added. The symbol ↻ represents the concatenation of the OFDM symbols, given by (4.7). The resulting signal is then converted to a carrier frequency and transmitted.

4.2.2 Channel

The channel is modeled as a time-domain complex-baseband transfer function, which may then be convolved with the transmitted signal to determine the signal at the receiver side. The channel baseband equivalent impulse response function for the u^{th} user, $\tilde{h}_u(t)$ is defined as:

$$\tilde{h}_u(\tau, t) = \sum_{l=0}^{L} h_{u,l}(t)\delta_c(\tau - \tau_l) \tag{4.10}$$

where $h_{u,l}(t)$ is the complex gain of the l^{th} multipath component for the u^{th} user at time t. The channel is assumed to be static for the duration of one OFDM symbol, and the path gain coefficients for each path contribution are assumed to be uncorrelated. No assumption is made

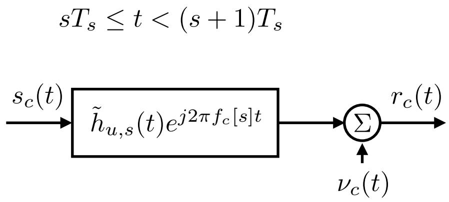

Figure 4.3: A diagram of the channel model given by (4.10)– (4.11). The transmitted signal passes through the channel, and noise is added.

for the autocorrelation properties of each path, except in the case of frequency hopping systems. In such systems, the channel is assumed to be completely uncorrelated between two frequency hops, provided that the distance in frequency is sufficiently large.

As the channel is assumed to be static over each OFDM symbol, (4.10) is redefined as:

$$\tilde{h}_{u,s}(t) = \sum_{l=0}^{L} h_{u,l}[s]\delta_c(t - \tau_l) \tag{4.11}$$

where

$$h_{u,l}[s] = h_{u,l}(t); \qquad sT_s \leq t < (s+1)T_s$$

The corresponding frequency-domain channel transfer function, $H_{u,s}$, can then be found using Fourier transformation:

$$\begin{aligned} H_{u,s}(\omega) &= \mathcal{F}\left\{\tilde{h}_{u,s}(t)\right\} \\ &= \int_{-\infty}^{\infty} \tilde{h}_{u,s}(t)e^{j\omega t}dt \end{aligned} \tag{4.12}$$

The time-domain channel model is illustrated in Figure 4.3.

4.2.3 Receiver

The signal at the receiver side consists of multiple echoes of the transmitted signal, as well as thermal (white gaussian) noise and interference. The RF signal received by the u^{th} user is written as:

$$r(t) = \Re e\left\{(\tilde{s}(t) * \tilde{h}_{u,s}(t))e^{j2\pi f_c[s]t}\right\} + \nu(t); \qquad sT_s \leq t < (s+1)T_s \tag{4.13}$$

where $\nu(t)$ is a real valued, passband signal combining additive noise and interference. The receiver now has to recreate the transmitted signal. Aside from noise and multipath effects, other imperfections in the receiver may also affect this process:

Timing error: In order to demodulate the signal, the receiver must establish the correct timing. This means that the receiver must estimate which time instant corresponds to $t = 0$ in the received signal (as seen from the transmitted signal point of view. As there are different uncertainties involved, a timing error of δt is assumed.

Frequency Error Similarly, the local oscillator of the receiver may oscillate at an angular frequency that is different from the angular frequency of the incoming signal. This difference is denoted $\delta\omega = 2\pi\delta f$.

The shifted time scale in the receiver is denoted $t' = t - \delta t$. Furthermore, due to the angular frequency error $\delta\omega$, the down-converted signal spectrum is shifted in frequency. The down-converted signal is therefore written as:

$$\tilde{r}(t) = (\tilde{s}(t') * \tilde{h}_{u,s}(t))e^{j\delta\omega t} + \tilde{\nu}(t'); \qquad sT_s \leq t < (s+1)T_s \tag{4.14}$$

where $\tilde{\nu}(t)$ is the complex envelope of the down-converted AWGN. The signal is divided into blocks of T_s each, and the CP is removed from each of them. The s^{th} received OFDM symbol block, $y'_s(t)$ is defined as:

$$\tilde{y}'_s(t) = \tilde{r}(t' - sT_s); \qquad 0 \leq t < T_s \tag{4.15}$$

The signal block corresponding to $\tilde{x}_s(t)$, $\tilde{y}_s(t)$ is found by removing the CP from the each $\tilde{y}'_s(t)$:

$$\tilde{y}_s(t) = \tilde{y}'_s(t + T_g); \qquad 0 \leq t < T_s - T_g \tag{4.16}$$

which can be rewritten as:

$$\begin{aligned}
\tilde{y}_s(t) &= \tilde{y}'_s(t + T_g); \qquad 0 \leq t \leq T_u \\
&= \tilde{r}(t' + T_g - sT_s) \\
&= (\tilde{s}(t' + T_g - sT_s) * \tilde{h}_{u,s}(t))e^{j\delta\omega t} + \tilde{\nu}(t' + T_g - sT_s) \\
&= (\tilde{x}'_s(t' + T_g) * \tilde{h}_{u,s}(t))e^{j\delta\omega t} + \tilde{\nu}_s(t') \\
&= (\tilde{x}_s(t') * \tilde{h}_{u,s}(t))e^{j\delta\omega t} + \tilde{\nu}_s(t')
\end{aligned} \tag{4.17}$$

where $\tilde{\nu}_s(t')$ is the noise signal block of duration T_u corresponding to the s^{th} OFDM symbol.

In order to recreate the transmitted subcarriers, N correlators are used, each one correlating the incoming signal with the k^{th} subcarrier frequency over an OFDM symbol period:

$$Y_s[k] = \frac{1}{\sqrt{T_u}} \int_0^{T_u} \tilde{y}_s(t')e^{j\Delta\omega kt}dt \tag{4.18}$$

In order to determine the correlator output, (4.18) may be seen as taking the continuous Fourier transform of (4.17) multiplied by the rectangular pulse $\Xi_{T_u}(t)$ and evaluating it at the corresponding subcarrier frequency. Assuming that the timing error is low enough to avoid ISI:

$$0 \leq \delta t < T_g - \max(\tau_l)$$

the continuous Fourier transform can be written as:

$$\begin{aligned}
Y_s(\omega) &= \mathcal{F}\left\{\tilde{y}_s(t)\Xi_{T_u}(t)\right\} \\
&= \mathcal{F}\left\{(\tilde{x}_s(t') * \tilde{h}_{u,s}(t))e^{j\delta\omega t} + \tilde{\nu}_s(t')\right\} * T_u e^{j\pi\frac{\omega}{\Delta\omega}}\operatorname{sinc}\left(\frac{\omega}{\Delta\omega}\right) \\
&= \mathcal{F}\left\{(\tilde{x}_s(t') * \tilde{h}_{u,s}(t))e^{j\delta\omega t}\right\} * T_u e^{j\pi\frac{\omega}{\Delta\omega}}\operatorname{sinc}\left(\frac{\omega}{\Delta\omega}\right) + N_s(\omega) \\
&= \mathcal{F}\left\{\tilde{x}_s(t') * \tilde{h}_{u,s}(t)\right\} * \delta_c(\omega - \delta\omega) * T_u e^{j\pi\frac{\omega}{\Delta\omega}}\operatorname{sinc}\left(\frac{\omega}{\Delta\omega}\right) + N_s(\omega) \\
&= e^{-j\omega\delta t}\mathcal{F}\left\{\tilde{x}_s(t) * \tilde{h}_{u,s}(t)\right\} * \delta_c(\omega - \delta\omega) * T_u e^{j\pi\frac{\omega}{\Delta\omega}}\operatorname{sinc}\left(\frac{\omega}{\Delta\omega}\right) + N_s(\omega) \\
&= e^{-j\omega(\delta t + \frac{\pi}{\Delta\omega})} \sum_{k'=N/2}^{N/2-1} \mathbf{X}_s[k']H_{u,s}(k'\Delta\omega)\operatorname{sinc}\left(\frac{\omega - k'\Delta\omega - \delta\omega}{\Delta\omega}\right) + N_s(\omega)
\end{aligned} \tag{4.19}$$

where

$$N_s(\omega) = \mathcal{F}\left\{\tilde{\nu}_s(t')\right\} * T_u e^{j\pi\frac{\omega}{\Delta\omega}}\text{sinc}\left(\frac{\omega}{\Delta\omega}\right) \tag{4.20}$$

is the Fourier transform of the AWGN contribution. The correlator output at the k^{th} correlator is then found as:

$$\begin{aligned} Y_s[k] &= Y_s(k\Delta\omega) \\ &= e^{-jk\Delta\omega(\delta t+\frac{\pi}{\Delta\omega})}\sum_{k'=N/2}^{N/2-1}\mathbf{X}_s[k']H_{u,s}(k'\Delta\omega)\text{sinc}\left(\frac{k\Delta\omega - k'\Delta\omega - \delta\omega}{\Delta\omega}\right) + N_s(k\Delta\omega) \end{aligned} \tag{4.21}$$

For zero frequency error, (4.21) reduces to:

$$Y_s[k] = e^{-jk\Delta\omega(\delta t+\frac{\pi}{\Delta\omega})}\mathbf{X}_s[k]H_{u,s}[k] + N_s[k]; \quad \delta\omega = 0 \tag{4.22}$$

where

$$N_s[k] = N_s(k\Delta\omega) \tag{4.23}$$

$$H_{u,s}[k] = H_{u,s}(k\Delta\omega) \tag{4.24}$$

From (4.21), it is seen that the k^{th} correlator output, $Y_s[k]$ corresponds to the transmitted subcarrier, $\mathbf{X}_s[k]$, with AWGN, ICI and a complex gain term (amplitude and phase shift) due to imperfect timing and channel effects. The analytical model for the receiver is illustrated in Figure 4.4.

When estimating the channel, the constant phase rotation term and the channel transfer function would be estimated jointly (as the receiver cannot discern between the two). In the following, the timing delay phase shift is omitted for clarity. Defining the equalization factor for the k^{th} subcarrier of the s^{th} OFDM symbol and u^{th} user as $Z_{u,s}[k]$, the subcarrier estimate is written as:

$$\begin{aligned} \hat{X}_s[k] &= Z_{u,s}[k]\,Y_s[k] \\ &= Z_{u,s}[k]\,H_{u,s}[k]\,\mathbf{X}_s[k] + Z_{u,s}[k]\,N_s[k] \end{aligned} \tag{4.25}$$

Assuming a zero-forcing, frequency-domain equalizer (as well as perfect channel estimation and zero frequency error), the corresponding equalizer gain is written as:

$$Z_{u,s}[k] = \frac{1}{H_{u,s}[k]}$$

and (4.25) is rewritten as:

$$\hat{X}_s[k] = \mathbf{X}_s[k] + \frac{N_s[k]}{H_{u,s}[k]} \tag{4.26}$$

It is observed, that although this is an unbiased estimator for $\mathbf{X}_s[k]$, the signal-to-noise ratio decreases drastically for subcarriers in deep fades.

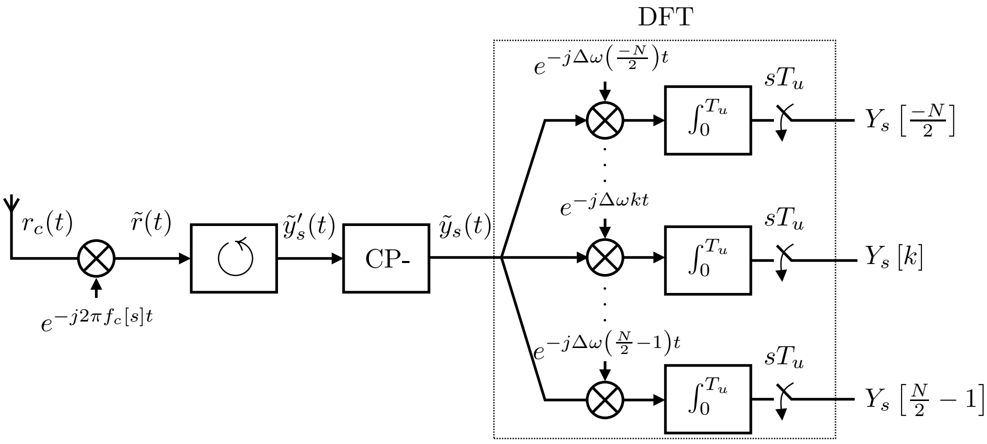

Figure 4.4: Receiver diagram for the OFDM analytical model, given by (4.13)–(4.22). The received signal (suffering from multipath effects and AWGN) is converted down to baseband. The symbol ⥁ represents the division of the received signal into blocks, given by (4.15). The CP is removed from each block, and the signal is then correlated with each subcarrier frequency, as shown by (4.18).

4.2.4 Sampling

Although the receiver may be modeled in the continuous time-domain, an OFDM receiver uses discrete signal processing to obtain the estimate of the transmitted subcarriers.

When the received signal is modeled as a Dirac impulse train, i.e. an ideally sampled signal, (4.17) is instead written as:

$$\tilde{y}_{s,d}(t) = \sum_{n=0}^{N-1} \tilde{y}_s[n]\delta_c(t - nT) \tag{4.27}$$

where:

$$T = \frac{T_u}{N} \tag{4.28}$$

is the sample duration and:

$$\tilde{y}_s[n] = \tilde{y}_s(nT); \quad n \in \{0, 1, ..., N-1\} \tag{4.29}$$

is the discrete sequence corresponding to the sampled values of $\tilde{y}_s(t)$. When (4.27) is inserted into (4.18), the correlation becomes the Discrete Fourier Transform of the received signal. It can be shown, however, that (4.21)–(4.26) are still valid in the discrete-time case.

4.3 Single OFDM Symbol Baseband Model in Matrix Notations

In this section, we explain the above analytical model in matrix model, so that it becomes easier to implement in simulation programs, such as in MATLAB simulations, we have to model all the components in the transmission chain in matrix format, thus the following model will be very useful in that regard. The baseband model for a single OFDM symbol s is shown in Figure 4.5.

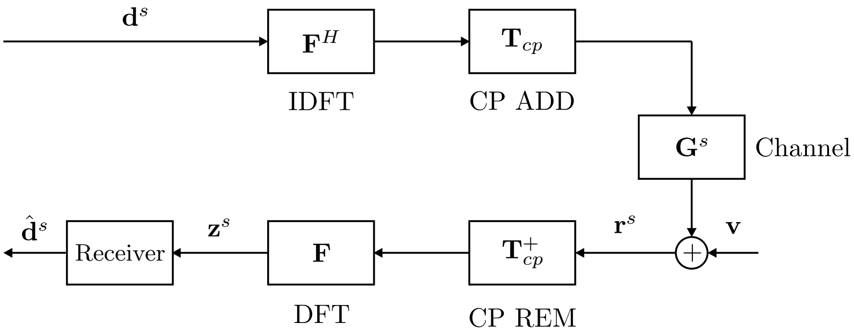

Figure 4.5: Single OFDM Symbol System Model

The single symbol received signal $\mathbf{r}^s \in \mathbb{C}^{[N_g+N+\delta_\mathbf{g} \times 1]}$ may be written as

$$\mathbf{r}^s = \mathbf{G}^s \mathbf{T}_{cp} \mathbf{F}^H \mathbf{d}^s + \mathbf{v} \tag{4.30}$$

with

$$\mathbf{T}_{cp} = \begin{array}{c} \begin{array}{cc} \overbrace{\hphantom{xxxxx}}^{N-N_g} & \overbrace{\hphantom{xxxxxxx}}^{N_g} \end{array} \\ \begin{bmatrix} & & 1 & & \\ & & & \ddots & \\ & & & & 1 \\ 1 & & & & \\ & \ddots & & & \\ & & 1 & & \\ & & & \ddots & \\ & & & & 1 \end{bmatrix} \end{array}$$

denoting the $[N_g + N \times N]$ cyclic prefix insertion matrix, where N_g is the number of samples in the guard period, and with

$$\mathbf{G}^s = \begin{bmatrix} g_n[0] & \mathbf{0} & \cdots & \cdots & \cdots & \cdots \\ g_{n+1}[1] & g_{n+1}[0] & \mathbf{0} & \ddots & \ddots & \ddots \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots \\ g_{n+L-1}[L-1] & g_{n+L-1}[L-2] & \cdots & g_{n+L-1}[0] & \ddots & \ddots \\ \mathbf{0} & \ddots & \ddots & \ddots & \ddots & \ddots \\ \vdots & \ddots & g_{n+N_g+N-1}[L-1] & \cdots & \cdots & g_{n+N_g+N-1}[0] \end{bmatrix}$$

denoting the $[N_g + N \times N_g + N]$ time-domain channel convolution matrix, where $g_n[l]$ represents the gain of the l^{th} sample delayed path in respect to the first path at time n of the time-varying channel impulse response, and where L is the span of samples from the first to the last considered path. The vector $\mathbf{v} \in \mathbb{C}^{[N_g+N\times 1]}$ represents complex valued circular symmetric white Gaussian noise with variance N_0, and the matrix $\mathbf{F}^H \in \mathbb{C}^{[N\times N]}$ corresponds to the IDFT operation and is the hermitian transposed of the DFT-matrix $\mathbf{F} \in \mathbb{C}^{[N\times N]}$.

The received vector $\mathbf{z}^s \in \mathbb{C}^{[N\times 1]}$ after the DFT-operation may be expressed as

$$\begin{aligned} \mathbf{z}^s &= \mathbf{F}\mathbf{T}_{cp}^{+}\mathbf{r}^s & (4.31)\\ &= \mathbf{F}\mathbf{T}_{cp}^{+}\mathbf{G}^s\mathbf{T}_{cp}\mathbf{F}^H\mathbf{d}^s + \mathbf{F}\mathbf{T}_{cp}^{+}\mathbf{v} & (4.32)\end{aligned}$$

with

$$\mathbf{T}_{cp}^{+} = \left[\begin{matrix} \underbrace{\qquad}_{N_g} & \underbrace{\begin{matrix} 1 & & \\ & \ddots & \\ & & 1 \end{matrix}}_{N} \end{matrix} \right]$$

denoting the $[N \times N_g + N]$ cyclic prefix removal matrix. Assuming that $N_g \geqq (L-1)$, the linear convolution of the transmitted sequence and the channel corresponds to a circular convolution. Assuming a time-invariant channel in the discrete time interval $[n; n + N_g + N - 1]$, the received vector $\mathbf{z}$ after the DFT-operation may be expressed as

$$\mathbf{z}^s = \mathbf{H}^s\mathbf{d}^s + \mathbf{n}, \tag{4.33}$$

with $\mathbf{H}^s \in \mathbb{C}^{[N\times N]}$ denoting the frequency domain diagonal channel matrix, where the $[k,k]^{\text{th}}$ element of $\mathbf{H}^s$ corresponds to the complex-valued channel gain of the k^{th} sub-carrier. The vector $\mathbf{n} \in \mathbb{C}^{[N\times 1]}$ represents complex valued circular symmetric white Gaussian noise with variance N_0. An estimate $\hat{\mathbf{d}}^s$ of the transmitted data symbols $\mathbf{d}^s$ may be calculated from $\mathbf{z}^s$ using various topologies as indicated in Figure 4.6.

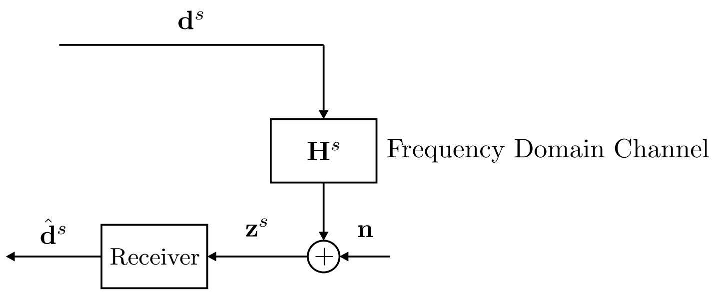

Figure 4.6: Simplified Single OFDM Symbol System Model

Chapter 5

Multi-Antenna OFDM Systems

Wireless transmission is impaired by fading and interference as we have discusses in Section 2. The increasing requirement of data rate and quality of service for wireless communications calls for new techniques to increase spectrum efficiency and to improve link quality [20]. OFDM has proved to be very effective in mitigating adverse multipath effects of a broadband wireless channel. The IEEE 802.11a WLAN standard specifies channel coding ($\frac{1}{2}$ rate convolutional coding with a constraint length of 7) and frequency interleaving to exploit the frequency diversity of the wideband channel, but this is efficient only if the channel is sufficiently frequency-selective, corresponding to long channel delay spreads. In a flat fading situation (or in relatively lesser frequency-selective fading situation), all or most subcarriers are attenuated simultaneously leading to long error bursts. In this case, frequency interleaving does not provide enough diversity to significantly improve the decoding performance [21]. So exploiting spatial diversity is necessary for any OFDM system [22].

Multiple Input Multiple Output (MIMO) technique has proved its potential by increasing the link capacity significantly via spatial multiplexing [23] and improving the link capacity via space-time coding [24]. Numerous research works are being published on MIMO enhanced OFDM based wireless systems. It is obvious that MIMO technique will be effectively used with OFDM based systems for providing mobile multimedia in future with reasonable data rate and quality of service (in terms bit error rate, BER).

5.1 Antenna Diversity

Diversity is the technique to improve link performance and/or increase data throughput by manipulating the statistical characteristics of the wireless link. There are different forms of diversity that are traditionally exploited in communications systems, such as temporal diversity in time-selective fading channels, spectral diversity in frequency-selective fading channels and spatial diversity in cases where the channel is neither time-selective nor frequency selective (i.e. when system constraints preclude the use of temporal or spectral forms of diversity, spatial diversity can be used to provide substantial improvement in system performance). For example, Interleaving makes use of temporal diversity; and spread spectrum communications and OFDM exploits spectral diversity.

Spatial diversity involves using of multiple antennas in transmitter and/or receiver. In a broad sense, the antenna diversity or spatial diversity can be classified in two categories: transmit diversity and receive diversity. The use of multiple antennas at the receiver is termed as receive

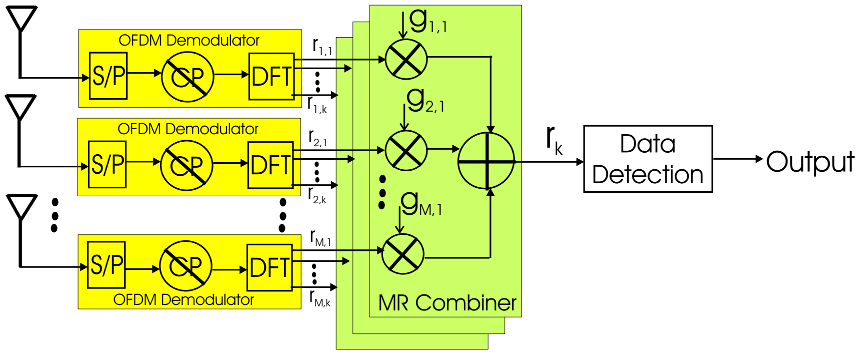

Figure 5.1: Multiple Antenna Receiver Diversity with MRC at subcarrier level

antenna diversity and the vice versa for transmitter.

5.1.1 Receiver Diversity

Receive diversity can be exploited fairly easily. In principle, multiple copies of transmitted signal are received at spatially separated antennas and they are appropriately combined according to specific signal processing (combining) techniques, such as *Selection Combining* (SC), *Maximum Ratio Combining* (MRC) or *Equal Gain Combining* (EGC). MRC is more complex compared to SC and EGC, but yields the highest Signal-to-Noise Ratio (SNR) [4]. As the number of receive antennas increases, the effective channel is seen as an additive Gaussian channel and the outage probability is driven to zero [25]. In case of OFDM reception, the selection combining can be separated into two different techniques, namely antenna selection combining (ASC) and subcarrier selection combining (SSC).

Here we discuss the MRC receiver diversity in an OFDM system. In MRC, the signals at the output of the receivers are linearly combined so as to maximize the instantaneous SNR. This is achieved by combining the co-phased signals, which requires that the CSI is known at the receiver. The SNR of the combined signal is equal to the sum of the SNRs of all the branch signals.

For an MRC-OFDM system as shown in Figure 5.1, the combining operations are performed at subcarrier level after the DFT operation, thus we denote the process as *Post-DFT MRC* or subcarrier combining receiver [26]. The received OFDM signals at different antenna branches are first transformed via M separate DFTs. Their outputs are assigned to N diversity combiners. In the linear combiner, the received signal of the m^{th} antenna and the k^{th} data subcarrier $r_{m,k}$ is multiplied by complex weight factors $g_{m,k}$, such that the signal branches are co-phased (i.e., all branches have zero phase). The magnitudes of these factors, having the squared sum of one, are assigned based on the instantaneous subcarrier power (or SNR) of each received signal branch. If the power is small in one particular branch, then a small gain factor will be assigned, and vice versa. The resultant signal envelope after adding up the branches is

$$r_k = \sum_{m=1}^{M} r_{m,k} g_{m,k} \tag{5.1}$$

In EGC, equal values are selected for the magnitudes of $g_{m,k}$.

In [27], a way to improve the radio link quality by introducing M number of receiver antennas in HiperLAN/2 model is presented. Signals from both antennas labelled as A and B, are demodulated and the values of the k data subcarriers in an OFDM symbol $R_{A,k}$ and $R_{B,k}$ are introduced in the combiner block. The combiner, according to a diversity algorithm, will merge the subcarrier values and the channel state information in order to form the signal R_k, which will pass through the channel equalizer and send to the inner receiver. The simulations in [27] show that MRC and SSC show a high reliability in dispersive channels. In another note, increasing the number of antennas also significantly improves the system performance, of course with the expense of higher hardware cost. It has been shown that the system performance drastically reduces under low correlation between transmission channels, and a good performance is obtained under highly correlated channels.

5.1.2 Transmitter Diversity

Receive diversity can be impractical in a number of applications. In those cases, using multiple antennas at transmitter, which is referred as transmit diversity, can be more attractive. Typical WLAN applications fall under this category. Transmit diversity assures that user mobile devices will become simple and cheaper [25].

S.M. Alamouti proposed a simple transmit diversity technique in [24] which was generalized by Vahid Tarokh et.el. to form the class of *Space-Time Block Codes* (STBC) [28]. Based on [24], [29] applied transmit diversity scheme to OFDM system with two transmit (Tx) antennas and one receive (Rx) antenna. [30] simulated an OFDM system with the transmit diversity scheme that [29] proposed. It was demonstrated that although this scheme has low complexity and is easy to implement, it provides considerable improvements in performance without any bandwidth expansion. Gains of between 9 to 14 dB were observed depending on the transmission mode and channel scenario at a BER of 10^{-4} for the case where two antennas are used in both the transmitter and receiver. We will discuss STBC in detail in Section 5.2.2.

One way of achieving transmit diversity is using a simple Delay Diversity (DD) scheme [31]. Using DD, the original signal is transmitted via the first antenna and (a) linearly time-delayed version(s) of original signal is (are) transmitted via one (or more) additional antenna elements. The limiting factor for such a diversity system is that the introduced delay is always shorter than the cyclic prefix (CP) to make sure that Inter-Symbol Interference (ISI) is avoided.

To overcome this limiting problem, Cyclic Delay Diversity (CDD) has been proposed in [32],[21],[22]. In this case, the signal is not truly delayed between respective antennas but cyclically shifted and thus, there are no restrictions for the delay times. The receiver structure in the DD and CDD schemes are similar to each other. All the signal processing needed is performed in time domain, so the duplication of the DFT operation for each receiving antenna branch is not a requirement any more, thus the receiver has lower computational cost compared to STBC and conventional diversity schemes. Moreover, it is possible to apply CDD as a transmit diversity technique without the knowledge of CSI at the transmitter side.

We will briefly discuss CDD as a mean for transmitter diversity in an OFDM system. Applying CDD in the transmitter randomizes a relatively flat channel, which is not unusual in indoor WLAN systems, in an efficient way. When we shift the OFDM signal cyclically and add them up in the receiver linearly, we actually insert some virtual echoes on the channel response. This effect increases the channel frequency-selectivity, thus higher order frequency diversity can be achieved, which is effectively exploited by a COFDM system [33]. On average, CDD does not increase or decrease the total number of errors in the received signal of an uncoded OFDM system, but it changes the error distribution in a beneficial way, yielding an increased coding gain.

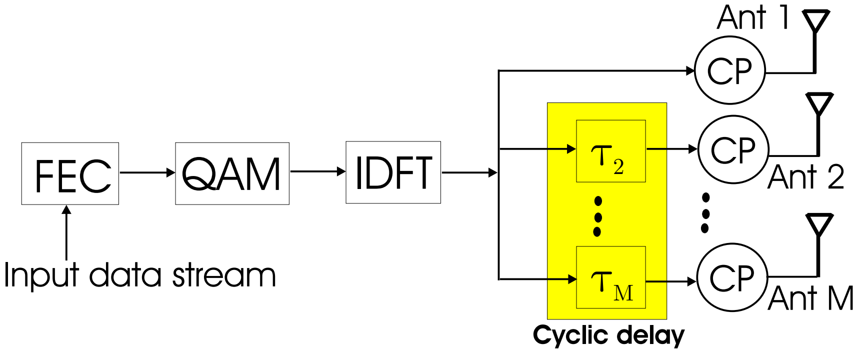

Figure 5.2: OFDM Transmitter with CDD; Cyclic shifts introduced in the original signal are fixed.

The transmitter diversity technique is presented in Figure 5.2. To prepare cyclically shifted transmission signal after the IDFT operation, a cyclic delay of τ_m (for $m = 2...M$) is introduced to the m^{th} antenna. M transmit antennas are available. The resulting CDD signal for m^{th} antenna branch is

$$x_m[n] = x[(n - \tau_m)modN] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} W_N^{-k\tau_m} d_k W_N^{kn} \tag{5.2}$$

τ_m is unique for all antenna branches. For the purpose of simple and efficient implementation, it is usually chosen to be an integer number of samples of the OFDM signal. τ_1 is taken to be zero, which means that the original signal is transmitted via the first antenna element. After inserting the cyclic delays, the CP is inserted in the conventional manner and the signals are transmitted via all the antennas simultaneously.

The optimum cyclic shift for transmitter CDD system has been analyzed in [34]. In that case, the channel is randomized as much as possible to introduce diversity via CDD. The CSI is not known at the transmitter, thus there is a possibility that the channel with CDD is worse than one of the branch channels.

5.1.3 Cyclic Delay Diversity in OFDM Receiver

Traditional receiver diversity combining schemes are explained in Section 5.1.1. Using any of these schemes in an OFDM system requires multiple DFT blocks in the receiver as seen in Figure 5.1. If the number of receiver antenna branches is M, then M DFT blocks are required. This is clearly an expensive arrangement. Thus, CDD as explained in Section 5.1.2 can be introduced at the receiver, where the diversity combining is performed prior to the DFT operation [33, Section 8.3], as shown in Figure 5.3. We can see that the receiver only requires one DFT block, regardless of the number of receive antenna branches. At the receiver, the antenna branch signals can be used for estimating the channel responses for each individual receiver antenna in order to optimize the diversity combining based on the instantaneous channel behavior. This allows for an optimized diversity combining using cyclic delays, τ_i and complex gain factors, g_i, where $i = 1, 2, ..., M$ and M is the number of diversity branches.

We denote this combining technique in the OFDM receiver as Pre-DFT Maximum Average (signal-to-noise) Ratio Combing (Pre-DFT MARC)[22]. For dual antenna system, where $g_1 = \sqrt{a}$ and $g_2 = \sqrt{1-a}$, $0 \leq a \leq 1$, are the magnitudes of the gain factors and $\tau_2 = n$ is the delay of

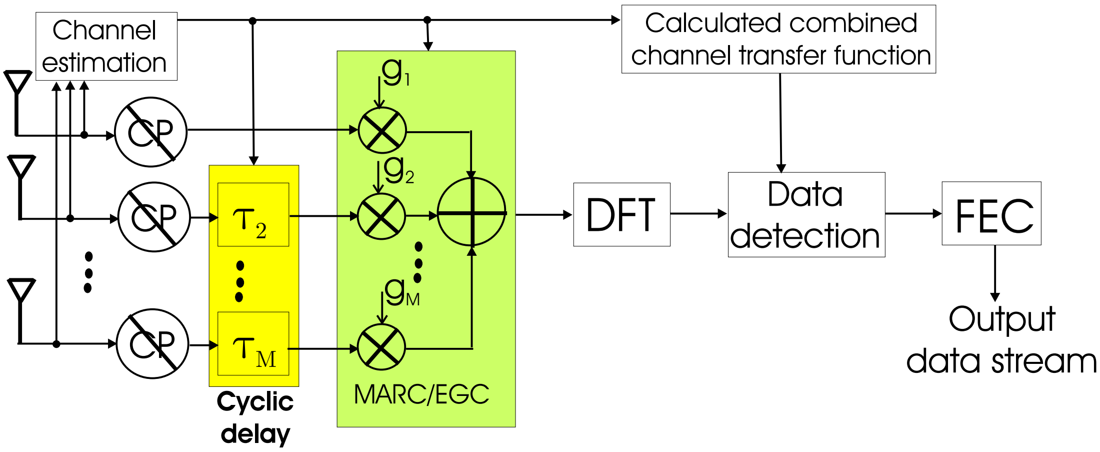

Figure 5.3: OFDM receiver with Pre-DFT Combining CDD. The instantaneous channel is estimated from the received signals to determine the optimum cyclic shifts (and gain factors, if MARC combining is performed).

branch two (in samples), we can obtain an expression that is proportional to the average SNR over all subcarriers in terms of a and n as [22]:

$$\begin{aligned} \mathrm{SNR}(a,n) &= a\mathbf{h}_1^H\mathbf{h}_1 + (1-a)\mathbf{h}_2^H\mathbf{h}_2 \\ &\quad + 2\sqrt{a}\sqrt{1-a}|\mathbf{h}_1^H\mathbf{W}_N^n\mathbf{h}_2|, \end{aligned} \tag{5.3}$$

where $\mathbf{h}_1$ and $\mathbf{h}_2$ are the channel transfer functions (CTF) of the 1^{st} and 2^{nd} diversity branches respectively, ^{H} denotes the conjugate complex transpose of a vector, N is the number of subcarriers, and $\mathbf{W}_N^n = \mathrm{diag}([W_N^0, W_N^n, W_N^{2n}, ..., W_N^{(N-1)n}])$, $W_N = e^{j2\pi/N}$, represents the cyclic delay. A phase rotation of the branch two signal by $\psi = -\angle(\mathbf{h}_1^H\mathbf{W}_N^n\mathbf{h}_2)$ leads to the result given in (5.3) [22]. The selection of a and n for maximizing the SNR (5.3) has been demonstrated in [22].

If we select equal values for the gain magnitudes (i.e. $g_1 = g_2 = \sqrt{0.5}$ for the dual antenna case), the combining technique is named as Pre-DFT Equal Gain Combining (Pre-DFT EGC).

A performance comparison is presented in [22] in terms of BER with and without coding, between MRC for receiver diversity and CDD for transmitter and receiver diversity in the context of IEEE 802.11a and/or HiperLAN/2 WLAN systems. CDD yields good diversity gains although the performance of Post-DFT MRC is not reached. On Ricean channels, the optimized schemes(i.e. Pre-DFT MARC as described in this section) perform much better than CDD with fixed delays(i.e. pure CDD at the transmitter as described in Section 5.1.2), while the gain in Rayleigh channels is surprisingly small.

5.2 MIMO Techniques

MIMO systems utilize space domain along with temporal and spectral domain to increase the capacity and link quality of wireless communications. The use of multiple antennas at both ends of a wireless link promises significant improvements in terms of spectral efficiency and link reliability. There are two techniques available in MIMO modelling, the first being *Spatial Multiplexing* (SM) and the other *Space-Time Coding* (STC).

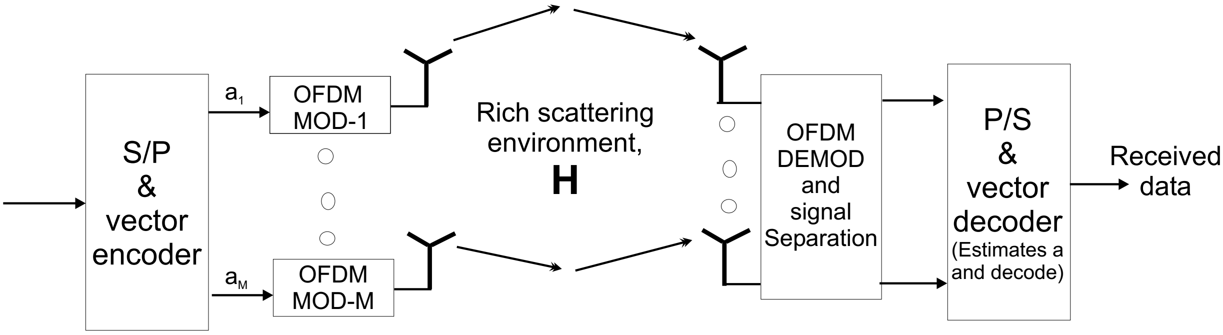

Figure 5.4: OFDM Based Spatial Multiplexing

5.2.1 Spatial Multiplexing Algorithms

Spatial multiplexing also known as the *Bell-labs LAyered Space-Time* (BLAST) system yields substantial increase in data rate in wireless radio link. Two types of BLAST algorithms are available; one is called *Vertical-BLAST* (VBLAST) [35] and the other is *Diagonal-BLAST* (DBLAST) [23]. V-BLAST as it is devised in [35] is presented in this section.

VBLAST increases data rate by transmitting independent information streams on different antennas. No channel knowledge at the transmitter is required. The capacity improvement is achieved by orthogonal channel matrix, H. In VBLAST, a single data stream is demultiplexed into M substreams, and each substream is then encoded into symbols and fed into respective transmitter. Tx-1 to Tx-M operate co-channel at symbol rate $\frac{1}{T}$ symbols/sec. Each transmitter is itself an ordinary QAM modulator. The receivers, Rx-1 to Rx-N, all operate in co-channel also, each receiving the signals from all M transmit antennas. A quasi stationary view point is taken into account, where the channel time variations are stationary for a burst period, and channel is estimated correctly for each burst period.

Narrowband model of V-BLAST MIMO channel can be expressed as:

$$\bar{H} = \begin{pmatrix} h_{1,1} & h_{1,2} & \dots & h_{1,M} \\ h_{2,1} & h_{2,2} & \dots & h_{2,M} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N,1} & h_{N,2} & \dots & h_{N,M} \end{pmatrix} \tag{5.4}$$

where the matrix transfer function of the channel is $\bar{H}^{[N,M]}$, $h_{i,j}$ is the complex transfer function from transmitter j to receiver i and $M \leq N$.

The input-output relationship of $[N, M]$ matrix channel is

$$\bar{r} = \bar{H}\bar{s} + \bar{n}, \tag{5.5}$$

where $\bar{r} = \begin{bmatrix} r_1 & r_2 & \dots & r_N \end{bmatrix}^T$ is a $[N, 1]$ received signal vector, $\bar{s} = \begin{bmatrix} s_1 & s_2 & \dots & s_M \end{bmatrix}^T$ is a $[M, 1]$ transmit signal vector, and $\bar{n} = \begin{bmatrix} n_1 & n_2 & \dots & n_N \end{bmatrix}^T$ is a $[N, 1]$ additive noise and interference vector where all components are random complex numbers.

The essential difference between D-BLAST and V-BLAST lies in the vector encoding process. In DBLAST, redundancy between the substreams is introduced through the use of specialized inter-substream block coding. The DBLAST code blocks are organized along diagonals in space-time. In V-BLAST the vector encoding process is simply a demultiplex operation followed by

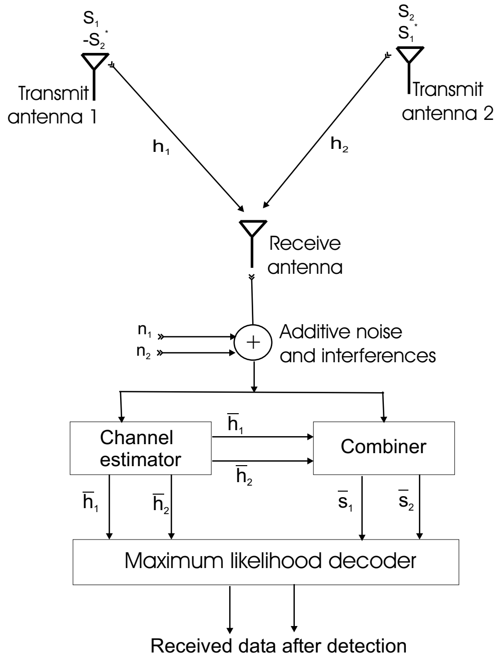

Figure 5.5: Alamouti's Space-Time Block Coding Scheme

independent bit-to-symbol mapping of each substream. No inter-substream coding, or coding of any kind is required, though conventional coding of individual substreams may certainly be applied [35].

Figure 5.4 presents an OFDM-based spatial multiplexing based on VBLAST algorithm. In an OFDM-based spatial multiplexing system, transmission takes place simultaneously from all M number of Tx antennas. The channel coded source data streams are first serial-to-parallel converted and vector encoded according to VBLAST algorithm. Later the respective symbol blocks are passed through OFDM modulator (OFDM MOD in Figure 5.4) and then OFDM symbol blocks are launched from the individual transmit antennas. In the receiver, the individual signals are OFDM demodulated, separated and then vector decoded to recover the original source data.

5.2.2 Space-Time Coding

[24] presented the transmitter diversity scheme for wireless communications which is a special form of STC. There are two forms STC available, *Space-Time Block Codes* (STBC) and *Space-Time Trellis Codes* (STTC). The original STC algorithm that [24] proposed is a two branch transmit diversity with one receiver, in another word it is a form of STBC [28].

This section describes the Alamouti 2Tx-1Rx (2 transmitters - 1 receiver) STBC scheme. In this scheme, two blocks of symbols namely s_1 and s_2 are transmitted at the same time from the transmitter antennas. In the first instance, s_1 and s_2 are transmitted from transmit antenna 1 (Tx1) and transmit antenna 2 (Tx2) respectively. In the following instance, $-s_2^*$ and s_1^* are sent from the antennas respectively (s_1^* means the complex conjugate of s_1). The channel between the Tx1 and receiver at time t,$h_1(t)$, can be modelled by a complex multiplicative distortion; similarly the channel between Tx2 and receiver can be termed as $h_2(t)$. Assuming that the multipath

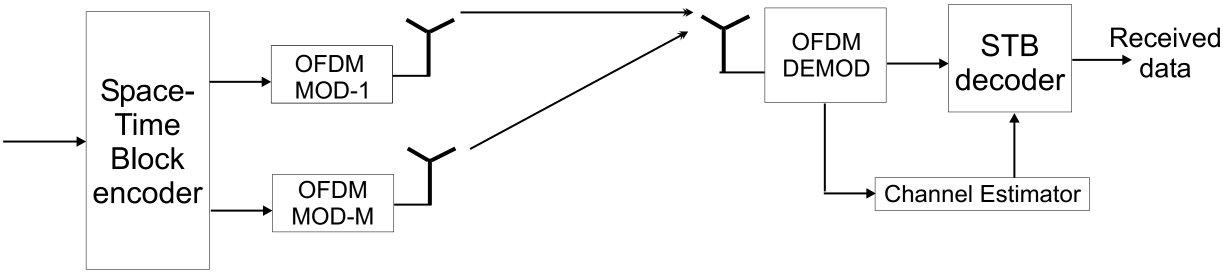

Figure 5.6: Space-Time OFDM System with STBC Algorithm

fading is constant (or time-invariant) across two symbol blocks (i.e. quasi-static channel), it can be written that

$$\begin{aligned} h_1(t) &= h_1(t+T) = h_1 = \alpha_1 e^{j\theta_1} \\ h_2(t) &= h_2(t+T) = h_2 = \alpha_2 e^{j\theta_2} \end{aligned} \tag{5.6}$$

In the (5.6), T is the symbol duration. Denoting the received signals as r_1 and r_2 at time t and $t+T$ respectively; similarly additive noise and interference as n_1 and n_2 respectively where n_1 and n_2 are complex random variables, the received signals can be expressed as

$$\begin{aligned} r_1 &= r_1(t) = h_1 s_1 + h_2 s_2 + n_1 \\ r_2 &= r_1(t+T) = -h_1 s_2^* + h_2 s_1^* + n_2 \end{aligned} \tag{5.7}$$

With the help of channel estimator, the combiner builds the following two combined signals that are sent to the maximum likelihood (ML) detector:

$$\begin{aligned} \tilde{s}_1 &= h_1^* r_1 + h_2 r_2^* = \left(\alpha_1^2 + \alpha_2^2\right) s_1 + h_1^* n_1 + h_1 n_2^* \\ \tilde{s}_2 &= h_2^* r_1 - h_1 r_2^* = \left(\alpha_1^2 + \alpha_2^2\right) s_2 - h_1 n_2^* + h_2^* n_1 \end{aligned} \tag{5.8}$$

These signals are then combined according to the ML principle as described in Figure 5.5 with the help of channel estimates.

It is observed that the diversity order in the 2Tx-1Rx STC is similar to that of 2-branch *Maximum Ratio Receiver Combining* (MRRC). [24] also prescribed the way to design 2Tx-MRx (2-branch transmit diversity with M receivers), to achieve a diversity order of $2M$ for the applications where a higher order of diversity is needed and multiple receive antennas at the remote units are feasible. The total radiated power from Tx antennas is kept proportional to $\frac{1}{M}$; in case of 2 Tx antenna, it is kept constant by transmitting half the power from each antenna. The simulations show a 3-dB loss of performance of the scheme compared to MRRC; this is caused by the fact that the transmitter does not know the channel, so no array gain is present in the system [20].

The role of OFDM modem in the STC is similar to OFDM based BLAST system. The OFDM modulation converts the frequency selective channel to frequency flat channels, so that STBC algorithm proposed can be applied to each flat fading subchannel over two consecutive OFDM symbols. High level block diagram of S-T OFDM system is shown in Figure 5.6. As in V-BLAST OFDM systems in Figure 5.4, source data is S-T block coded and later OFDM modulated over two consecutive OFDM symbols. These two symbols are sent to the receiver via the transmitter antennas. In the receiver, the signals are jointly received and OFDM demodulated, before they are separated and source data is recovered according to algorithm described in Figure 5.5.

5.2.3 Space-Frequency Coding

Space-Frequency Coding (SFC) is actually implementation of Alamouti scheme [24] on OFDM subcarrier basis. In this case, adjacent subcarriers are coded in a similar fashion as it is done in STBC between two complete OFDM symbols. SFC is very interesting for severely frequency-selective scenario. This topic will further be studied and included in the later versions of this report.

Chapter 6

Synchronization Issues

OFDM signal in time domain looks like a Gaussian noise because it is the superposition of many subcarriers with random phases and amplitudes depending on user data. There are hundreds or thousands of samples per OFDM symbol, since the number of samples necessary is proportional to the number of subcarriers [13]. The subcarriers are spaced closer and closer together in frequency domain when the number of subcarriers is increased to provide better data rate considering that the available bandwidth is the same. This brings the need for stricter synchronization in the system. Thus, synchronization is one of the most critical topics since OFDM demodulation is virtually impossible with minute synchronization error.

There are three aspects of OFDM synchronization, namely subcarrier frequency synchronization, receiver sampling frequency synchronization and symbol timing synchronization. The first one can be termed as frequency error and the last two can be grouped as timing error.

Subcarrier frequency offset error between the transmitter and the receiver can cause havoc in OFDM reception. It can happen for two reasons, first RF oscillator frequency mismatch between the transmitter and the receiver and the channel Doppler shift. The frequency offset causes two problems, one is the reduction of signal amplitude and the other is introduction of ICI [36]. The carrier frequency offset must be less than 2% of inter-carrier distance [37]. This restriction is obtained for normal Gaussian and typical Rician channels.

Symbol timing relates to the problem of detecting the start of a symbol. The requirement for this is somewhat relaxed when CP is used to extend the symbol [13]. The symbol timing result defines the DFT window; i.e. the set of samples used to calculate DFT of each received OFDM symbol.

A quite different timing estimation problem is tracking the sampling clock frequency. The oscillators used in DAC and ADC in transmitter and receiver respectively will never have exactly the same period. Thus the sampling instants slowly shift relative to each other [25]. This will result in incorrect sampling and thus increase the BER.

6.1 Symbol Timing Synchronization

Different OFDM systems have different requirements for symbol timing; for example, WLANs cannot spend more time beyond the preambles whereas a broadcast system can spend several symbols to acquire accurate symbol timing estimate. We will concentrate on WLAN case.

WLAN standards such as IEEE 802.11a and HiperLAN/2 specifies a preamble signal at the beginning of the transmission. The preambles of 802.11a standard are presented in Figure 6.2

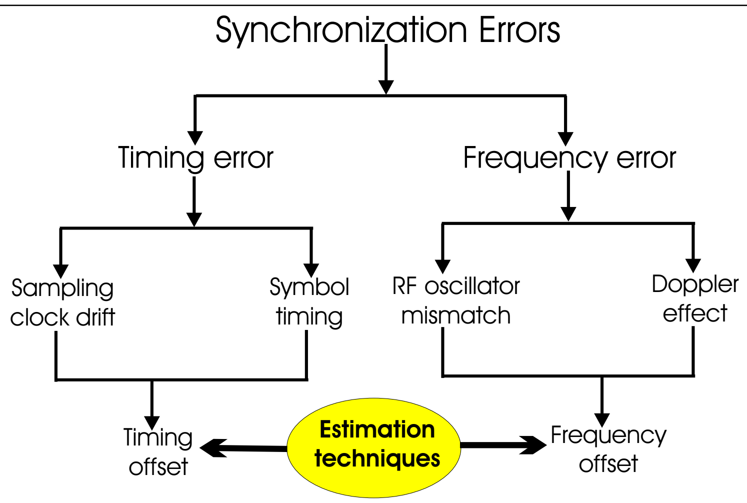

Figure 6.1: Synchronization Error in any OFDM System

[38]. In both the standards, the preambles are designed such that the start of the symbols can be easily determined at the beginning of the transmission. The first 10 parts starting from A_1 to A_{10} are all short training symbols; all of them are 16 samples long. The lasttwo parts C_1 and C_2 are long training symbols that span 64 samples as it is for a regular OFDM symbol. The middle part CP is 32 samples long and saves the long training symbols from multipath interferences.

The knowledge of the preamble is available to WLAN receiver, thus it can easily make use of a simple cross-correlation technique for symbol timing. After the packet detection algorithm signals the start of a packet, the symbol timing algorithm refines the estimation to sample level precision. This is done by using the cross-correlation between the received signal r_n and a known reference t_k; for example the end of the short training symbols or the start of the long training symbols. The value of n that corresponds to the maximum absolute value of the cross-correlation is the symbol timing estimate [25]. Mathematically it is shown in equation 6.1:

$$\hat{i}_s = arg \max_n \left(\left| \sum_{k=0}^{L-1} r_{n+k} t_k^* \right|^2 \right) \tag{6.1}$$

In a multipath environment the symbol timing estimation can be improved if an estimate of the multipath taps h_n of the channel impulse response is available [6]. In a multipath environment, the first arrived path is not necessarily the strongest path. Thus cross-correlation algorithm may take a high correlation value corresponding to paths that arrive later than first path. In that case, the signal energy at the first path is lost. If the knowledge of h_n is available, then the timing point can be changed such that the energy of h_n inside the DFT window can be maximized.

The method that is mentioned above is non-data aided method, where only CP is used to minimize the symbol timing error. There are data aided methods available also, which are used in practical OFDM systems. In data-aided methods, pilots are inserted in various locations of the signals, and symbol timing is derived based on the known pilots.

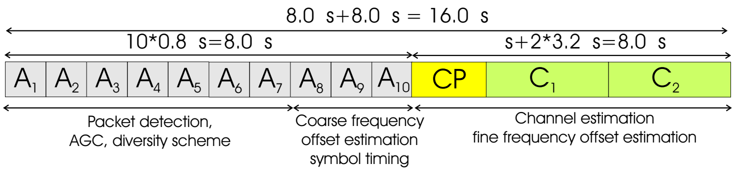

Figure 6.2: OFDM Preamble Structure Specified in IEEE 802.11a Standard

6.2 Sampling Clock Synchronization

Considering a relative sampling clock offset between the transmitter and the receiver of

$$\gamma = \frac{T^{'} - T}{T}, \tag{6.2}$$

where T and $T^{'}$ are the sampling times in transmitter and receiver respectively. With this, the overall effect in the frequency domain (i.e. after the DFT block) on the k^{th} received subcarriers of i^{th} OFDM symbol, $R_{i,k}$, is [39]:

$$R_{i,k} = exp\left(j2\pi k\gamma i\frac{T_{sym}}{T_u}\right) X_i(k) sinc(\pi k\gamma) H_{i,k} + N_\gamma(i,k) \tag{6.3}$$

Here, $H_{i,k}$ refers to channel channel frequency response of i^{th} OFDM symbol in k^{th} subcarrier and $N_\gamma(i,k)$ is the additional interference due to the sampling frequency offset. The power of this term is approximated by

$$P_\gamma \approx \frac{\pi^2}{3}(k\gamma)^2 \tag{6.4}$$

(6.4) shows that the degradation increases with the square of the amount of the offset and it is the highest for outer most subcarriers [25].

Using pilot subcarriers for estimating the sampling frequency error is one of the techniques that are mostly found in the literature. The sampling frequency offset is estimated by using the knowledge of the linear relationship between the phase rotation caused by offset and the pilot subcarrier index [25].

A method that is presented in [40] introduces two sets of pilot subcarriers; p_1 corresponds to pilots on negative subcarriers and p_2 corresponds to pilots on positive subcarriers. The received pilots can be represented by:

$$R_{i,k} = H_k p_{i,k} exp\left(j2\pi k\gamma i\frac{T_{sym}}{T_u}\right) \tag{6.5}$$

Now, if we calculate the rotation between two consecutive pilots, we find that

$$\begin{aligned}\Delta_{i,k} &= R_{i,k}R^*_{i-1,k} = H_k p_{i,k} e^{j2\pi k\gamma i\frac{T_{sym}}{T_u}} \left(H_k p_{i-1,k} e^{j2\pi k\gamma(i-1)\frac{T_{sym}}{T_u}}\right)^* \\ &= H_k p_{i,k} e^{j2\pi k\gamma i\frac{T_{sym}}{T_u}} H^*_k p^*_{i-1,k} e^{-j2\pi k\gamma(i-1)\frac{T_{sym}}{T_u}} \\ &= |H_k|^2 |p_{i,k}|^2 e^{-j2\pi k\gamma\frac{T_{sym}}{T_u}} \end{aligned} \tag{6.6}$$

The cumulative phases of $\Delta_{i,k}$ is calculated next for the sets of pilots p_1 and p_2, which is explained in (6.7):

$$\varphi_{1,i} = \angle\left[\sum_{k\in p_1} \Delta_{i,k}\right] \quad \& \quad \varphi_{2,i} = \angle\left[\sum_{k\in p_2} \Delta_{i,k}\right] \tag{6.7}$$

Now the sampling frequency offset can be estimated from the equation above:

$$\hat{\gamma} = \frac{1}{2\pi}\frac{T_{sym}}{T_u}\frac{1}{\min_{k\in p_2}(k) + \max_{k\in p_2}(k)}\left(\varphi_{2,i} - \varphi_{1,i}\right) \tag{6.8}$$

6.3 Carrier Frequency Synchronization

The Local Oscillators at the Transmitter and the Receiver may not be at the same frequency. Since they are analogue high frequency components they cannot be at the same frequency. There will almost always be some error say of the order of a few ppm be there. This is referred to as carrier frequency offset. This error results in monotonically increasing or decreasing phase error with time. The accumulated phase error will rotate of the symbol constellation in the receiver causing the error in the decision during the demodulation of the symbol. This error goes on accumulating with time. It is very difficult to correct the oscillator frequency, instead most of the implementations allow the LO to run at its frequency but estimate the frequency offset and apply compensation accordingly to nullify the phase accumulation. IEEE 802.11a allows the use of short training sequence to perform the coarse frequency estimation [14]. Further there are long training sequence in IEEE 802.11a for fine frequency synchronization.

Various algorithms that have been developed to compensate carrier frequency offset can be classified under three different groups; namely *Data Aided algorithms* (DA) which are based on special training symbols embedded in the transmission signal, *Non-Data Aided algorithms* (NDA) that analyzes the signal in frequency domain and CP based algorithms that use the inherent structure of the OFDM signal provided by the CP [41]. Out of these three, DA methods are most suited for WLANs, as the WLAN standards specify preambles that contain training symbols and can easily be used for the synchronization purpose.

There are principally two different approaches for this, one is time domain approach which is performed in received signal before it goes through the DFT block and the other one is frequency domain approach which is performed after the DFT operation. A very good description of these two approaches is placed in [25]. Here we present the time domain approach in detail.

Let the transmitted signal be S_n, then the complex base band model of the passband signal y_n is

$$y_n = S_n e^{j2\pi f_{tx} n T_{sym}} \tag{6.9}$$

where f_{tx} is the transmitter carrier frequency. After down conversion at the receiver using carrier frequency f_{rx} the received complex baseband signal r_n, ignoring the impairments due to noise and channel is

$$r_n = x_n e^{j2\pi f_\delta n T_{sym}} \tag{6.10}$$

where

$$f_\delta = f_{tx} - f_{rx} \tag{6.11}$$

The short training sequence is periodic after say λ. Then the estimation method for frequency offset is

$$c = \sum_{l=0}^{\lambda-1} r_l r_{l+\lambda}^* \tag{6.12}$$

after simplification of (6.12) we get

$$c = e^{-j2\pi f_\delta \lambda T_{sym}} \sum_{l=0}^{\lambda-1} |x_l|^2 \tag{6.13}$$

Thus we get the estimate of the carrier frequency offset as

$$f_\delta = -1/(2\pi\lambda T_{sym})\angle c. \tag{6.14}$$

This is a common time domain estimation of coarse frequency offset. The long training sequence of IEEE 802.11a is used for fine frequency estimation. The method followed is similar except that the the value of λ changes.

Chapter 7

Channel Estimation

Current OFDM based WLAN standards such as IEEE802.11a and HiperLAN2 use variations of Quadrature Amplitude Modulation (QAM) schemes for symbol mapping which require a coherent detection method in the receiver. And naturally, data detection in coherent OFDM receivers require an accurate (or near accurate) estimate of Channel State Information (CSI). In non-coherent methods, the detection is performed based on the differential information available between successive symbols.

There are two major kinds of channel estimators that are found in literature, namely pilot assisted and blind estimation. A mixture of these two, where a blind method with limited training symbols is used, is called semi-blind technique.

Traditional one-dimensional channel estimation techniques for the OFDM systems can be summarized as follows: *Least Squares* (LS), *Minimum Mean Squared Error* (MMSE) and *linear MMSE* (LMMSE). LS estimators are very simple to constitute, but they suffer from MSE in low SNR conditions. MMSE, based on time domain estimations, are high complexity estimators that provide good performance in sampled-spaced channels, but limited performance in non-sample-spaced channels and high SNR conditions. The third one, LMMSE provides good performance in both sampled and non-sampled channels [42].

7.1 Exploiting Channel Correlation Properties for CSI Estimation

In OFDM systems, the Doppler effects are kept smaller by making sure that the symbol duration is much smaller compared to the channel coherence time. In this case, the channel attenuations at successive symbol durations experience sufficiently higher time correlation. Similarly, if subcarrier spacing is chosen in a way that the spacing is much smaller than the coherence bandwidth of the channel, the channel attenuations at the adjacent subcarriers will be highly frequency correlated. So, the estimator can exploit both of these two correlation properties [43].

Figure 7.1 emphasizes the role of channel estimation in coherent detection of a WLAN OFDM receiver. Channel estimation is performed after the FFT processing, and prior to data detection. In this case, the estimated CSI is only used for equalization. Since each subcarrier is flat fading, all the techniques suitable for single carrier flat fading systems are directly applicable to OFDM subcarrier level.

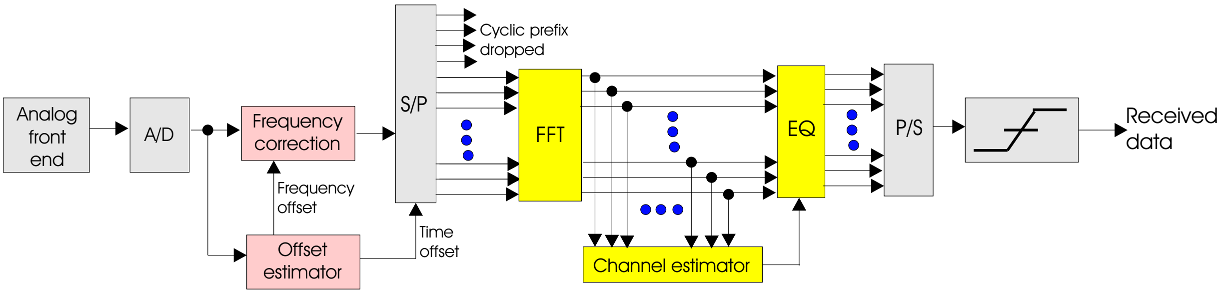

Figure 7.1: OFDM Receiver with Coherent Detection (using Channel Estimation)

Channel estimation of a SISO-OFDM[1] system can be done by using complete training symbols after certain OFDM data symbols, or by inserting some training pilot tones in every OFDM symbol. In the first case, the CSI is estimated with the training symbol and interpolated for the consecutive symbol before the next training symbol appears. This technique renders unacceptable results when the channel variation time is comparable to OFDM symbol duration. The second method is suitable in these kinds of fast varying channels. The CSI is estimated for all the pilot tones using the pilot subcarriers from that particular symbol and later CSI for all other subcarriers are obtained by interpolation. In that way, perfect or near perfect estimates are achievable. But the cost is paid in significant throughput reduction.

When channel is considered to be quasi-static, (i.e. the channel does not change much between two consecutive OFDM symbols, in another words the channel coherence time is higher than OFDM symbol duration), then the temporal correlation between two consecutive OFDM data symbols can be used to improve the estimates. If the channel estimates for k^{th} subcarrier at l^{th} time instance is denoted by $\hat{H}(k,l)$ and $\hat{H}(k,l+1)$ is the estimate for next time instance, that is $(l+1)^{th}$ time instance for same k^{th} subcarrier, then the best MSE estimator for $\hat{H}(k,l+1)$ given $\hat{H}(k,l)$ and $H(k,l+1)$ is [44]

$$\hat{H}(k,l+1) = a\hat{H}(k,l) + bH(k,l+1) \tag{7.1a}$$

$$\text{where} \quad a = \frac{R_{hh}^2(l) - R_{hh}^2(l+1)}{R_{hh}^2(l) - R_{hh}^2(l+1) + \sigma^2 R_{hh}^2(l)} \tag{7.1b}$$

$$\text{and} \quad b = \frac{\sigma^2 R_{hh}^2(l+1)}{R_{hh}^2(l) - R_{hh}^2(l+1) + \sigma^2 R_{hh}^2(l+1)} \tag{7.1c}$$

In fact, using (7.1a), the temporal correlation properties are used for better channel estimations. In a similar way, the channel frequency correlations can also be used between subcarriers of an OFDM symbol.

7.2 Channel Estimation Based on Pilots

With an OFDM system, the wideband channel is sliced to a number of narrow band channels, which are tagged to the respective subcarrier frequencies. Thus the trivial task of channel equalization is reduced to simply estimating the channel transfer function, which are the narrow bands. Such a channel can be estimated by inserting pilot symbols with known modulation

[1]SISO refers to Single Input Single Output system, where the channel is modeled with one transmitter and one receiver. SISO can be regarded as the counterpart of MIMO.

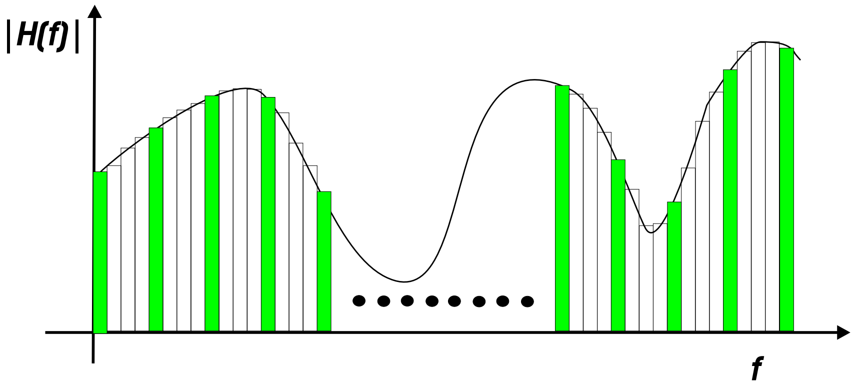

Figure 7.2: Channel Estimation with Pilot Symbols

scheme into the transmitted signal. Based on these pilot symbols, the receiver can measure the channel transfer factors for each subcarrier using interpolation techniques [10].

Figure 7.2 illustrates pilot tone based channel estimation technique for OFDM systems. The channel is estimated with the pilot symbols that are inserted (in dark color in the figure). The attenuations of the pilot symbols are measured and the attenuations of the data symbols between these pilot symbols are typically estimated/interpolated using time correlation property of fading channel [43].

A simple mathematical description of optimum pilot-assisted channel estimation is presented in [43]. In a matrix form, the observed symbols after the DFT operation in the receiver can be written as

$$\bar{r} = \bar{X}\bar{h} + \bar{n} \tag{7.2}$$

where the diagonal matrix $\bar{X}$ contains the transmitted symbols on its diagonal (either known pilot symbols receiver decisions of information symbols which are assumed to be correct), the channel attenuations of one OFDM symbol (i.e. Fourier transform of $h(t)$ evaluated at the frequency f_k) is collected in vector $\bar{h}$ and the vector $\bar{r}$ contains the observed outputs of the DFT.

If we maximize the channel estimates in the *Least-Square* (LS) sense (minimizing $\left|\left|\bar{r} - \bar{X}\hat{\bar{h}}\right|\right|^2$ for all possible $\hat{\bar{h}}$), then

$$\hat{\bar{h}}_{ls} = \bar{X}^{-1}\bar{r} = \begin{bmatrix} \frac{r_0}{X_0} & \frac{r_1}{X_1} & \cdots & \frac{r_{N-1}}{X_{N-1}} \end{bmatrix}^T \tag{7.3}$$

This is a straight forward estimation technique where the received symbol on each subcarrier is divided by the transmitted symbol to obtain the estimate. In the next step, the frequency correlation can be used to smooth and improve the LS channel estimate.

The optimal *Linear Minimum Mean-Square Error* (LMMSE) estimate of $\hat{\bar{h}}$ (minimizing $E\left(\left|\left|\hat{\bar{h}} - \bar{h}\right|\right|^2\right)$ for all possible linear estimators $\hat{\bar{h}}$) becomes

$$\hat{\bar{h}}_{lmmse} = \bar{A}\hat{\bar{h}}_{ls} \tag{7.4}$$

where $\bar{A} = \bar{R}_{hhls}\bar{R}_{hlshls}^{-1} = \bar{R}_{hh}\left(\bar{R}_{hh} + \sigma_n^2\left(\bar{X}\bar{X}^H\right)^{-1}\right)^{-1}$ and $\bar{R}_{hh} = E\left(\bar{h}\bar{h}^H\right)$ is the channel autocorrelation matrix, that is the matrix that contains the correlations of the channel attenuations

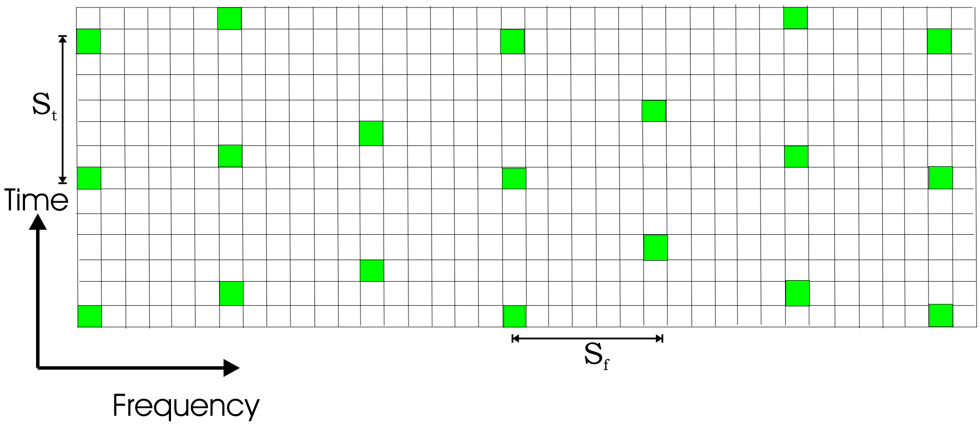

Figure 7.3: An Example of Pilot Symbol Insertion Method

of the subcarriers. Similarly $\bar{R}_{hhls}$ denotes the correlation matrix between channel attenuations and their LS-estimates, and $\bar{R}_{hlshls}$ denotes the autocorrelation matrix of the LS estimates.

The solution that is described above is computationally very extensive, because we need N number of complex multiplications per estimated attenuation and also we need the knowledge of SNR. Thus, practically LMMSE solution is not feasible to implement, but it is observed that an optimal solution can be obtained starting from this solution. [45] has developed generic complexity approximations of equation 7.4 whose performance can be made very close to LMMSE solution.

7.2.1 Design of Pilot Based Channel Estimator

There are mainly two problems in designing channel estimators for wireless OFDM systems. The first problem concerns the choice of how pilots should be inserted. The second problem is the design of the estimator as a low complexity with good channel tracking ability.

The pilot symbols should be inserted properly, so that it successfully estimates the frequency response of the channel. The difference between two consecutive pilot symbols in time and frequency domain, S_t and S_f respectively, can be represented as

$$S_t \leq \frac{1}{B_{doppler}} \quad \text{and} \quad S_f \leq \frac{1}{\tau_{max}} \tag{7.5}$$

Here $B_{doppler}$ is the Doppler spread. Figure 7.3 shows an example of such a pilot insertion scheme [6].

Assuming that the pilot pattern is chosen, the optimal linear channel estimator in terms of *Mean-Square-Error* (MSE) is a 2-D Weiner filter. Knowing statistical properties of the channel, such a channel can be designed using standard techniques [16].

7.3 Channel Estimation Based on Training Symbols

Conventional estimation schemes send a stream of transmitted symbols with a modulation scheme known to the receiver, and the receiver analyzes the effect of the channel on the known symbols by observing the deviations on the received known symbols. These symbols are called Training

Symbols (TS). As this estimation technique solely depends on the TS, this type of channel estimations are usually termed as TS-based channel estimation. The accuracy of the CSI is dependent on the length of TS. It is worth mentioning here that the transmission of training symbols reduces the spectral efficiency of the system [6].

In WLAN systems, the training sequences as shown in Figure 6.2 are provided to facilitate straightforward and reliable channel estimation. The long training sequence provided in the WLAN IEEE 802.11a standard is to aid in obtaining the channel estimation. Since there are two long training sequences, it can be averaged to improve the quality of the channel response. Since the DFT is a linear operation, the average of the two sequences can be performed before DFT, saving the computational resources. After the DFT processing, the received training symbols $R_{1,k}$ and $R_{2,k}$ are a product of the training symbol L_k and the channel Hk plus the additive noise $W_{l,k}$,

$$R_{1,k} = H_k L_k + W_{l,k} \tag{7.6}$$

Since the long training sequence L_k is known, we can compute the Channel Frequency response as following:

$$\hat{H}_k = \frac{1}{2}\left(R_{1,k} + R_{2,k}\right) L_k^* \tag{7.7}$$

$$\hat{H}_k = \frac{1}{2}\left(H_k L_k + W_{1,k} + H_k L_k + W_{2,k}\right) L_k^* \tag{7.8}$$

$$\begin{aligned}\hat{H}_k &= H_k |L_k|^2 + \frac{1}{2}\left(W_{1,k} + W_{2,k}\right) L_k^* \\ &= H_k + \frac{1}{2}\left(W_{1,k} + W_{2,k}\right) L_k^* \end{aligned} \tag{7.9}$$

It is assumed here that the long training sequence has a magnitude of unity and the noise samples $W_{l,k}$ and $W_{2,k}$ are statistically independent so there variance is half the variance of the individual noise sample thus enhancing the SNR by 3 dB [14].

7.4 Blind Channel Estimation

Contrary to pilot based and training symbol based channel estimation techniques, blind channel estimation requires no training symbol, instead certain known properties of the transmitted symbols are observed to obtain a perfect or near perfect CSI. In this way, the spectral efficiency in increased, but the cost is paid in terms of increased computational cost. Furthermore, blind algorithms does not converge very fast, thus limited amount of training symbols are used to converge the computation faster, in which case the algorithms are termed as semi-blind algorithms.

There are some good numbers of blind algorithms thar are studied for OFDM systems. One of the mostly used algorithm is well-known *Constant Modulus Algorithm*CMA [46]. This topic will further be studied in near future and some discussions will be added with future versions of this report.

7.5 Channel Estimation in CDD-OFDM System

We have discussed CDD-OFDM system in Section 5.1.2 and Section 5.1.3. As shown in Figure 5.3, the channels are estimated immediately after the reception at the receiver front end in case of Receiver CDD-OFDM system. In this section, we consider dual antenna receiver diversity scheme with CDD. We denote the estimated value of CTFs $\mathbf{h}_1$ and $\mathbf{h}_2$ as $\tilde{\mathbf{h}}_1$ and $\tilde{\mathbf{h}}_2$ respectively. The estimated combined channel frequency response as a function of n and g can be written as

$$\tilde{\mathbf{h}} = \sqrt{g}\tilde{\mathbf{h}}_1 + \sqrt{1-g}e^{j\psi}\mathbf{W}_N^n\tilde{\mathbf{h}}_2, \tag{7.10}$$

where the phase of the weight factors of 1^{st} and 2^{nd} branch signals are given by 0 and ψ, respectively. g denotes the complex gain factors for combiner branches and n denotes the cyclic delay (please refer to Section 5.1.3 for details).

With a 16-QAM or higher order constellations, it is necessary to know the CSI at the receiver, i.e. the phase and gain of the CTF should be estimated correctly for correct data detection. Because we are using narrow-band tones within multi-carrier system, we only need to estimate a single value [47]. We assume that the channel is wide sense stationary in a sense that it is constant in time for at least one complete OFDM symbol duration. Given a Minimum Mean Square-Error(MMSE) channel estimator designed to cope with worst-case channel conditions, the channel estimates for l^{th} received OFDM symbol for i^{th} diversity branch is well modelled as the true CTF sample $\mathbf{h}_{l,i}$ disturbed by AWGN $\mathbf{n}_{l,i}^H$ [48]

$$\tilde{\mathbf{h}}_{l,i} = \mathbf{h}_{l,i} + \mathbf{n}_{l,i}^H, \tag{7.11}$$

where the power $\sigma_h^2(\sigma_n^2)$ of the estimation noise $\mathbf{n}_{l,i}^H$ is a function of the channel noise variance. A simple one-tap Recursive Least Squares(RLS) filter, which first estimates the channel and then normalizes the channel effect, can be used [49]. Mathematically,

$$\tilde{\mathrm{h}}_{k,l}^i = \alpha\frac{r_{k,l}^i}{\tilde{s}_{k,l}} + (1-\alpha)\tilde{\mathrm{h}}_{k,l-1}^i \tag{7.12a}$$

$$\mathrm{z_{k,l}} = \frac{r_{k,l}^1\tilde{\mathrm{h}}_{k,l-1}^{1*} + r_{k,l}^2\tilde{\mathrm{h}}_{k,l-1}^{2*}}{|\sqrt{g}\tilde{\mathrm{h}}_{k,l-1}^1|^2 + |\sqrt{1-g}\tilde{\mathrm{h}}_{k,l-1}^2 W_N^k|^2} \tag{7.12b}$$

$$\tilde{s}_{k,l} = Quantize(z_{k,l}), \tag{7.12c}$$

where $\tilde{s}_{k,l}$, $\tilde{h}_{k,l}$ and $z_{k,l}$ are estimate of constellation points, estimate of demodulated channel and decision variable (a noisy estimate of constellation value $s_{k,l}$) for k^{th} subcarrier and l^{th} OFDM symbol respectively, α is the update factor in the estimate of the channel. $\tilde{s}_{k,l}$ is either a training symbol or the quantized decision variable. Assuming no decision errors, the error in the estimate of the subcarrier $h_{k,l}$ is then [47]:

$$\sigma_{\mathbf{n}_{l,i}^H}^2 = \sigma_{N_S}^2\frac{\alpha}{2-\alpha}E[\frac{1}{|x|^2}], \tag{7.13}$$

where N_S is the noise on the received signal.

7.6 Channel Estimation in MIMO Enhanced OFDM Systems

In MIMO-OFDM system, the number of channels that need to be estimated is $[M, N]$, where M represents the number of transmit antennas and N represents the number of receiver antennas. Naturally the complexity with the estimation increases with increasing number of antennas at any side of the channel. Practically all signals are transmitted simultaneously from the all M transmit antennas, thus the received signal at any one of the N receivers is a superposition of the transmitted signals of all the transmitter antennas at that instance of time. Thus, the system should be modelled as follows: for the channel between i^{th} transmit antenna and j^{th} receive antenna, only the signal from i^{th} transmit antenna is the desired signal and the rest are interfering signals. To avoid this problem, some schemes suggest that the pilot tone that is used by one antenna is not at all used by other antennas [50]. So interferences in pilot tones can be avoided and correct CSI can be estimated. This method takes note that the time duration between two consecutive OFDM symbols are higher than the channel coherence time, or no channel temporal correlation information is available.

Chapter 8

Research Challenges

Before discussing the research challenges, let us consider the user requirements and the bottlenecks to realize such requirements. User requirements, such as higher link capacity, better QoS and ubiquitous access (access to internet from anywhere and with any type of device), are always present for discussion in any wireless system design process. There is also a requirement of robust communication over wireless channel even at low data rates. The bottlenecks to these requirements being primarily the channel itself, the stability, correctness and transient response of radio frequency components, the fixed word length of the hardware and limitation of battery life for portable devices.

8.1 Wireless Channel Modelling

Wireless channel has always been a challenge to researchers and scientists. The hostility of wireless channel can never be eradicated, thus it is always the goal of system designers to overcome the effects of channel on the received signal. In this respect, there are several parameters that characterize a wireless multi-antenna communication system, such as antenna separation, propagation terrain and/or foliage, *Base Transceiver Station* (BTS) antenna height, polarization, CCI, range, wind speed /traffic and mobile antenna height, etc, need to be studied and examined. Several well-known channel models are found in the literature, but each addresses only some of the parameters of wireless propagation, which are specific to the applications, to name a few, the SISO models, the path loss model,average delay spread model, the K-factor model, the cluster model approach, the UTRA model, the MIMO models, the ray tracing model, the METRA model and the scattering model. There is definite need for a unified all encompassing channel model that can be used in the development of MBWA. Such a model will be able to cover the situation of fixed wireless access (indoor office model), low mobility situation (Pavement traffic) and high vehicular mobility.

8.2 Synchronization Issues

In Chapter 6, the details of the effect of synchronization issues have been dealt with. In WLAN systems, the carrier frequency synchronization and symbol synchronization are done using the preamble transmitted as the header of the packet. Some algorithms use the CP data for symbol synchronization as well. In the case of broadcast systems a lot of pilot sub-carriers have to be sent for synchronization purpose. So there is lot of forced wastage of spectrum.

Also if there are some estimation errors in carrier frequency or symbol timing or sampling clock, all the errors translate as rotation of the constellation points. So the success of OFDM system depends strongly on synchronization. The higher the data rates are, the stricter the synchronization requirements become. So in order to build systems to support higher and higher data rates, there is a need for a range of algorithms and system design that facilitates robust stimulation of the synchronization parameters with minimum computational complexity. Our literature survey gave us the idea that the currently available algorithms are limited in their performance in case of doppler shift and larger delay spread environment. Such systems with one time estimate (non-real time estimate) of the synchronization parameters will not fit the mobile environment. New and efficient techniques are required to enable improved tracking of environmental parameters that hinder the system.

Sampling clock synchronization is another parameter that effect the OFDM system. A drift in the sampling clock manifests itself in the form of ICI. There are methods to estimate the sampling clock drift for WLAN packet based systems, but these are either too complex to implement or are not that robust in their performance. When the packet length is big, with very low SNR and very high delay spread along with doppler shift, the accuracy and stability of these algorithms comes to question. Another problem often encountered due to mismatch in sampling clock frequencies is, the missing of a sample or over sampling of the incoming data. If this condition is not traced and compensated, then the system performance will fall drastically in case of packet based WLAN systems. A suitable system design to enable continuous tracking of sampling clock can amazingly improve the overall error rate of the system.

8.3 Channel Estimation Issues

We have discussed some issues related to channel estimation in Chapter 7. Channel estimation is a primary requirement of any receiver that perform coherent reception. The capacity of a system is largely depended on the channel estimation scheme used in the system. The more accurate the channel estimate is, the better the quality of service. OFDM offers a very simple frequency domain channel estimation Scheme. Even though the scheme is simple enough it does not perform accurately under very low SNR conditions. There are different techniques in Single Carrier systems, such as *Linear Equalizer* (LE), *Zero Forcing Equalizer* (ZFE), *Decision Feedback Equalizer* (ZFE) etc, but these are all time domain equalizers. Efficient yet robust techniques can be thought of that use both frequency domain and time domain processing. There could be possibilities of implementing Dynamic Cyclic Prefix Length in order to dynamically increase the data rate. The wireless channel though is a fading one, yet there is some coherence bandwidth and the fading characteristics are not necessarily random, this fact can be exploited for implementing better systems. One may optimize the performance by dynamic pilot allocation as well. There can be optimization of TDD systems where the down link and up link data path can share the channel information and thus can improve the throughput by pre-equalization.

The case of mobile environment throws up quite a few alternatives. Obviously, the channel transfer function changes during the packet interval. Thus the estimation of the channel by preamble may not be valid even after every symbol. A constant tracking mechanism is needed. The DVB-T system uses many pilots in the symbol for this purpose, so a mix of pilot based tracking plus preamble based initial estimate can be thought of as an alternative.

MIMO systems are being thought upon for implementation in MBWA IEEE 802.20. With MIMO systems, new ways should be found for channel estimations. MIMO systems bring with them the additional overload of multiple channel estimations, so it will be required that the computation cost of such channel estimations need to be reduced. To conclude, the focus of the channel estimation techniques should be low complexity and power efficient solutions.

8.4 Capacity Enhancement via MIMO

MIMO algorithms are usually derived from a narrow-band point of view. As the wideband OFDM modulation converts the subcarriers to a narrowband channel, it should be possible to implement all kinds of MIMO algorithms on subcarrier by subcarrier basis. MIMO channel modelling promises an increase in data throughput of overall transmission channel, so using OFDM with an amalgamation of MIMO concepts will definitely play a significant role in future WLAN systems. MIMO techniques should be studied more extensively to find efficient ways for increasing the data rate of the current OFDM systems. For example, SM opens multiple data pipes between transmitter and receiver. Now when we increase the number of transmitter and receiver antennas, we effectively increase the possibility of reaching higher data rate. But this comes with some side effects, with increasing potentials for super data rate, the processing loads also get increased which is not a pleasant situation. So, it is required that techniques should be found to introduce MIMO advantages in an OFDM system in more efficient and reasonable way. STFC techniques for MIMO systems is another important issue that can be looked into.

8.5 System Implementation

The role of OFDM as the possible physical layer contender for high data rate future wireless communications is somewhat established by WLANs adoption of it as their PHY standard. Understanding the demand for OFDM based system, there is a need to design OFDM based systems for future wireless applications. OFDM involves complex system operations, which must coincide with numerous constrains, such as timing, synchronization, RF impairments etc. Thus, OFDM deserves a thorough study for better and more efficient designs. Numerous theoretical researches are done on WLANs. Nevertheless, hardware implementations have not yet achieved the maximum limit of the transmission data rate specified by the standards. So, developing OFDM hardware for future systems is an immense need.

On the other hand, tremendous growth of reconfigurable hardware in recent years has made it possible to implement complex telecommunications related digital signal processing algorithms. There are several of FPGAs[1] available now with multimillion gates featured on them. FPGAs have grown in their logic densities and have shrunk in the logic area tremendously. Recent FPGAs are made of millions of logic gates with special multipliers banks, where highly computational signal processing, such as OFDM, can be implemented. The recent FPGAs are complete and concise in supporting OFDM operation. Moreover there are user-friendly tools available that can be used to program the FPGAs in very short time. So, it is an immense requirement to design an OFDM based Multicarrier Modulation (MCM) testbed in FPGA, where several different systems can be tested in real life scenario.

Huge efforts are required on designing power efficient implementation of such systems, since most of the systems require large amount of computations. May be a more power efficient system with moderate data rate is the need of the day than a system that consumes a lot of energy to

[1] Field Programmable Gate Arrays

give very high data rates. The field is totaly open and one may go in any direction just carrying only one objective to develop better but usable communication system.

8.6 Peak to Average Power Reduction

The peak to average power issue of OFDM systems was discussed in Section 3.5.2. Lot of effort have been made to reduce the PAPR of OFDM systems. Let us first discuss the steps that have already been taken and the performance achieved by them so that we can find out the area or the path that may be tried out for further improving the situation.

There are a range of parameters, all trying to quantify the variation of the amplitude of the OFDM symbol. Finally the factor that was accepted and used in most of the papers is $PF_2 = \frac{max(Power)}{rmsPower}$ [51]. Mostly non-linear amplitude clipping algorithm was used to reduce the PAPR of OFDM systems. Amplitude clipping resulted in out-of-Band spectrum spillage. Also the BER and PER performance was degraded by the implementation of the amplitude clipping algorithms [6]. There are several clipping techniques discussed in [6] where each of the techniques tries to minimize the out-of-band radiation, the BER and PER.

An alternative direction is to use coding of the incoming bit-stream such that the resulting IFFT output had a lower PAPR [6, 51]. *Shapiro-Rudin* sequences were the early ones. These codes assumed that the bit stream had to be BPSK modulated, i.e. the trigonometric series would have coefficients of either +/-1 only. Thus multilevel coding situation was not considered, even though it achieved great performance for BPSK. It established an upper and lower bound for the PAPR of coded OFDM. It was later shown that Shpairo-Rudin codes were a special case of Golay complementary codes. For 16 carrier systems theoretical PAPR is about 12 dB , but by using Shapiro-Rudin codes it can be brought down to 2.8 dB. Shephard [52] proposed the use of Shapiro-Rudin codes for $\frac{1}{2}$-rate QPSK-OFDM. Golay codes were also proposed. *Gloay* codes, found in 1961 are a set of complementary codes. Popović showed that Golay codes provide the benefit that the PAPR is bounded by 2 (or 3) dB . Popović also showed that Shapiro-Rudin codes were a special case of Golay codes. It has been found that with Golay codes coding rate falls much lower than $\frac{1}{2}$, when number of carriers exceed 16. So it cannot be efficiently implemented in case of high number of sub-carriers. Another class of codes called *M-sequence* (Maximum Length) codes and partial M-sequence codes. Selected mapping was another technique that was tried. Selected mapping technique had low overhead , but it was computationally intensive since one had to find the best mapping out of the set of randomly mapped codes.

All these codes mentioned above are designed for BPSK modulation systems. Not much literature is found on techniques to reduce the PAPR for QPSK, QAM-16, QAM-64 and higher order modulated OFDM systems with very large number of sub-carriers. So with the insight of the already tried and proven methods for BPSK we can investigate and deisgn new coding shcemes that may have the capability of reducing the PAPR for high data rate OFDM systems with higher order modulation and very large number of sub-carriers. Combined codes can be designed so that the Error Correcting Codes and PAPR reducing codes may not be seen separately, so as to optimize the transmission/encoding and reception/decoding procedure.

8.7 Dynamic CP Length

The current standards are "too" robust with the channel impairments. The channel is strictly taken as stationary indoor channel. By using the term stationary it is meant that between few

OFDM symbols, the channel is expected to become static, or not changing. Thus, it may be wise to "measure" the channel maximum delays and insert the CP according to the measurement. In that case, a technique has to be found to measure the maximum channel delays. This can be done using the pilots that are inserted for channel estimation and synchronization process.

CP plays an important role in combating multipath effects by reducing ICI to maintain orthogonality between subcarriers and eliminating the ISI. It is also the fact that inserting CP has its own cost, a part of signal energy is lost since CP carries no information. The loss is measured as [6]

$$SNR_{loss_CP} = -10log_{10}\left(1 - \frac{T_{CP}}{T_{symbol}}\right) \tag{8.1}$$

According to IEEE 802.11a standard, CP length is $0.8\mu s$ and data part of OFDM symbol is $3.2\mu s$, thus complete OFDM symbol duration is $4.0\mu s$, which is 5 times the CP length. Using the above equation, we can see that 0.97dB in SNR is lost to accommodate such amount of CP. The standard specifies $0.8\mu s$ as the CP duration considering the fact that the maximum RMS delay spread at worse case condition can be up to $200ns$, and the CP is taken as 4 times that value. But in practice, the usual delay spread values in static indoor channels are $40 - 70ns$ [53]. Thus it is clear that to cater the worst case scenario, actually a large part of signal energy is always being wasted.

Now we recount the loss of SNR due to CP insertion considering that we have a mechanism in place that can measure the RMS delay spread dynamically. For example, we assume that in average the maximum delay spread is somewhere in between $50ns$ to $100ns$, then if we take $200ns$ as the CP length, then $T_{CP} = 0.2\mu s$ and $T_{symbol} = 3.4\mu s$. With these new values, $SNR_{loss_CP} = -10log_{10}\left(1 - \frac{0.2}{3.4}\right) = 0.26dB$, so in this case by reducing the symbol length with smaller CP length, $(0.97 - 0.26)dB = 0.71dB$ of SNR can be saved.

In another way, if we maintain the OFDM symbol length so that $T_{CP} = 0.2\mu s$ and $T_{symbol} = 4.0\mu s$, then we should be able to enhance the user data rate. New maximum data rate using best case modulation (64-QAM) and channel coding according to the standard will be $\left(1 + \frac{0.8-0.2}{4.0}\right) * 54Mbps = 62.1Mbps$. Of course, this will mean that the sub-carrier spacing will be little bit smaller than the one specified in the standard, but this should cause any problem in the system performance. This is a significant improvement indeed. A suitable algorithm to perform this task can be found by making use of the intelligent channel estimation algorithm that is envisioned in Chapter 7. In this way, SNR gain or capacity enhancement can be assured without increasing any processing.

8.8 OFDM Based Multi-User Systems

OFDM is now being stuied for multi-user systems, such as cellular wide area networks. Basically, orthogonal frequency resources (i.e. orthogonal subcarriers) can be shared among users, which is the simplest multiple access technique based on OFDM modulation. Besides, there are a number of *hybrid* multiple access techniques that can be found in literature. Here *hybrid* means an amalgamation of OFDM and multiple access techniques (with the main accent to the spread spectrum) to provide an efficient multi-user scenario with very high data rate. The following is a list of some of the most known multiple access techniques that can be found in the literature:

1. OFDMA (Orthogonal Frequency Division Multiple Access)
2. OFDMA-FSCH (OFDMA Fast SubCarrier Hopping, downlink of Flash-OFDM)

3. OFDMA-SSCH (OFDMA Slow SubCarrier Hopping, uplink of Flash-OFDM)

4. MC-CDMA (Multi-Carrier CDMA) [54]

5. OFDM-CDMA-SFH (OFDM-CDMA Slow Frequency Hopping) [55]

6. VSF-OFCDMA (Variable Spreading Factor - Orthogonal Frequency and Code Division Multiple Access)

A comprehensive analysis of these access techniques can be found in [19]. At this moment, tremendous research efforts are being spent on these techniques all over the world. It is undoubtedly a very important issue that still requires a lot more attentions and efforts.

8.9 Miscellaneous Research Directions

1. Other than the basic OFDM system for improvement, researchers in the field of OFDM have been working on improved channel coding techniques which can be designed for special use with OFDM. New coding schemes are being thought of which shall be integrated with OFDM. A lot of research is being done on turbo-coded OFDM for uplink and LDPC coded OFDM for downlink.

2. Work is also in progress in using Walsh-Hadamard-OFDM or even DCT-OFDM system. An additional transformation is being tried out to achieve improved performance in these systems. Additionally a new line of thought is coming up towards pseudo orthogonal subcarrier spacing.

3. Smart antennas could still be thought of for OFDM systems since MIMO systems increases the computational complexities several times. Smart antennas can be used for interference mitigation and range extension of a cellular system. It can be very good solution, considering that MIMO complexities may be prohibitive to implement in the MS receiver.

4. Hybrid systems could be the next wave of communication technology after OFDM. A lot of interest is gathering around MC-CDMA, UWB-OFDM and a mix of frequency hopping, CDMA and OFDM.

Chapter 9

Conclusion

This report intended to give a first hand guide to understand the basics of OFDM. We specially emphasized on PHY layer AI design related issues. Our main goal was to present the ins and outs of any OFDM system, so we concentrated on several important issues like synchronization and channel estimations etc. One of the primary aims was to identify possible future scopes for research in OFDM system design. So considerable amount of efforts was spent on studying existing wireless standards based on OFDM. We strongly believe that with the success OFDM PHY standards such as IEEE 802.11a, OFDM will continue to play a great role in all future generations of wireless systems. As we see it, MBWA systems will be one of the most dominating wireless systems in future and OFDM will be the core technology for MBWA systems, we will be concentrating our future efforts in designing a robust and power efficient MBWA systems for future applications.

As we have mentioned in the preface, this report is written at the very beginning of our PhD studies. As we proceed along with our studies, and we grow with knowledge and competencies in OFDM, this report will be constantly updated and improved.

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