Universidade de Aveiro Departamento de Electrónica, Telecomunicações e Informática,
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Eduardo Manuel Diogo Conde

Técnicas de detecção para o sistema LTE no sentido ascendente

Detection techniques for the uplink of LTE

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Dissertação apresentada à Universidade de Aveiro para cumprimento dos requisitos necessários à obtenção do grau de Mestre em Engenharia Electrónica e Telecomunicações, realizada sob a orientação científica do Prof. Dr. Adão Paulo Soares da Silva, Departamento de Electrónica, Telecomunicações e Informática, Universidade de Aveiro; e do Prof. Dr. Atílio Manuel da Silva Gameiro, Departamento de Electrónica, Telecomunicações e Informática, Universidade de Aveiro.

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Palavras Chave LTE, OFDM, OFDMA, SC-FDMA, MIMO, ZF, MMSE, IB-DFE

Resumo

Nos dias de hoje a internet móvel está em rápida expansão devido aos *smartphones* e outros dispositivos móveis. Cada vez mais o acesso à Internet é feito por ligações sem fios, permitindo que os utilizadores se liguem à Internet em qualquer lugar. Devido ao aumento do número de utilizadores e de serviços e por consequente o aumento de largura de banda, é necessário implementar novas arquitecturas que consigam acomodar essas taxas de transmissão e que também sejam rentáveis para as operadoras de telecomunicações.

O LTE é assim uma das tecnologias promissoras para o aumento de tráfego verificado nestes últimos anos nas redes celulares. Por detrás da tecnologia LTE estão sistemas de acesso múltiplo que usam múltiplas portadoras ortogonais baseadas no OFDM, sendo o OFDMA usado para o downlink e o SC-FDMA para o uplink. Outro conceito usado pelo LTE são os sistemas MIMO que aumentam a eficiência espectral e a taxa de transmissão de dados.

Esta dissertação tem por base a camada física do LTE (camada 1 do modelo OSI), nomeadamente a comunicação de dados no sentido ascendente baseada na técnica de acesso múltiplo SC-FDMA. Foi implementada numa plataforma de simulação um receptor com um equalizador no domínio da frequência iterativo, onde foram usadas várias configurações MIMO com o meio partilhado por vários utilizadores. Os resultados obtidos permitem comparar a eficácia deste tipo de equalizadores versus os equalizadores lineares como o ZF e MMSE tanto na eliminação das interferências como na atenuação do efeito *near-far*.

Os resultados obtidos mostraram-se bastante satisfatórios conseguindo-se remover grande parte dos efeitos da interferência entre símbolos e/ou de múltiplo acesso obtendo-se um desempenho muito próximo do obtido considerando um filtro adaptado. Para além disso a diferença de potência entre os utilizadores não afectou o desempenho do equalizador.

Keywords LTE, OFDM, OFDMA, SC-FDMA, MIMO, ZF, MMSE, IB-DFE

Abstract

Nowadays, the mobile web is in an accelerated expansion brought by the introduction of smartphones and other mobile devices. Access to the Internet by wireless networks is more and more used because it allows the users to connect to the Internet anywhere. Thus, with the increase of the number of users and services whose by consequence increase the used bandwidth, there is the need to implement new architectures to cope with the demanded transmission rates while being rentable for mobile network carriers.

LTE is one of the most relevant technologies to the observed increase of traffic in the last years in cellular networks. Behind this technology, multiple access systems which use multiple orthogonal subcarriers based on OFDM are used, being OFDMA used for the downlink and SC-FDMA for the uplink. Another concept used by LTE is the MIMO systems which increase the spectral efficiency and the data rate.

This thesis work is based on the physical layer of LTE (layer 1 of OSI model), namely the uplink communication with its multiple access technique SC-FDMA. It was implemented in a simulation platform an iterative frequency-domain receiver where multiple MIMO configurations were used in a multiple access environment. The obtained results allows comparison of the efficacy in removing interference and also the attenuation of the near-far effect between linear frequency domain receivers such as ZF and MMSE and iterative frequency domain receivers.

The results were successful in the cancellation of inter-symbol interference and multiple access interference. The obtained performance is very close to the one obtained by the matched filter. Moreover, the difference of power between the users did not affect the performance of the equalizer.

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Acronyms

1xEV-DO	One Carrier Evolved, Data Optimized. 3
1xRTT	One Carrier Radio Transmission Technology. 3
2G	Second Generation. 1, 2, 7, 11
3G	Third Generation. 3, 5, 7, 11, 39, 56
3GPP	Third Generation Partnership Project. 2–4, 8–10
3GPP2	Third Generation Partnership Project 2. 3, 4, 7
4G	Fourth Generation. 5, 56
8-PSK	8 - Phase Shift Keying. 2
ADC	Analog to Digital Converter. 27
AES	Advanced Encryption Standard. 4
AWGN	Additive White Gaussian Noise. 2
BER	Bit Error Rate. 2, 49, 51, 52, 54, 55
BPSK	Binary Phase Shift Keying. 25
BS	Base Station. 5, 13, 21, 25, 29, 30, 40, 41, 49, 51, 53, 56, 57
CAPEX	Capital Expenditure. 8
CDMA	Code Division Multiple Access. 2, 3, 5, 7, 56
CDMA2000	Code Division Multiple Access 2000. 3, 7, 9, 10
CP	Cycle Prefix. 5, 16, 18–20, 22–24, 40, 47
CPC	Continuous Packet Connectivity. 3
CTC	Convolutional Turbo Code. 46, 49, 55, 57
DAC	Digital to Analog Converter. 25
DFE	Decision Feedback Equalization. 27
E-UTRA	Evolved UMTS Terrestrial Radio Access. 4, 13, 21
E-UTRAN	Evolved UMTS Terrestrial Radio Access Network. 9–14
EAP	Extensible Authentication Protocol. 4
EDGE	Enhanced Data rates for GSM Evolution. 2, 7
eNodeB	Evolved Node B. 10, 11, 22
EPC	Evolved Packet Core. 9–11

EPS	Evolved Packet System. 9
FDD	Frequency Division Duplexing. 8, 21
FDE	Frequency Domain Equalizer. 36–38, 43, 45, 48
FDM	Frequency Division Multiplexing. 16
FFT	Fast Fourier Transform. 16, 20, 21, 24–27, 34, 47, 48
GERAN	GSM EDGE Radio Access Network. 9, 13
GPRS	General Packet Radio Service. 2, 10
GSM	Global System for Mobile communication. 2, 3, 7, 9
HSDPA	High Speed Downlink Packet Access. 3, 5, 12
HSPA	High Speed Packet Access. 3, 7
HSPA+	HSPA Evolution. 3, 7
HSS	Home Subscriber Server. 10
HSUPA	High Speed Uplink Packet Access. 3, 5, 12
IB-DFE	Iterative Block Decision Feedback Equalizer. 39, 42, 47, 49, 57
ICI	Inter Carrier Interference. 18, 19
IFFT	Inverse Fast Fourier Transform. 16, 19, 20, 25–27, 47
IMS	IP Multimedia Subsystem. 8
IMT-2000	International Mobile Telecommunications-2000. 3
IP	Internet Protocol. 7, 9–11
IS-95	Interim Standard 95. 2, 3, 7
ISI	Inter-Symbol Interference. 5, 16–20, 27, 35, 37, 39, 42, 43, 45, 49, 51, 57
ITU	International Telecommunications Union. 3
LMSC	LAN/MAN Standard Committee. 3
LSTI	LTE/SAE Trial Initiative. 7
LTE	Long Term Evolution. 4–12, 15–17, 20, 22, 23, 25, 28, 30, 36, 39, 56
MAI	Multiple Access Interference. 5, 45, 49, 51, 52, 57
MBMS	Multimedia Broadcast Multicast Service. 13
MF	Matched Filter. 49, 51, 52, 54, 57
MIMO	Multiple Input Multiple Output. 3–7, 12, 15, 28–31, 39, 40, 51, 53, 55, 56
MISO	Multiple Input Single Output. 28, 32
MME	Mobility Managment Entity. 10–12
MMSE	Minimum Mean Square Error. 6, 27, 35, 37, 38, 49
OFDM	Orthogonal Frequency Division Multiplexing. 3, 5, 12, 16, 18–20, 22–24, 34, 46, 49, 56

OFDMA	Orthogonal Frequency Division Multiple Access. 4, 6, 8, 15, 20, 21, 25, 26, 35, 56
OPEX	Operational Expenditure. 8
P-GW	Packet Data Network Gateway. 10, 12
PAPR	Peak-to-Average Power Ratio. 5, 8, 12, 21, 25, 27
PCRF	Policy Control and Charging Rules Function. 10
PDN	Packet Data Network. 9, 10
PIC	Parallel Interference Cancellation. 43
PLMN	Public Land Mobile Network. 12
PRB	Physical Resource Block. 22–24
QAM	Quadrature Amplitude Modulation. 3, 8, 17, 24, 25, 40, 46, 48
QoS	Quality of Service. 10, 13, 14
QPSK	Quadrature Phase Shift Keying. 2, 8, 15, 24, 25, 40, 46, 48, 49
RAN	Radio Access Network. 11
RF	Radio Frequency. 25, 26
RRM	Radio resource management. 11
S-GW	Serving Gateway. 10–12
SAE	System Architecture Evolution. 9, 10
SC-FDE	Single Carrier Frequency Domain Equalizer. 5
SC-FDMA	Single Carrier Frequency Division Multiple Access. 5–8, 12, 15, 21, 25–27, 35, 36, 46, 47, 56
SFBC	Space-Frequency Block Code. 33–35
SFC	Space-Frequency Coding. 33
SIMO	Single Input Multiple Output. 28
SINR	Signal to Interference-plus-Noise Ratio. 29–31, 45
SISO	Single Input Multiple Output. 28, 30, 49
SNR	Signal Noise Ratio. 19, 20, 29, 30, 35, 37
STBC	Space Time Block Coding. 31–34
STC	Space Time Coding. 31
TD-DFE	Time-Domain Decision Feedback Equalizer. 39
TDD	Time Division Duplexing. 8, 21
TDMA	Time Division Multiple Access. 2, 7
UE	User Equipment. 5, 8–11, 13, 21, 23, 25, 39, 40, 43, 45, 49, 51, 53, 56, 57
UMB	Ultra Mobile Broadband. 4, 7, 20
UMTS	Universal Mobile Telecommunications System. 7, 9, 10
UTRA	UMTS Terrestrial Radio Access. 8

UTRAN	UMTS Terrestrial Radio Access Network. 9, 13
voIP	voice over Internet Protocol. 8
W-CDMA	Wideband Code Division Multiple Access. 3, 9
WiMAX	Worldwide Interoperability for Microwave Access. 3, 4, 8–10, 20
ZF	Zero Forcing. 6, 35, 37, 38, 49

Chapter 1

Introduction

1.1 Evolution of Cellular Systems

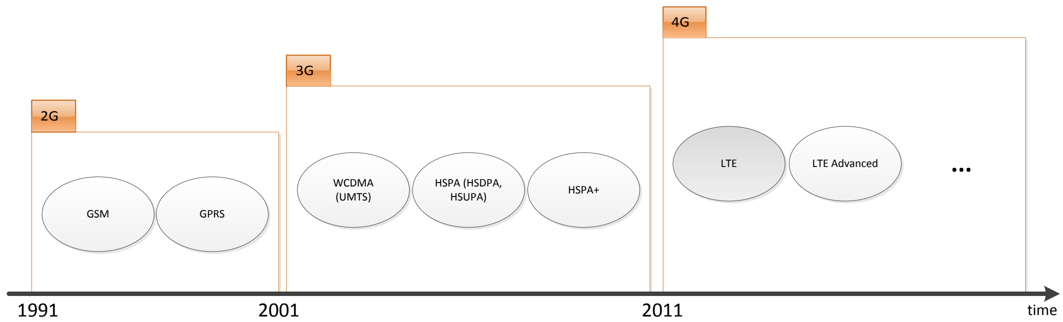

Figure 1.1: Evolution of Mobile Technologies

Since introduction of cellular systems in early 1980s, characterized by voice analog communications, and then followed by the Second Generation (2G) in the 1990s where analog communications were replaced by the digital counterparts, cellular communications have comprised several generations (see Fig. 1.1) and experienced an exponential growth where it is believed that there are over 6 billion subscriptions at the first quarter of 2012 [1] as shown in Fig. 1.2.

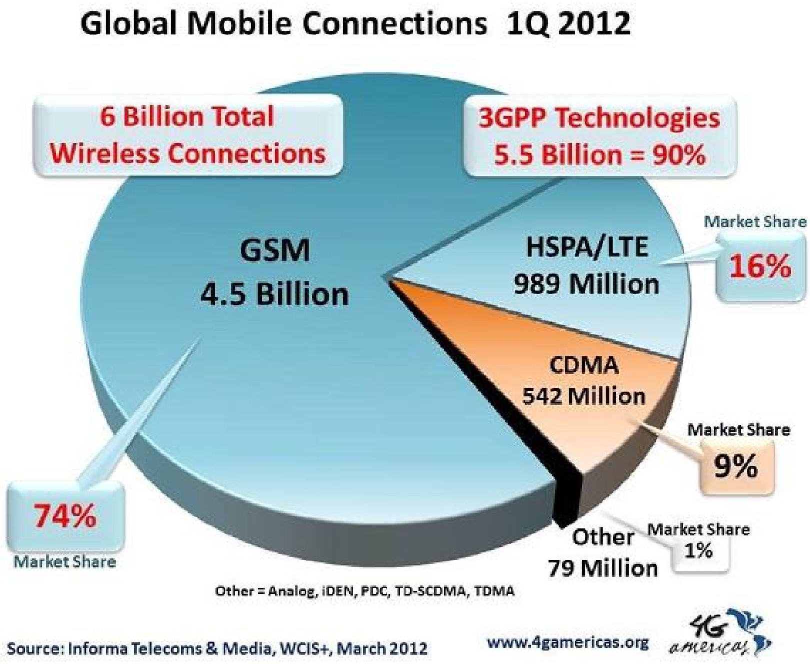

Figure 1.2: Subscriptions of Cellular Systems [1]

The second generation not only brought better voice quality but also greater capacity to allow more mobile terminals, and thus mobile communications become more widespread with Time Division Multiple Access (TDMA) based Global System for Mobile communication (GSM) and Code Division Multiple Access (CDMA) based Interim Standard 95 (IS-95) or CDMAone, among other 2G technologies. Despite of the second generation being designed to support only voice communications, in the later releases, data transmission were implemented and those standards become the 2.5 generation of mobile communications. One of the biggest examples of the new data transmission standard is the evolution of GSM, the General Packet Radio Service (GPRS) a best-effort service that used packet switched data transmissions and could provide data rates from 56 up to 114 kbps [2]. Packet switched transmissions would become more and more important with the rise of the mobile internet. In addiction to the 2G standards, in 2003 Enhanced Data rates for GSM Evolution (EDGE) was standardized by Third Generation Partnership Project (3GPP) being an upgrade of the GSM/GPRS that achieved more capacity by using a better method for coding (8 - Phase Shift Keying (8-PSK) instead of Quadrature Phase Shift Keying (QPSK)) that allowed 3 bits per symbol and symbol rates up to 270 kbps. Moreover by using EDGE channel coding and a 2-FSK/8-PSK modulation scheme the data rate can be increased by 33% with just 2 dB degradation in Eb/N0 ratio with equal Bit Error Rate (BER) in the Additive White Gaussian Noise (AWGN) channel [3].

Although the 2nd generation already had packet switching in the later releases, the voice communications were the majority of the network traffic, so there was the need for a new infrastructure to support multimedia as the main service to attract users for data based traffic.

The Third Generation (3G) was first introduced when there was a demand for higher bandwidth by the International Telecommunications Union (ITU) initiative on the International Mobile Telecommunications-2000 (IMT-2000). In the IMT-2000 it were set several requirements like the need of a 3G technology to provide peak data rates of at least 200 kbps. For the 3rd generation both GSM and Code Division Multiple Access 2000 (CDMA2000) created their own standards based on the CDMA technology. 3GPP standardized Wideband Code Division Multiple Access (W-CDMA) and Third Generation Partnership Project 2 (3GPP2) standardized CDMA2000-1x and CDMA2000-3x which use three 1.25 Mhz subcarriers. While CDMA2000 transmits on 1.25 MHz radio channels and is backwards compatible with IS-95, W-CDMA uses 5 Mhz bandwidth channels and isn't backwards compatible with GSM. That was one of the main reasons that hindered the penetration of W-CDMA in the North American market, since there was a large number of IS-95 devices and only a few allocations of bandwidths with 5 Mhz [4].

W-CDMA 3GPP's standard was later upgraded into High Speed Packet Access (HSPA), the combination of High Speed Downlink Packet Access (HSDPA) on Release 5 and High Speed Uplink Packet Access (HSUPA) on Release 6 that offered speeds up to 14.4 mbps downlink and 5.76 mbps uplink, while still using the same frequency bands. In release 7, HSPA Evolution (HSPA+) speeds were increased further into 168 mbps for the downlink and 22 mbps for the uplink and brought significant improvements such as Continuous Packet Connectivity (CPC), Multiple Input Multiple Output (MIMO) techniques and 64-Quadrature Amplitude Modulation (QAM) [5]. In Release 8 Dual Cell HSDPA was introduced. Dual Cell is characterized by the use of joined scheduling between two HSDPA carriers. The main advantage of dual-cell is that it improves the user throughput for a given load, for any location in the cell, even at cell edge, and thus increasing the overall capacity [6].

On 3GPP2's side, CDMA2000 1x, also known as One Carrier Radio Transmission Technology (1xRTT) has the same bandwidth as the second generation IS-95 and uses a duplex pair of 1.25 Mhz channels adding more 64 traffic channels to the original set of 64 while supporting downlink speeds up to 153 kbps. This standard was not considered in third generation because in practice it didn't fulfil the data rates that were considered in the standards. The 3G project that was standardized by 3GPP2 was the CDMA2000 One Carrier Evolved, Data Optimized (1xEV-DO). The revision A of 1xEV-DO supports peak data rates of 3.1 Mbps for the downlink and 1.8 Mbps for the uplink and introduced the Orthogonal Frequency Division Multiplexing (OFDM) waveform that allowed mobile network carriers to offer lower cost multicast services. This allowed new services and protected the investments of the service providers that employed CDMA2000 1x networks in the past [7].

In parallel with the development and deployment of HSPA, LAN/MAN Standard Committee (LMSC) introduced the standard IEEE 802.16, that has been commercialized under the Worldwide Interoperability for Microwave Access (WiMAX) label by the WiMAX Forum. The mobile version of the standard IEEE 802.16e is referred as Mobile WiMAX and uses the multiple access technology Orthogonal Frequency Division Multiple Access (OFDMA) improving multi-path performance in non-line-of-sight environments. Mobile WiMAX offer scalability in both radio access technology and network architecture. The main features of Mobile WiMAX are the inclusion of MIMO techniques which brings higher data rates, scalability from different channelizations (from 1.25 to 20 MHz), improved security with Extensible Authentication Protocol (EAP) based authentication and Advanced Encryption Standard (AES) based authenticated encryption and optimized handover schemes with latencies less than 50 milliseconds [8].

With the introduction of mobile WiMAX, 3GPP started to develop Evolved UMTS Terrestrial Radio Access (E-UTRA) while 3GPP2 developed Ultra Mobile Broadband (UMB), both based on OFDMA and similar network architecture of Mobile WiMAX. The 3GPP2's UMB was dropped out in November 2008 by Qualcomm in the favor of 3GPP's E-UTRA [9] that is better known as Long Term Evolution (LTE). It is expected that the number of LTE subscribers will increase exponentially in the near future as shown in Fig. 1.3. This system will be presented in detail in Chapter 2.

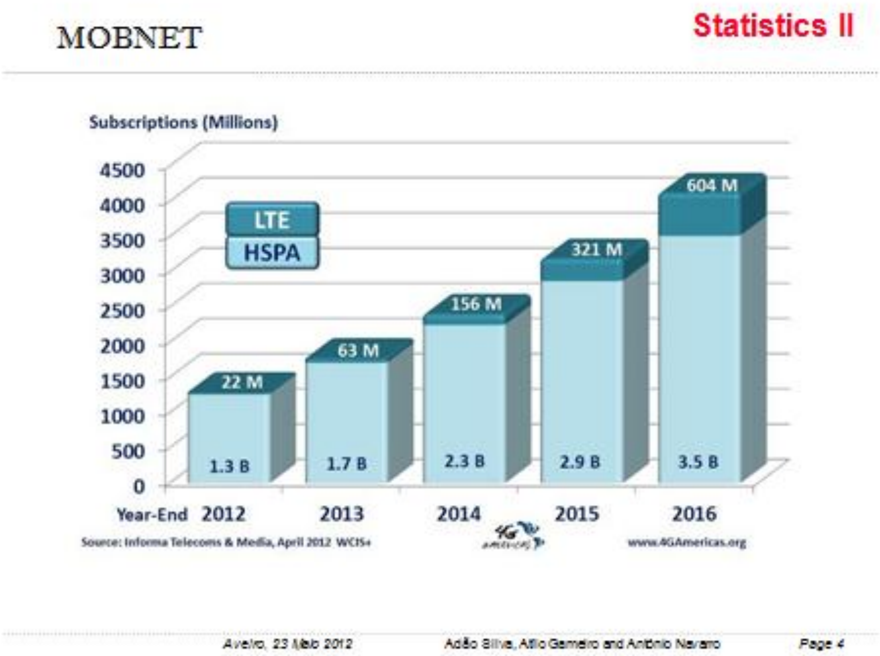

Figure 1.3: Estimation of Future Subscriptions of LTE [1]

1.2 Motivation and Objectives

Since there is a trend for the increase in the demand for broadband services, higher and higher transmission rates are required. The 3Gs technologies based on CDMA are unable to cope with the increasing demands, because they are matured to the point where they have limitation in bit rate and relatively high latency which is too big to handle traffic with big data rate such as mobile TV [10], despite of the improvements brought by HSDPA and HSUPA.

Long Term Evolution (LTE) was chosen for the next generation of mobile communications labeled as a Fourth Generation (4G) and is based on new technologies like Single Carrier Frequency Division Multiple Access (SC-FDMA) and MIMO departing from 3G based CDMA. LTE provides higher data rates (over 100 Mbps), brings higher spectral efficiency than CDMA based systems, simpler network architecture and less operation costs for the mobile network carriers.

Regarding the uplink, a new air interface technique such as SC-FDMA was developed. SC-FDMA is a hybrid modulation scheme that combines multipath resistance by OFDM technology with low Peak-to-Average Power Ratio (PAPR) of traditional single-carrier (SC) formats. As with OFDM-based schemes, SC-FDMA includes a Cycle Prefix (CP), long enough to cope with maximum channel delay, so that linear frequency-domain equalization techniques can be used at the receiver side. These Single Carrier Frequency Domain Equalizer (SC-FDE) schemes are closely related to the ones employed in multicarriers based systems. In the current LTE standard only single user and/or multi-user closed-form equalizers are considered. However, in a severe time-dispersive channels, where high levels of interference are experienced, the performance of these equalizers is very poor [11].

The main objective of this thesis is to develop and assess an iterative multi-user SC-FDE for the uplink LTE system. For that, a simulation platform based on LTE specification was implemented. It is assumed that a set of single antenna User Equipment (UE)s share the same frequency to transmit data to the Base Station (BS) equipped with multiple antennas, forming a "virtual" MIMO system. The implemented algorithm removes efficiently both the Inter-Symbol Interference (ISI) and Multiple Access Interference (MAI). The implemented platform is flexible since it allows to configure several parameters such as number of users, number of antennas at the BS, number of subcarriers, different modulation and channel coding schemes. The results are compared against both the single user and multi-user closed form frequency domain equalizers employed on the LTE standard.

1.3 Outline

After the presentation of the evolution of cellular systems and the motivation, objectives and contributions of this thesis, we will proceed with the following Chapters:

In Chapter 2 we introduce the LTE system as a whole, by describing the basic concepts, its main targets and specifications and finally the network architecture.

Following, in Chapter 3 we introduce the multicarrier systems that are the base of the multiple access schemes of LTE, Orthogonal Frequency Division Multiplexing (OFDMA) and Single-Carrier Frequency Division Multiple Access (SC-FDMA) which are described further in this chapter.

In Chapter 4 we first introduce the MIMO schemes and the concepts of diversity. Later in this chapter we will present transmit diversity based on time and frequency block codes and linear receiver diversity equalizers (Zero Forcing (ZF) equalizer and Minimum Mean Square Error (MMSE) equalizer) for SC-FDMA.

In Chapter 5 we develop the Iterative Frequency-Domain Receivers which is the core of this thesis. The conditions and parameters used to obtain the results are defined. Also, the proposed scheme is assessed in several scenarios.

For the last but not the least in Chapter 6 we will conclude the thesis and provide guidelines for future research.

Chapter 2

Long Term Evolution System

LTE is part of the evolutionary chain of the 3G standards, starting from EDGE, then Universal Mobile Telecommunications System (UMTS) and HSPA and immediately after HSPA+.

In the previous cellular networks there has been fragmentation of the cellular technologies, for example, the second generation had TDMA based GSM or CDMA based IS-95, the third generation had UMTS and CDMA2000 and now finally with the introduction of LTE, it may be the first truly global standard since 3GPP2's UMB was abandoned [9], however LTE frequency bands will be different depending on the countries which LTE is implemented. The operators that used different technologies finally reached consensus due to the all Internet Protocol (IP) network architecture and the new radio access techniques such as SC-FDMA and MIMO that offers higher performance beyond what the CDMA technology could offer. LTE isn't the end of previous technologies allowing the coexistence with other 2G and 3G technologies.

First LTE proposal was in 2004 by NTT DoCoMo of Japan but the standard was only created by means of 2005. The main goal set for LTE was the raise of both the speed and capacity of mobile data networks using relatively new modulation and digital processing techniques. Also the network architecture was simplified into a IP based system making it incompatible with 2G and 3G networks. In 2007 the LTE/SAE Trial Initiative (LSTI) alliance was created with the intent of accelerating the acceptance and deployment of LTE/SAE as the logical choice of the industry for next generation networks [12]. The LTE standard was finished in December of 2008 and it was launched for the first time by TeliaSonera in Oslo and Stockholm on December 14th of 2009.

Within the specifications of LTE, for the radio access network it was specified downlink and uplink peak rates of 100 Mbps and 50 Mbps respectively in a 20Mhz spectrum, less than 5 ms user plane latency, the ability to manage faster moving LTE devices (up to 350Km/h) and it defined cell coverage between 5 and 100km. Also defined in the standard there are several modulation schemes available for the transmission of data. Higher modulation orders increase the data rate with the inconvenient of being more susceptible to signal errors and path loss. This means that distance between the base station and the mobile terminal matters in the selection of the modulation scheme. For downlink the main modulation schemes supported are QPSK, 16-QAM and 64-QAM while for the uplink 64-QAM is optional at the UE [13].

To separate downlink and uplink traffic LTE supports Time Division Duplexing (TDD), Frequency Division Duplexing (FDD) or both at same time. LTE also supports scalable carrier bandwidths (1.25, 2.5, 5, 10, 15 and 20MHz) instead of UMTS Terrestrial Radio Access (UTRA)'s static 5 Mhz channels by varying the number of subcarriers for transmission to achieve high flexibility in channelization.

While IEEE 802.16e (WiMAX) uses OFDMA for both the downlink and uplink, one of the main advantages of LTE is that it uses SC-FDMA for the uplink, instead of OFDMA. Since OFDMA has high PAPR it requires inefficient power amplifiers, therefore it would reduce the autonomy of mobile devices significatively. SC-FDMA was then chosen for the uplink having lower PAPR over OFDMA [14] and shares most of the OFDMA architecture.

In a nutshell, some of the main targets of LTE are listed below [15]:

- Achieving higher peak data rate - 100Mbps Downlink and 50Mbps Uplink
- Increasing the bit rate in the edge of the cells with the same sites as today
- Significatively Improved spectrum efficiency
- Latency less than 10ms
- Scalable bandwidth
- Compatibility with other 3GPP and non 3GPP systems
- Reduced costs Capital Expenditure (CAPEX) and Operational Expenditure (OPEX) and cost effective migration from Release 6 of UTRA radio interface and architecture
- Balanced system, terminal complexity and power consumption
- Support for further enhanced IP Multimedia Subsystem (IMS) and core network
- Support for different kinds of services like voice over Internet Protocol (voIP)
- Optimized system for low mobility but supporting high mobility
- Operation in both paired and unpaired spectrum

2.1 Network Architecture of LTE

LTE being the evolution of UMTS not only brought us a new radio access interface (Evolved UMTS Terrestrial Radio Access Network (E-UTRAN)) but also a simplified network architecture under the term System Architecture Evolution (SAE) where the Evolved Packet Core (EPC) belongs. Evolved Packet System (EPS) is composed by both LTE and SAE.

The new IP-based EPC network architecture is based on an evolution of the existing GSM/W-CDMA core network with simplified operations and cost-efficient deployment. This new EPC network is designed to optimize network performance, improve cost-efficiency and facilitate the uptake of mass-market multimedia services [16].

LTE has dropped the circuit-switched model present in older cellular system to support only packet-switched services, i.e., the UE connects with the Packet Data Network (PDN) including voice traffic based only on the IP. LTE then must provide such connection without any disruption when the user is in mobility, higher data rates, lower latency and previously defined Quality of Service (QoS) [17]. Finally it supports both interworking and mobility with heterogeneous networks like 3GPP's systems (UMTS Terrestrial Radio Access Network (UTRAN) and GSM EDGE Radio Access Network (GERAN)) and non 3GPP systems like WiMAX and CDMA2000.

2.1.1 Core Network

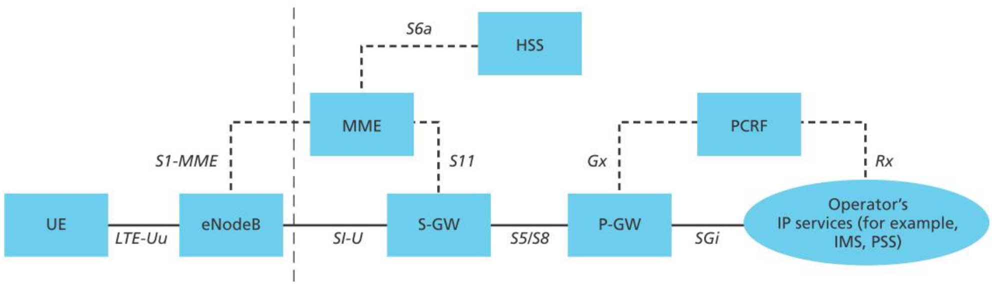

Figure 2.1: EPS Elements of LTE Architecture [17]

The overall network is comprised by the Core Network (EPC) and the Access Network (E-UTRAN). The core network manages the control of the UE and establishes the bearers. The EPC is further composed by the following logical nodes represented in the Fig. 2.1 and described below [17], [18]:

- **Packet Data Network Gateway (P-GW)** - Allocates the IP address for the UE, enforces Quality of Service (QoS) to guarantee the bit rate and filters downlink user IP packets into different QoS based bearers. Another feature of the P-GW is the ability to interwork with non-3GPP architectures, such as CDMA2000 and WiMAX networks.

- **Mobility Managment Entity (MME)** - This control node processes the signalling between the UE and the Core Network. It supports several functions that are classified as functions to aid bearer management and functions for connection management. The first function is used for establishment, maintenance and release of the bearers and the latter for establishing connections and is responsible for the security. Both of this function types are handled by different layers of the protocol between the UE and CN (NAS protocol), the session manager layer and the connection and mobility layer respectively.

- **Serving Gateway (S-GW)** - When the UE moves between different Evolved Node B (eNodeB)s, the Serving Gateway acts as a mobility anchor for the data bearers and retains them when the UE is in idle state. When the MME initiates the paging of the UE, it buffers the downlink data. While the P-GW allows interworking with non 3GPP's architectures, S-GW allows interworking with technologies such as GPRS and UMTS which are 3GPP's architectures. S-GW also performs administrative functions for lawful interception and logging data for charging proposes.

- **Policy Control and Charging Rules Function (PCRF)** - This function is responsible for the control of policy and decision making. It also provides the QoS authorization based on user's subscription profile.

- **Home Subscriber Server (HSS)** - Contains the user's SAE subscription data like the QoS profile and access restrictions for roaming. This function also holds the information about the PDN where the user can connect in the form of access point name or an PDN address (based on subscribed IP address(es)). Dynamic information is also stored like identity of the MME which the user is attached and registered. For the last but not the least, it may integrate the authentication server to generate vectors of authentication and security keys.

2.1.2 Access Network

The simplified LTE access network (E-UTRAN) only consists of a network of eNodeBs without a centralized controller for normal traffic. The eNodeBs are usually interconnected by the X2 interface and with the EPC by the S1 interface. The S1 interface is divided into S1-MME for interworking with the MME and S1-U for connections with the S-GW. The protocols used between the different eNodeBs are known as the "AS Protocols". The access network of LTE is depicted in the Fig 2.2.

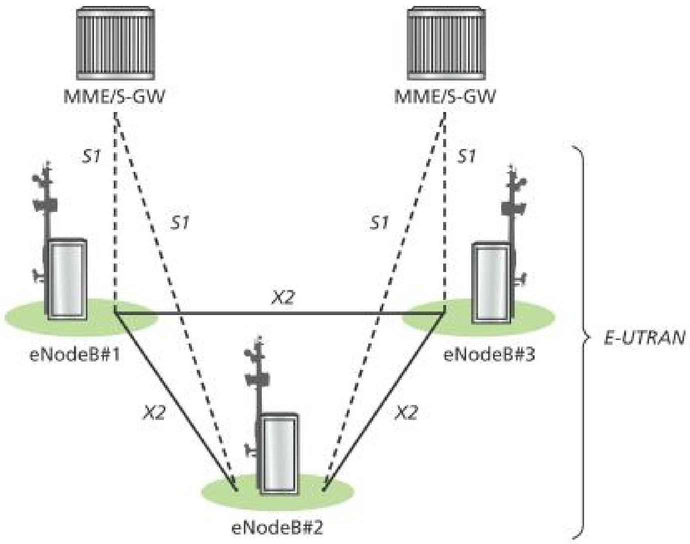

Figure 2.2: Overall E-UTRAN Architecture [17]

The functions of E-UTRAN are [17] [18]:

- **Radio resource management (RRM)** - Controls radio-related functions such as the control of the radio mobility and admission, scheduling or dynamic allocation of the resources for downlink and uplink traffic.

- **Header Compression** - Compress the IP packet headers, lowering the size of the overhead.

- **Security** - Encryption of data sent through the radio interface

- **Connectivity and Signalling to the EPC**

The functions described above, unlike the previous 2G and 3G technologies reside only in the eNodeBs allowing tight interaction between the layers of the Radio Access Network (RAN), therefore it reduces latency and improves system efficiency. Without a processing-intensive centralized controller there is potential to reduce costs and isn't needed at all because LTE does not support soft handover. However there is an inconvenient of the lack of a centralized controller, if the UE moves all the information related to the UE must be transferred from one eNodeB to another making it susceptible to data losses. LTE posses mechanisms to prevent such losses in the operation of the X2 interface [17].

There is an important feature of the S1 interface, the "S1-Flex" which allows load sharing and reduces single points of failure for the Core Network Nodes by using multiple core network nodes (MME/S-GWs) to form a mesh network in a common geographical area (pool area). An eNodeB is then served by multiple MME/S-GWs and the UE uses the same MME if it is located within the pool area.

2.1.3 Roaming

Each operator has a Public Land Mobile Network (PLMN) and users move between those PLMNs, for example when they change between countries. After changing the PLMN they are connected to the E-UTRAN, MME and S-GW of the visited LTE network. One of the important roaming features of LTE is the ability to use the home network P-GW to be used allowing the user to access the home network services in a visited network [18].

2.2 Summary

The following table will summarize the requirements and specifications of LTE [19] [20]:

Table 2.1: Requirements of LTE Release 8

High Spectrum Efficiency	**OFDM in Downlink**	• Robust against multipath interference • Use of MIMO Techniques • Frequency Domain Channel Dependent Scheduling
	SC-FDMA in Uplink	• Low PAPR • Orthogonality between users in Frequency Domain
Mobility	• Optimized (0 - 15 Km/h) • High Performance (15 - 120 Km/h) • Maintain Connection (120 - 350 Km/h)	
Coverage	• Cells up to 5 km should met the throughput, spectrum efficiency and mobility targets) • Cells with 30 km may have a slight degradation • Cells with 100 km should be supported	
Spectrum Efficiency	**Downlink**	• In a loaded network, target for spectrum efficiency (bits/sec/Hz/site), 3 to 4 times Release 6 HSDPA
	Uplink	• In a loaded network, target for spectrum efficiency (bits/sec/Hz/site), 2 to 3 times Release 6 HSUPA
User Throughput	**Downlink**	• Average user throughput/MHz, 3 to 4 times Release 6 HSDPA
	Uplink	• Average user throughput/MHz, 2 to 3 times Release 6 HSUPA

Peak Data Rate	**Downlink**	• 100 Mbps within 20 Mhz spectrum allocation (5bps/Hz)
	Uplink	• 50 Mbps within 20 Mhz spectrum allocation (2.5bps/Hz)
Latency	**User Plane**	• From 5 ms in unload condition, up to 10 ms round-trip delay between UE and the BS
	Control Plane	• Less than 100 ms from camped state (Release 6 Idle mode) to active state (Release 6 CELLDCH)
Control Plane Capacity	• At least 200 users per cell in the active state	
Spectrum Flexibility	• Spectrum allocations in E-UTRA should be flexible in different pre-defined sizes such as 1.25, 1.6, 2.5, 5, 10, 15 and 20 Mhz for downlink and uplink in both paired and unpaired spectrum. • The system shall support content delivery over an aggregation of resources including Radio Band Resources (all spectrum available to the operator) in the same and different bands for both the downlink and uplink in adjacent or non adjacent channel arrangements.	
Co-existence and Inter-working with 3GPP Radio Access Technology	• Co-existence in the same geographical area and co-location with GERAN/UTRAN on adjacent channels. • E-UTRAN terminals which support operation in UTRAN and/or GERAN should be able to support measurement of, and handover to those radio access interfaces. • The interruption time while an handover ocurrs between E-UTRA and UTRAN (or GERAN) should be less than 300 ms.	
Radio Resource Management Requirements	• Enhanced support for end to end QoS. • Efficient support for transmission in higher layers. • Support of load sharing and policy management across different	
Complexity	• Minimize the number of options • No redundant mandatory features	
Further Enhanced MBMS	• While reducing terminal complexity: same modulation, coding, multiple access approaches and UE bandwidth than for unicast operation. • Provision of simultaneous dedicated voice and Multimedia Broadcast Multicast Service (MBMS) services to the user. • Available for paired and unpaired spectrum arrangements.	

Architecture	• E-UTRAN architecture is only packet based • It should support real-time and voice conversation class traffic in the packet based connections • E-UTRAN architecture shall minimize the presence of "single points of failure" • E-UTRAN architecture shall support an end-to-end QoS • Backhaul communication protocols should be optimized

Chapter 3

Multiple Access Schemes of LTE

As already mentioned LTE, uses new multiple access schemes for the radio access network, for the downlink it uses the multicarrier system OFDMA (Orthogonal Frequency Division Multiple Access) and for the uplink it uses SC-FDMA (Single-Carrier Frequency Division Multiple Access). OFDMA was chosen instead of CDMA because it has less complexity and it has greater robustness to reduce multipath interference by spreading the carriers all over the attributed spectrum. It can also achieve higher spectral efficiency with MIMO technologies and be easily scaled depending on the system requirements. Furthermore LTE does not suffer cell breathing issues [21]. The Fig. 3.1 shows the differences between OFDMA and SC-FDMA transmission. While the OFDMA sends the 4 QPSK symbols in parallel one for each subcarrier, the SC-FDMA sends them in series with 4 times the rate occupying the same bandwidth as 4 OFDMA symbols.

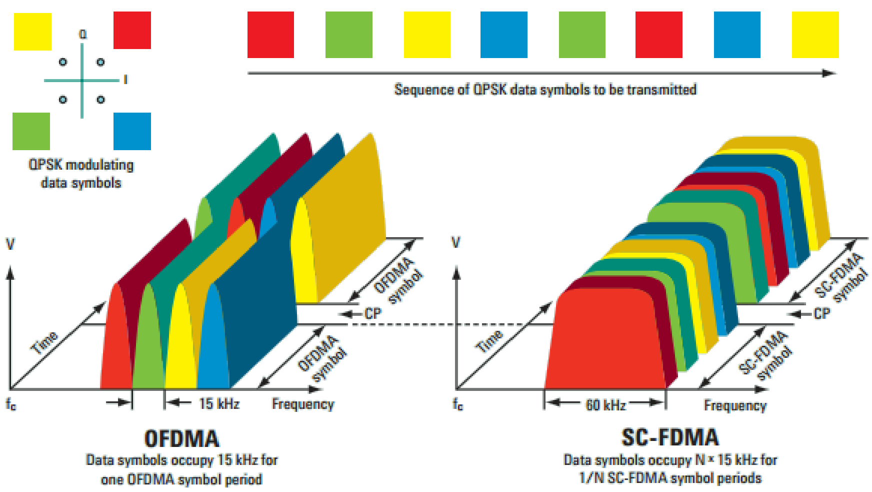

Figure 3.1: OFDMA vs SC-FDMA regarding the transmission of data symbols [22]

In this chapter before we describe the new multiple access schemes we start to describe the OFDM modulation principles that are fundamental for the overall comprehension of the new multiple access schemes.

3.1 Orthogonal Frequency Division Multiplexing

OFDM comes as an evolution of the conventional Frequency Division Multiplexing (FDM) where the subcarriers are overlapped because they are orthogonal to each other and thus there is no interference between the subcarriers reducing the needed spectrum up to 50%. The existence of guard bands in FDM to prevent such interference is not needed in the OFDM. The main difference between OFDM and FDM is that while in FDM the transmitting data is modulated in a single carrier, in OFDM the data is transmitted in parallel in several subcarriers where the data rate for each one is inversely proportional to the number of subcarriers used for transmission. One key advantage of OFDM is that we can use error correcting codes and interleavers in the frequency domain that can provide better gain than using those techniques in the time domain [23].

By using narrowband channels in OFDM, when we have a frequency selective fading response only some of the OFDM subcarriers are affected by the fades and thus instead of losing the entire symbol we just lose portions of the symbol and recovery may be possible by using proper coding and error correction.

For LTE the subcarrier spacing of OFDM is 15 khz that makes 66.7 μs for the symbol length. Since the symbols can overlap overtime leading to inter-symbol interference (ISI) a guard period is needed, known as CP and is inserted between each transmitted data symbol. CP is chosen to be slightly longer than longest expected delay spread in the radio channel. That length was set to 4.49 μs for the standard CP making the system able to cope with path delay variations up to 1.4 Km by just having a 7% loss of capacity [22].

The OFDM system is based in digital technology, using the Fast Fourier Transform (FFT) to move the signal to the frequency domain and the Inverse Fast Fourier Transform (IFFT) to move it back to the time domain without any losses of the original information if the requirements for digital processing are fulfilled like the requirement of minimum sampling rate. The length of FFT should be a multiple of 2 to reduce the amount of multiplications needed.

In a nutshell, the main characteristics of OFDM is the modulation using multicarriers, the property of orthogonality between the multicarriers, use of a cyclic prefix and easy implementation of frequency domain equalizers.

3.1.1 Modulation using Multicarriers

Single carrier systems modulates symbols by using just one carrier with data symbol period $T_b = 1/R$, where R is the baud rate in symbols/sec. In multi carrier systems, the available bandwidth B is split into N_c subbands, where each subband is a subcarrier and the space between those subcarriers is $\Delta f = W/N_c = 15$ Khz for LTE. The data stream is split into blocks of N_c that are distributed in each subcarrier and then sent in parallel increasing the data symbol period to $T_b = N_c/R$.

When we add more subcarriers the symbol period increases and therefore the data stream becomes more robust to channel distortion, impulse noise and fading. Those channel effects could result in inter-symbol interference (ISI) when the time dispersion becomes significant when compared to the symbol period [24].

3.1.2 Orthogonality

To ensure that the receiver can separate the previously overlapped subcarriers without interference, those subcarriers must be orthogonal to each other. The property of orthogonality allows simultaneous transmission of additional sub-carriers in a tight frequency space without interference from each other [25], however loss of orthogonality may occur and consequently degradation of the transmitted signal. Orthogonality is defined by the following equation:

$$\int_0^T S_i(t).S_j(t)dt = \begin{cases} C & \text{if } i = j \\ 0 & \text{if } i \neq j \end{cases} \tag{3.1}$$

If we have two different functions and we multiply each other and integrate them over a symbol period, the result is null if the functions are orthogonal to each other.

The resulted spectrum of any given subcarrier using a given QAM modulation is the $sinc(fT)$ function and it is characterized by having maximum amplitude at the centre and null amplitude in multiples of 1/T, where T is the duration of the QAM symbol . Since the main lobe of the $sinc(fT)$ function is narrow, there is the possibility of overlapping subcarriers making the peak of each subcarrier matching the null of the other subcarriers as shown in Fig. 3.2.

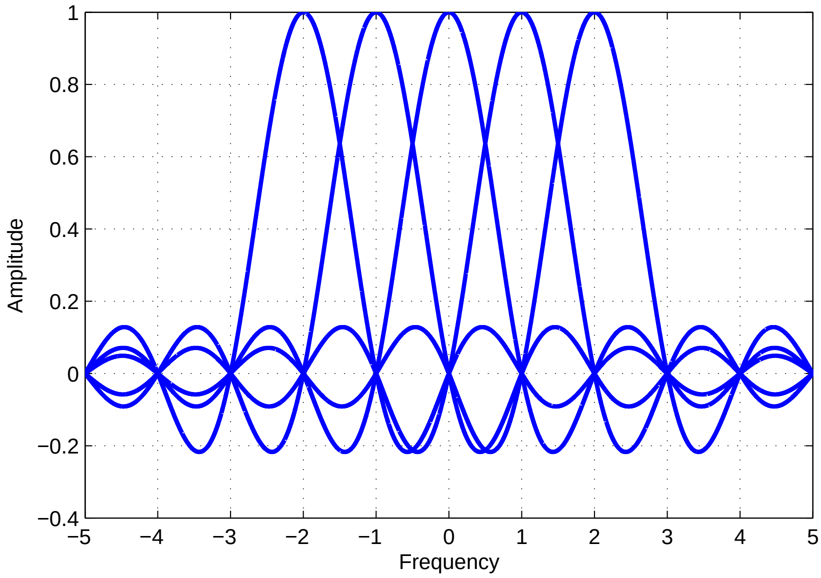

Figure 3.2: Subcarriers normalized in relation to 1/T

3.1.3 Cyclic Prefix

Since OFDM consist in transmitting in parallel using a large number of subcarriers, each one in a fraction of the available bandwidth, the symbol rate of the OFDM signal is N_c times lower than in an equivalent single carrier transmission. That property of having low symbol rate makes OFDM naturally resistant to ISI caused by multipath propagation and thus preventing loss of orthogonality. Nevertheless, loss of orthogonality may occur in a time-dispersive channel, the slower subcarriers will delay the symbols and those subcarriers may carry the remainders of the previous symbol making both symbols mixed in the DFT outputs, resulting in ISI. The delayed subcarriers will also be susceptible to Inter Carrier Interference (ICI).

Loss of orthogonality may also happen when the amplitude and phase of the subcarriers are not constant because the spectral shape of the subcarriers will have the nulls shifted into other frequencies, so the subcarriers will interfere with each other causing ICI.

To prevent one subcarrier from interfering with another, a CP is added. CP is no more and no less than the copy of the last portion of the data symbol into the front of the symbol in a time interval known as guard interval. By adding the CP the channel behaves as if the transmitted waveforms were from time minus infinite giving it a continuation, converting the linear convolution with the channel impulse response into a cyclic convolution, and thus ensuring orthogonality [26]. This is accomplished because the amount of time dispersion from the channel is smaller than the duration of the cyclic prefix .

Within each OFDM symbol a guard period is added increasing the length of the symbol waveform. The total length of symbol is now:

$$T_{symbol} = T_g + T_b \tag{3.2}$$

where T_g is the length of guard period and T_b the useful symbol time corresponding to the size of the IFFT which generated the OFDM signal. The addiction of CP is represented in the Fig. 3.3 below:

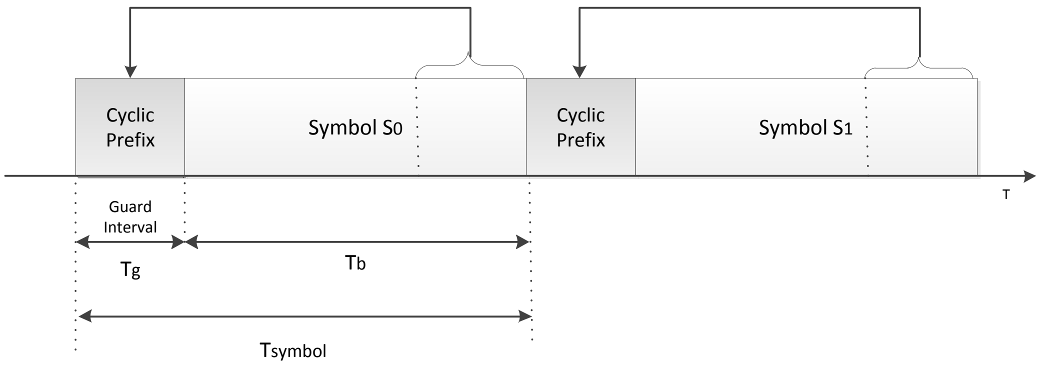

Figure 3.3: Addiction of Cyclic Prefix

The length of the Cyclic Prefix must be chosen carefully because there is a compromise between the increase of energy transmitted with the increase of CP length and having it's length corresponding at least to the offset between the slowest and the fastest subcarrier to make the received subcarrier unaffected by any dispersion to avoid both ISI and ICI.

The following equation give the Signal Noise Ratio (SNR) loss in dB caused by the insertion of CP:

$$SNR_{loss[dB]} = -10 \log_{10} \left(1 - \frac{T_g}{T_{symbol}} \right) \tag{3.3}$$

The linear form of the previous equation is given by:

$$SNR_{loss} = \frac{T_{symbol}}{T_{symbol} - T_g} \tag{3.4}$$

The duration of the short CP for LTE is 4.69 μs and the long CP is 16.67 μs with the LTE OFDM symbol period of $T_b = 66.67\mu s$. Knowing that $T_{symbol} = 66.67{+}T_g$, using the equation 3.4 we get SNR_{loss} of 1.0708 and SNR_{loss} of 1.25 for the short and long CP respectively. In terms of spectral efficiency defined by $\eta = SNR_{loss}^{-1}$ the short CP has a spectral efficiency of 93.4 % and the long CP has 80 % of spectral efficiency.

3.2 OFDMA

First we started to describe the multicarrier multiplexing technique OFDM, now we are going to describe multiple access technique chosen for the downlink traffic of LTE based on OFDM - the OFDMA.

OFDMA distributes subcarriers between the users so they can transmit and receive at the same time within the same channel that is split into subchannels. The subcarriers are distributed among users depending on several factors like the location and the characteristics of propagation for each user, so its possible to increase the performance by selecting the best subcarriers suited for each user with less fading and interference [27]. OFDMA isn't used only in LTE, other systems such as WiMAX and the defunct UMB use OFDMA too, the only difference is that they use other factors like different subcarrier spacing, bandwidth, symbol length, and so on.

Another characteristic of the OFDMA used in LTE is that the subcarriers have a narrow spacing of just 15 Khz regardless of the total available bandwidth. Since the subcarriers are orthogonal to each other when the sampling occurs the other subcarriers will have zero value.

At the transmitter of OFDMA the modulated data source is sent to a serial-to-parallel converter and each output of the converter goes to each input of the IFFT block after mapping the symbols with the corresponding subcarriers. After all subcarriers are mapped a cyclic prefix extension is added which is the copy at the end of the symbol added to the beginning of the symbol. Instead of using a breaking in the transmission the cyclic prefix makes the OFDM symbol periodic allowing a discrete Fourier spectrum, enabling the use of IFFT at the transmitter and FFT at the receiver. As it was previous mentioned, the guard interval is designed to exceed the delay spread in the environment to prevent loss or orthogonality caused by the channel effects.

At the receiver the previously added guard interval must be removed. While the guard interval protects the symbols from ISI each subcarrier may be affected by different amplitude and phase changes. To account for those changes, pilot symbols are added to the OFDM symbols with proper placement in both time and frequency domain. To estimate the channel the receiver interpolates the effect of the channel to the different subcarriers from the grid which contains the time and frequency domain pilot symbols [28].

The receiver uses a frequency domain equalizer to revert the channel effects for each subcarrier. The equalizer multiplies each subcarrier based on the estimated channel frequency response adjusting the phase and amplitude each subcarrier has experienced. Also the OFDMA receiver needs to account for time and frequency synchronization. While time synchronization is obtained by correlation between the known data samples, typically the pilot symbols, and the actual received data, for frequency synchronization the frequency offset between the transmitter and the receiver is estimated by using an estimation of the frequency offset between the UE and the BS, where the UE locks to the frequency obtained from the base station.

If we consider the time domain, one of the main drawbacks of OFDMA transmission is that the signal envelope varies strongly because there are multiple sinusoidal waves with different frequencies with steps of 15 Khz leading to a gaussian distribution of different peak amplitude values. Since the maximum power efficiency is achieved when the amplifier operates at saturation point, low PAPR is needed to achieve higher efficiency [14]. While in BS power is available, for mobile terminals it means the battery is consumed faster. This is the main reason why SC-FDMA was chosen for the uplink instead of OFDMA.

3.2.1 E-UTRA Frame Structure

In FDD the downlink and uplink transmissions the frames are separated in the frequency domain while in TDD the frames are either allocated for the downlink or the uplink. The E-UTRA frame structure consists in 10 ms radio frames subdivided in ten 1 ms sub-frames.

Being the basic time unit given by the maximum FFT size of 2048 with 15 Khz spacing between the subcarriers $Ts = 1/(15000*2048)s$, the slot has a duration of $T_{slot} = 307200*T_s = 0.5ms$ where T_s is the sampling period. The combination of two of those slots is a subframe with the duration of 1 ms. The Fig. 3.4 below shows the constitution of a frame.

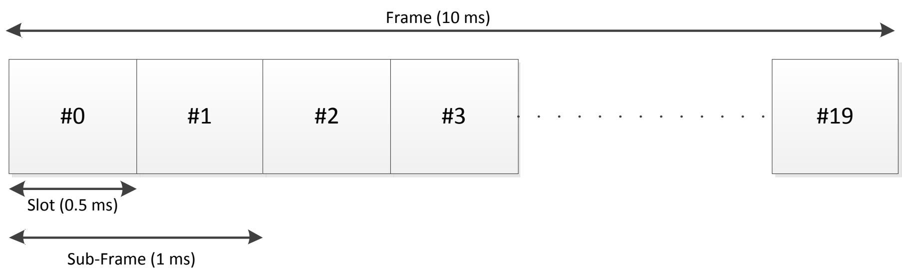

Figure 3.4: LTE Generic Frame Structure

The basic scheduling unit for both downlink and uplink is called Physical Resource Block (PRB) and its assigned by the base station (eNodeB) scheduler. The PRB may contain 6 OFDM symbols for the long CP or 7 OFDM symbols for the short CP in the time domain. Comparing the duration of resource block of short CP $(7*66.67+7*4.69) \simeq 500\mu s$ with long CP $(6*66.67+6*16.67) \simeq 500\mu s$ we will get equal duration which was chosen to preserve the 0.5ms slot timing [29]. This is represented in the Fig. 3.5.

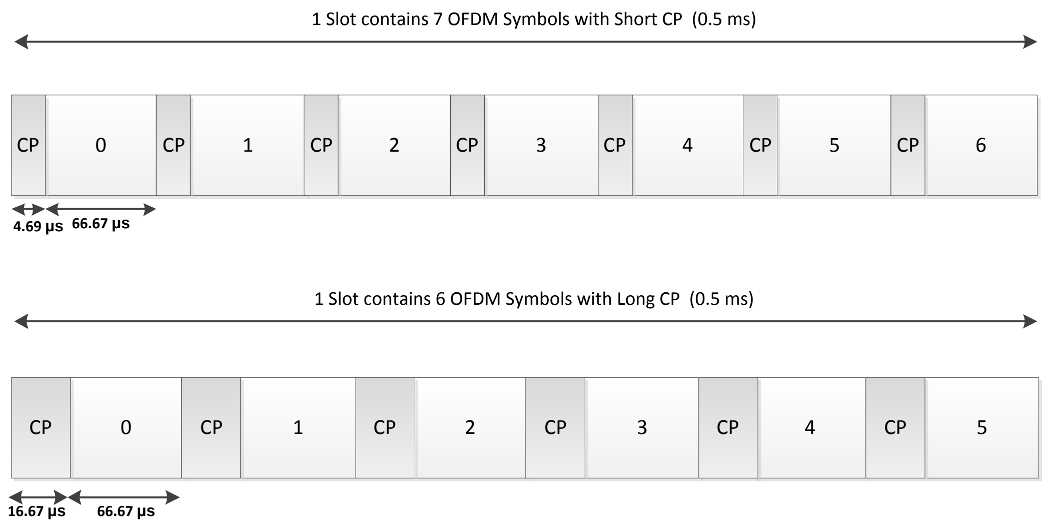

Figure 3.5: Short vs Long Cyclic Prefix

The allocation of subcarriers in the frequency domain in done in the PRBs with 12 consecutive sub-carriers of 15 Khz each, resulting in a minimum bandwidth of 180 Khz [29]. The transmitted signal is represented by a resource grid in Fig. 3.6 where each cell inside the grid is a resource element which represents a single carrier with the duration of 1 symbol period. Although in the frequency domain scheduling different sub-carriers could have different modulations, it is not practical and the the parameters such as modulation are fixed on the resource block basis [28].

Regarding the overall system bandwidth for LTE, it varies by defined parameters (from 1.25 Mhz to 20 Mhz), therefore the total number of subcarriers is variable.

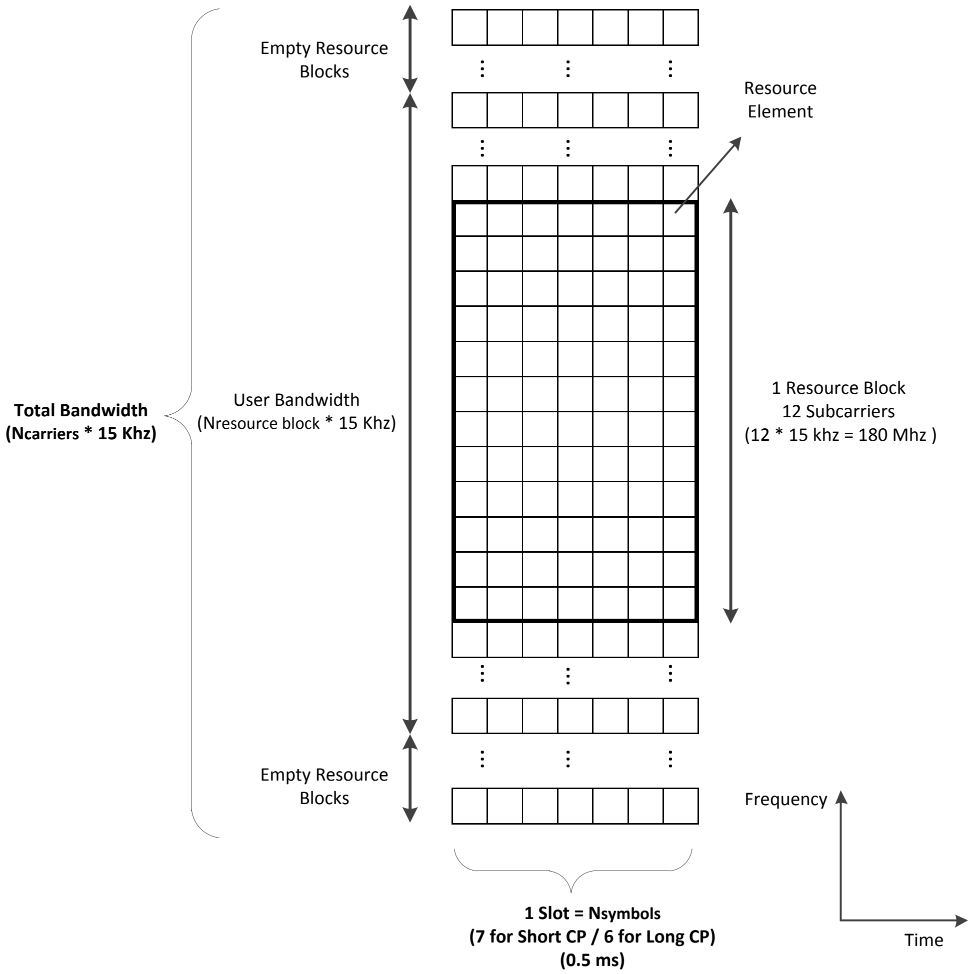

Figure 3.6: Resource Grid of LTE

Reference symbols, also known as pilot symbols, are inserted in the PRBs within the first OFDM symbol and the fifth for short CP or fourth for long CP for a LTE system with one antenna. There is also a frequency domain staggering of three subcarriers between the first and second reference symbols, thus there are four reference symbols within each reference block. The insertion of reference symbols is depicted in further detail in Fig. 3.7. As previously mentioned, the UE will interpolate over multiple reference symbols to estimate the channel.

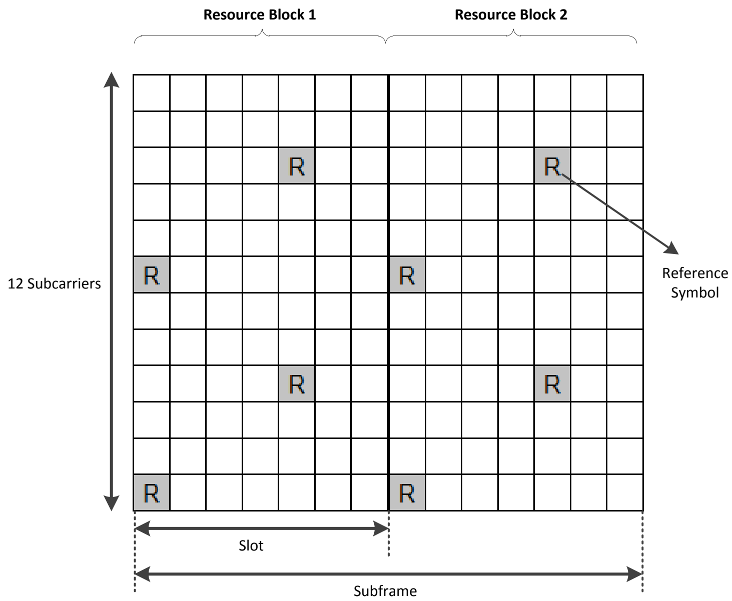

Figure 3.7: Position of Reference Symbols in a Resource Block

If two transmit antennas are used the reference symbols are inserted from each antenna, where the reference symbols on the second antenna are offset in the frequency domain by three subcarriers. Furthermore, when one antenna transmits a reference symbol in a time/frequency location the other doesn't transmit anything at that location [30].

The modulation parameters used in LTE Release 8 are shown in the table below [29]:

Table 3.1: Modulation Parameters of LTE Release 8

Modulation	**Downlink**	QPSK, 16-QAM, 64-QAM					
	Uplink	BPSK, QPSK, 16-QAM					
Bandwidth (Mhz)		1.4	3	5	10	15	20
Number of available PRBs		6	12	25	50	55	100
FFT Size		128	256	512	1024	1536	2048
Sampling Frequency (Mhz)		1.92	3.84	7.68	15.36	23.04	30.72
Slot duration		0.5 ms					
Sub-frame duration		1 ms					
Subcarrier Spacing		15 Khz					
PRB Bandwidth (Mhz)		180 Khz					
OFDM Symbol per Slot	**Short CP**	7					
	Long CP	6					
Cyclic Prefix Duration	**Short CP**	160 for l=0 (5.21 μs)					
		144 for l=1,2...5 (4.69 μs)					
	Long CP	512 (16.6μs)					
		1024 (33.33 μs)					

3.3 SC-FDMA

SC-FDMA is the uplink technique for multiple access of LTE. Unlike OFDMA that has high PAPR which requires expensive and inefficient power amplifiers, since SC-FDMA uses only one carrier for transmitting, its PAPR is lower which comes as an advantage when using battery powered UEs. SC-FDMA uses less power and therefore brings higher autonomy to UEs. On the other hand, SC-FDMA isn't used for the downlink because it requires additional complexity that lowers the performance of BSs where multiple users are processed simultaneously [31].

The main difference between OFDMA and SC-FDMA is that the subcarriers aren't independently modulated and different users are mapped into different subcarriers in the frequency domain but transmitted signal is sent using only one carrier, thus combining orthogonality between the users and the low PAPR of the single carrier modulation. Some of the architecture of OFDMA is shared in the SC-FDMA architecture.

SC-FDMA has 3 main key features, described below [32]:

- Lower power variations in the transmitted signal (low PAPR), associated to single carrier transmission
- Possibility to use equalization with low complexity but with high quality in the frequency domain
- Flexible distribution of the bandwidth.

3.3.1 Architecture of SC-FDMA

The transmitter of SC-FDMA aggregates the previously modulated symbols (in Binary Phase Shift Keying (BPSK), QPSK or 16-QAM modulation depending on the channel conditions for each terminal), in blocks of N symbols. In general the same modulation format is used for all subcarriers to keep the overhead small [14].

Each block of N symbols is sent for a FFT to spread the symbols in the frequency domain. In the frequency domain, the symbols are mapped in orthogonal subcarriers and afterwards the IFFT will convert them back to time domain so they can be sent, after the addiction of cyclic prefix. Finally the linear operation of pulse shaping is needed to reduce the energy outside the band to avoid spectrum regrowth, just before the signal is sent to a Digital to Analog Converter (DAC) and Radio Frequency (RF) conversion.

The blocks of the architecture of SC-FDMA are represented in Fig. 3.8 below:

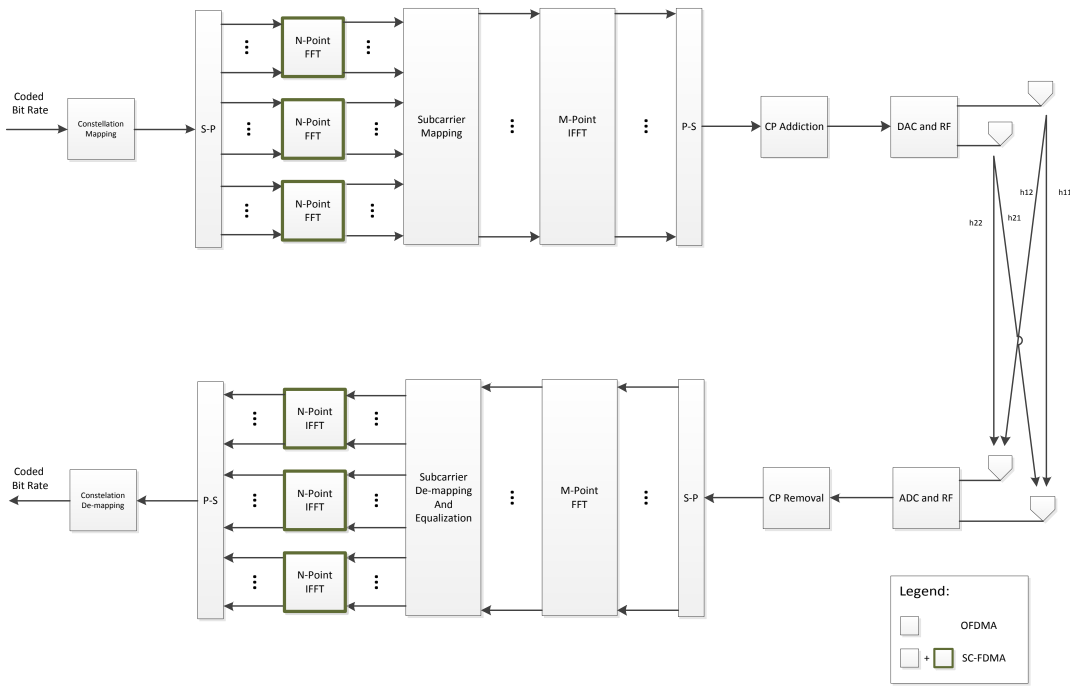

Figure 3.8: Architecture of SC-FDMA vs OFDMA

The blocks used for the transmission of a SC-FDMA symbol are the following:

- Constellation Mapping
- S-P Converter
- N-point FFT
- Subcarrier mapping
- M-point IFFT
- P-S Converter
- Cyclic Prefix Addiction and Pulse Shaping
- RF block

If you look closely you will notice that the main difference between the OFDMA and SC-FDMA in the architecture is the existence of the N-point FFT block in the transmitter which performs the frequency spreading of the data blocks. At the receiver the N-point IFFT block reverts the effect of the transmitter's N-point FFT.

At the receiver the process is the opposite of the transmitter, after the down conversion and the Analog to Digital Converter (ADC) the cyclic prefix is removed which may contain ISI of the previous symbol. The FFT block moves the subcarriers to frequency domain so that frequency domain equalization can be performed for channel correction. Some examples of used equalizers are the mean square error equalization (MMSE), Decision Feedback Equalization (DFE) and turbo equalization. The already equalized symbols are converted back to the time domain by the IFFT blocks for detection and decoding.

3.3.2 Subcarrier Mapping

There are two ways of subcarrier mapping in SC-FDMA, the localized and distributed mapping, which are depicted in Fig. 3.9. With localized mapping each user use a set of adjacent subcarriers to transmit its data. It is possible to get frequency selective gain because there is the possibility of scheduling the uplink frequencies so that the mobile terminal gets favorable propagation conditions in frequency selective channel conditions. However this frequency mapping scheme cancels frequency diversity. In the distributed mapping the subcarriers used by the user are spread over the entire bandwidth. This frequency mapping scheme is more robust with respect to frequency selective fading and offers additional frequency diversity gain [14]. However better synchronization is required in terms of frequency for the distributed mapping.

Distributed mapping has inherently lower PAPR than localized mapping. However, pulse shaping is necessary to curtail out-of-band spectrum components prior to radio transmission [33].

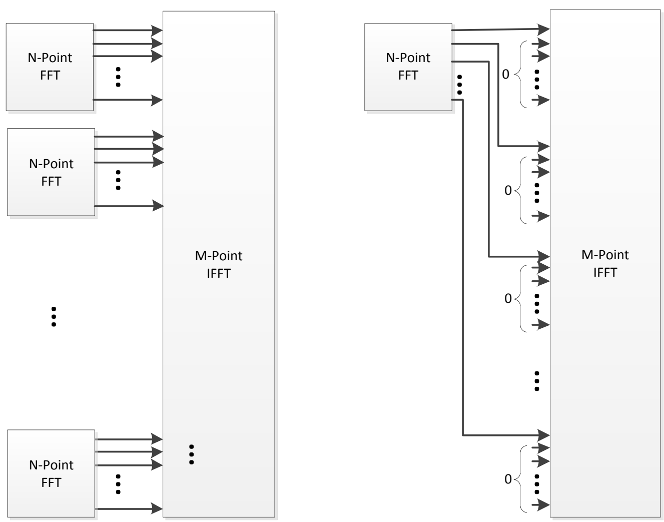

Figure 3.9: Localized vs Distributed Mapping

Chapter 4

MIMO Systems

The Release 7 of LTE specified MIMO communications, where multiple antennas are used at the receiver and/or the transmitter. MIMO systems improve the capacity and throughput of the wireless link when compared to Single Input Multiple Output (SISO) systems by using diversity techniques and transmitting multiple streams of data in parallel. MIMO systems can also reduce the interference between the users by using smart antennas. The physical layer was the most affected by the introduction of MIMO.

Beside MIMO systems that uses multiple antennas at the transmission and the reception, other schemes are used in LTE. While SISO is already used in traditional radio systems where a single antenna transmits to one receiving antenna, Multiple Input Single Output (MISO) have multiple antennas at the transmitter side but just one at the receiver. Finally the Single Input Multiple Output (SIMO) is characterized by one antenna transmitting to several antennas at the receiver's side. The different schemes of MIMO are represented in the Fig. 4.1 below:

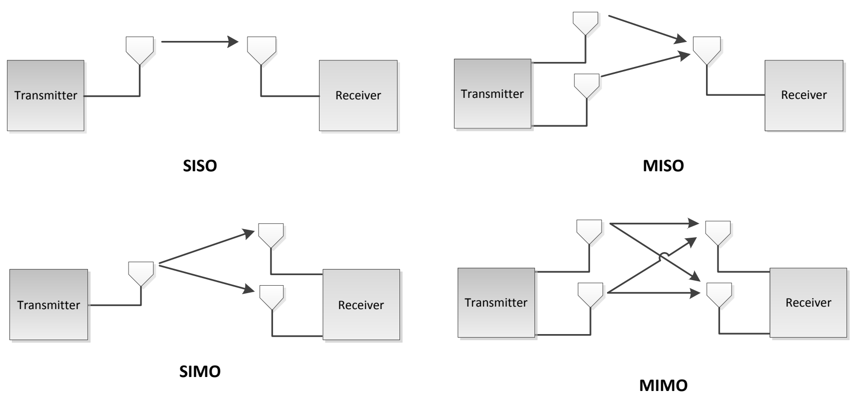

Figure 4.1: MIMO Schemes

In a transmission channel due to the existence of obstacles between the transmitting and the receiving antennas the effects of reflection, diffusion or refraction may occur. Those effects will generate different paths for the transmitted data which are received with temporal variation of amplitude and phase. The effectiveness of diversity schemes lies in the fact that at the receiver we must provide independent samples of the basic signal that was transmitted. Since the probability of two or more relevant parts of the signal undergoing deep fades will be very small we will get a diversity gain . The receiver must combine the received diversified waveforms to maximize the resulting signal quality. Transmit diversity is easier to implement at the BS and consists in introducing controlled redundancies at the transmitter, which can be then exploited by appropriate signal processing techniques at the receiver [34].

MIMO Systems are the only configurations capable of spatial multiplexing where the transmitted data is multiplexed by multiple antennas. Those multiple antennas transmit simultaneously several multiplexed data streams which are received by the multiple receiver's antennas to be de-multiplexed so that the original stream can be recovered. With the same transmitting power and bandwidth the transmission rates increase proportionally with the number of transmit-receive antenna pairs [35].

Array gain is obtained by a coherent combining of the signal from multiple antenna elements at the transmitter and/or receiver which increase the average SNR. To obtain array gain channel knowledge is necessary [36], and it is made available via channel estimation. If the transmitter does not have any information about the channel, we have a open-loop system. The system is closed loop if the transmitter receives information about the channel conditions, in a return channel, from the receiver to improve the performance. In a closed-loop system, the array gain is achieved by processing at the transmitter and the receiver [37]. The main drawback of closed loop spatial multiplexing is that it requires the channel conditions at the receiver to be steady, so when the receiver is in high mobility closed loop spatial multiplexing is not suitable.

4.1 MIMO Scenarios

In MIMO systems we have two different scenarios, one when the Signal to Interference-plus-Noise Ratio (SINR) is low and the another when the SINR is high. For the first scenario the aim of MIMO systems is to improve the SINR, while for the second one is to share the SINR.

In the scenarios with low SINR, beamforming (generating a narrow beam in the direction of the receiver) and transmit diversity techniques are used, for example the Alamouti space-time block code or space-frequency block codes. Although transmit diversity does not improve the average SINR, it reduces the variations in the SINR experienced by the receiver, therefore we can use higher order modulations and increase the coding rate, thus higher link throughput is achieved. [38]

On the other hand, when we have scenarios with high SNR if the SINR level becomes high enough, the SINR vs throughput curve saturates, i.e. when we increase the SINR further we won't get a considerable amount of throughput. To prevent such effect, multiple antennas are used at the transmitter and at the receiver of the radio link. This allows spatial multiplexing where different information bits are sent from each antenna, sharing overall SINR.

Considering a MIMO 2x2 system, we can model it using two parallel channels where each one has a specific SINR neglecting the interference between the streams. If we keep the total transmitted power the same, that is, to reduce the power used in each stream by 3 dB and then if we consider the same amount of noise in both streams, the SINR value for each stream will be half the SINR value of a single stream transmission (in a SISO system) using the same total power. This means that we can send two different data streams using the same power and when we have sufficient high SINRs we can double the throughput. However we have to take account of the interference between the streams because it's orthogonality depends a lot on the actual MIMO channel realization.

In Multi User MIMO, different users transmit their respective information simultaneously to the a BS equipped with multiple antennas. The BS will then separate the signals from the different users. Multi User MIMO increases the capacity of the cell and is often used when the LTE network is under high load [38].

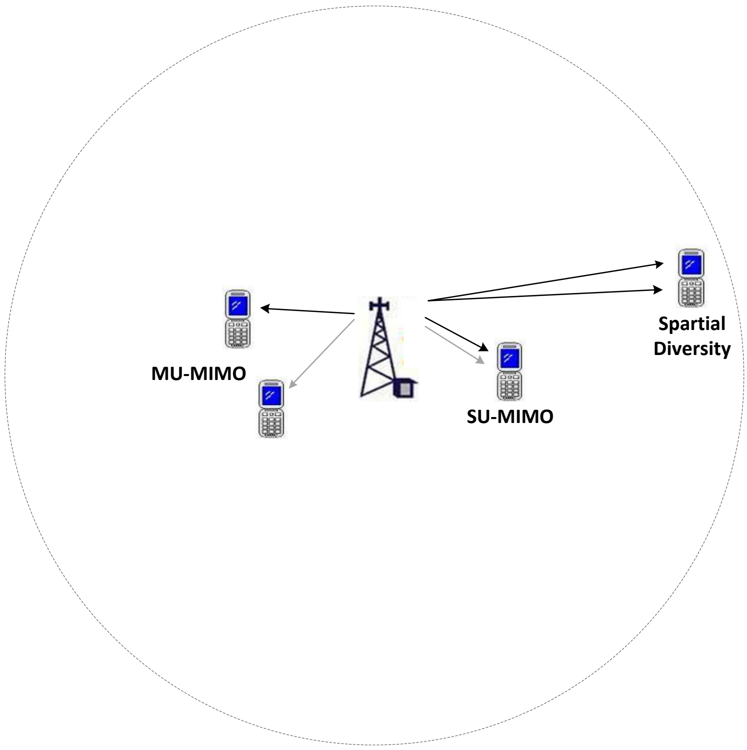

Figure 4.2: MIMO Scenarios

In cellular networks we have interference from other cells that is higher on the edge than in the center and the received energy behaves opposite way. So in the edge of the cells we can use beamforming and spatial diversity techniques to improve the SINR and in the middle of the cell where we get high SINR, the SINR is shared by means of spatial multiplexing. This concludes that improving SINR techniques are used to give coverage while spatial multiplexing techniques are used for sharing SINR to improve peak rate.

Now lets present transmit and receiving diversity techniques.

4.2 Transmit Diversity

4.2.1 Space Time Coding and The Alamouti Concept

Space Time Coding (STC) uses transmit diversity to takes advantage of the spatial diversity present in MIMO systems where different copies of the transmitted signal are sent thought separate channels with different fading characteristics to get spatial diversity gain. The obtained gain is proportional to the number of antennas at the transmitter and the receiver [35]. Since after channel coding, the data is sent in different periods of time, a coding gain is also obtained in complement of the spatial diversity gain. This technique offers low complexity in channel coding/decoding.

The Space Time Block Coding (STBC) is represented by a two-dimensional space-time matrix. The symbols after being mapped are transmitted using the space-time matrix where the columns represent the transmitting antennas and the lines the period of time where the symbol will be transmitted. Considering that we have n antennas and T time slots we have the following space-time matrix:

$$
\begin{array}{cc}
 & \xrightarrow{\text{Transmitting Antennas}} \\
\begin{array}{c}\text{Time} \\ \text{Slots}\end{array}\Big\downarrow & \begin{bmatrix} S_{1,1} & S_{1,2} & \dots & S_{1,n} \\ S_{2,1} & S_{2,2} & \dots & S_{2,n} \\ \vdots & \vdots & \ddots & \vdots \\ S_{T,1} & S_{T,2} & \dots & S_{T,n} \end{bmatrix}
\end{array}
$$

Figure 4.3: Space-Time Block Code

Alamouti introduced a simple STC which contains a method for achieving spatial diversity with two antennas allowing the same data rate as on a single antenna [39]. Moreover, it was assumed that the channel response is flat fading, for each transmitting antenna the channel experienced is different from the other antenna and despite of the channel varying in time, it is constant over two time slots.

The scheme proposed by Alamouti follows: considering that we have MISO system with two transmitting antennas and one receiver, a transmission sequence is sent $\{S_1, S_2\}$ where each subscript are time slots that are grouped in groups of two. In the first time slot the first antenna will send the symbol S_1 and the second one will send S_2. In the second time slot it will be sent replicas of the first symbols but with some differences, the symbol S_1 is conjugated (S_1^*) and will be sent by the second antenna and the symbol S_2 is conjugated and inverted $-(S_2^*)$ and will be sent by the first antenna. Since 2 slots are used to send 2 symbols there is no change of data rate. This is represented in the following table:

Table 4.1: Alamouti Space-Time Block Code for MISO 2x1

	antenna 1	antenna 2
time slot n	S_n	S_{n+1}
time slot $n+1$	$-S_{n+1}^*$	S_n*

where the STBC used is given by the following matrix:

$$\mathbf{S} = \begin{bmatrix} S_n & S_{n+1} \\ -S_{n+1}^* & S_n^* \end{bmatrix} \tag{4.1}$$

Considering H_1 the channel response between the first transmitting antenna and the receiver, H_2 the channel response between the second transmitting antenna and the receiver which both form the vector $H = [H_1\ H_2]^T$. The representation of the Alamouti scheme is shown in the Fig. 4.4 below:

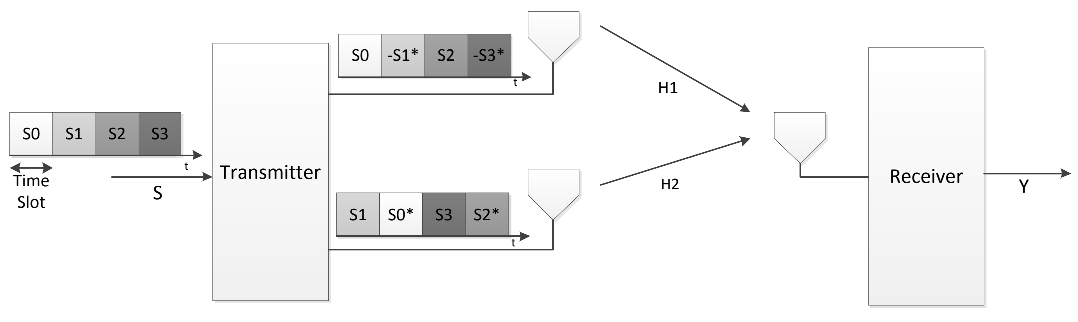

Figure 4.4: Alamouti Scheme for MISO 2x1 for 4 transmitting symbols

The received signal is given by

$$\mathbf{Y} = \frac{1}{\sqrt{2}}\mathbf{SH} + \mathbf{N} \tag{4.2}$$

considering that N is the channel noise vector at the receiver. That equation can be decomposed in:

$$
\begin{cases}
Y_n = \frac{1}{\sqrt{2}} H_{1,n} S_n + \frac{1}{\sqrt{2}} H_{2,n} S_{n+1} + N_n \\
\\
Y_{n+1} = \ - \frac{1}{\sqrt{2}} H_{1,n+1} S_{n+1}^* + \frac{1}{\sqrt{2}} H_{2,n+1} S_n^* + N_{n+1}
\end{cases}
\tag{4.3}
$$

Considering that the channel response doesn't varies between 2 consecutive time slots, the equation above is equivalent to

$$
\begin{cases}
Y_n = \frac{1}{\sqrt{2}} H_1 S_n + \frac{1}{\sqrt{2}} H_2 S_{n+1} + N_n \\
\\
Y_{n+1}^* = \frac{1}{\sqrt{2}} H_2^* S_n - \frac{1}{\sqrt{2}} H_1^* S_{n+1} + N_{n+1}^*
\end{cases}
\tag{4.4}
$$

In the matrix form, the previous equation is

$$
\begin{pmatrix} Y_n \\ Y_{n+1}^* \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} H_1 & H_2 \\ H_2^* & -H_1^* \end{pmatrix} \begin{pmatrix} S_n \\ S_{n+1} \end{pmatrix} + \begin{pmatrix} N_n \\ N_{n+1}^* \end{pmatrix}
\tag{4.5}
$$

The short notation is given by:

$$
\mathbf{Y} = \frac{1}{\sqrt{2}} \mathbf{H}_v \mathbf{S} + \tilde{\mathbf{N}}
\tag{4.6}
$$

We conclude that the resulting virtual channel matrix rows and columns $\mathbf{H}_v$ are orthogonal to each other.

$$
\mathbf{H}_v^H \mathbf{H}_v = \mathbf{H}_v \mathbf{H}_v^H = (H_1^2 + H_2^2)\mathbf{I}_2 = \mathbf{H}^2 \mathbf{I}_2
\tag{4.7}
$$

where $\mathbf{I}_2$ is the (2x2) identity matrix and $(.)^H$ the hermitian operation.

4.2.2 Space Frequency Coding

Space-Frequency Block Code (SFBC) is a variant of the Alamouti STBC. Space-Frequency Coding (SFC) instead of being applied on adjacent time symbols, it is applied on subjacent subcarriers where it is assumed that the channel is constant between those subcarriers. SFBC achieves similar diversity gain as STBC-OFDM in slow fading but SFBC performs better in fast fading channels [40]. Moreover in SFBC there is almost no decoding delay which is an important advantage compared to STBCs. To exemplify this spatial diversity scheme we considered a system with 2 antennas transmitting that sends a code block in two adjacent subcarriers.

The transmitted signal is:

Table 4.2: Alamouti Space-Frequency Block Code for MISO 2x1

	antenna 1	antenna 2
subcarrier n	S_n	$-S_{n+1}*$
subcarrier $n+1$	S_{n+1}	S_n*

Note that the original data information is sent without modification in the first antenna to make it compatible with systems without SFBC where the second antenna is not used.

At the receiver the signals on the subcarriers n and $n+1$ after removing the guard interval and the FFT are:

$$\begin{cases} Y_n = \frac{1}{\sqrt{2}} H_{1,n} S_n - \frac{1}{\sqrt{2}} H_{2,n} S_{n+1}^* + N_n \\ \\ Y_{n+1} = \frac{1}{\sqrt{2}} H_{1,n+1} S_{n+1} + \frac{1}{\sqrt{2}} H_{2,n+1} S_n^* + N_{n+1} \end{cases} \tag{4.8}$$

where $H_{k,n}$ is the flat fading coefficient of the nth channel for transmit antenna k and N_n the noise at the subcarrier n. Since OFDM system are designated so that the fading in each subcarrier is flat we can assume that $H_{k,n} = H_{k,n+1}$.

The estimated symbols are given assuming the previous equality by:

$$\begin{cases} \hat{S}_n = \frac{1}{\sqrt{2}} H_1^* Y_n + \frac{1}{\sqrt{2}} H_2 Y_{n+1}^* \\ \\ \hat{S}_{n+1} = -\frac{1}{\sqrt{2}} H_2 Y_n^* + \frac{1}{\sqrt{2}} H_1^* Y_{n+1} \end{cases} \tag{4.9}$$

Resulting in

$$\begin{cases} \hat{S}_n = \frac{1}{2}(|H_1|^2 + |H_2|^2) S_n + \frac{1}{\sqrt{2}} H_1^* N_n + \frac{1}{\sqrt{2}} H_2 N_{n+1}^* \\ \\ \hat{S}_{n+1} = \frac{1}{2}(|H_1|^2 + |H_2|^2) S_{n+1} + \frac{1}{\sqrt{2}} H_2 N_n^* + \frac{1}{\sqrt{2}} H_1^* N_{n+1} \end{cases} \tag{4.10}$$

where we conclude that like in the STBC case the received matrix lines and columns are orthogonal to each other and the SNR is given by

$$SNR = \frac{1}{2}\frac{(|H_1|^2 + |H_2|^2)}{\sigma^2} \tag{4.11}$$

that means that a diversity order of 2 can be achieved with the SFBC.

4.3 Receive Diversity

Receive diversity is easier to achieve than transmit diversity since it doesn't require special modulation and coding schemes. A list of the most known receive diversity schemes follow:

- **Selection Combining** - The receiver selects the antenna with the highest received signal, ignoring the received signal from the other antennas.
- **Switched combining** - A threshold of SNR is set at the receiver. If the received signal at a determined antenna is above the threshold then it is selected until it drops again under that threshold.

  Equalizers in frequency domain are also used:

- **Equal Gain Combining** - The received signals from the multiple receiving antennas are summed up coherently, compensating the phase rotation caused by the channel.
- **Maximum Ratio Combining** - The received signals from the different antennas are weighted individually depending on the reliability of the signal in each antenna and then the phase distortion is compensated by calculating the complex conjugate of the frequency response of the channel. Afterwards the signals are aligned and combined.
- **Zero Force Combining** - This equalizer forces ISI to zero restoring the orthogonality between users by inverting the channel frequency response. However, this equalizer boosts the noise when the frequency response of the channel has deep notches.
- **Minimum Mean Square Error Combining** - This equalizer minimizes the mean squared error between the transmitted signal and the estimated signal. This equalizer is equal to ZFC, if the $SNR \to \infty$.

From the four equalizers, Equal Gain Combining and Maximum Ratio Combining are only suitable for the downlink when the OFDMA scheme is used because they can't remove the ISI present in uplink's SC-FDMA which is caused by the channel frequency response and noise. For the uplink the ZF and MMSE equalizers can be used to remove such ISI.

4.3.1 Linear Frequency Domain Equalizers

In the uplink scheme of LTE frequency equalization is used instead of time domain equalization because it offers huge complexity savings as compared with equivalent time domain approaches resulting in less processing requirements [41]. Moreover, the complexity of time domain equalization increases with the number of symbols that are spanned by the channel impulse response.

Considering a SC-FDMA system where data blocks with N modulation symbols per block $\{s_n : n = 0, 1, ..., N-1\}$ are converted into the frequency domain by a DFT resulting in $\{S_k : k = 0, 1, ..., N-1\}$, the received signal is given by

$$Y_k = S_k H_k + N_k \tag{4.12}$$

If a ZF equalizer is used then $F_k = 1/H_k = H_k^*/|H_k|^2$, where

$$\tilde{S}_k = F_k Y_k = \frac{H_k^*}{|H_k|^2} Y_k. \tag{4.13}$$

If we have R receiving antennas then

$$Y_k^{(r)} = S_k H_k^{(r)} + N_k^{(r)} \tag{4.14}$$

and the corresponding estimated symbols is equal

$$\tilde{S}_k = \sum_{r=1}^{R} F_k^{(r)} Y_k^{(r)} \tag{4.15}$$

where $F_k^{(r)}$ is a Frequency Domain Equalizer (FDE) coefficient for each receiving antenna R and it is calculated by

$$F_k^{(r)} = \frac{H_k^{(r)*}}{\sum\limits_{r'=1}^{R} |H_k^{r'}|^2} \tag{4.16}$$

This results after replacing (4.14) and (4.16) in (4.15) in

$$\tilde{S}_k = S_k + \frac{\sum\limits_{r=1}^{R} H_k^{(r)*}}{\sum\limits_{r'=1}^{R} |H_k^{r'}|^2} N_k^{(r)} \tag{4.17}$$

The ZF will remove the ISI but in a typical frequency-selective channel, there are deep notches in the channel frequency response and the ZF equalizer will increase the noise when attempting to invert the channel impulse response. To prevent such thing to happen MMSE equalizers are used.

Since the mean square error is given by

$$\Theta(k) = \frac{1}{N^2}\sum_{k=1}^{N-1}\Theta_k \tag{4.18}$$

where

$$\Theta_k = E[|\tilde{S}_k - S_k|^2] = E[|Y_k F_k - S_k|^2]. \tag{4.19}$$

To obtain the optimized coefficients of the FDE we need to minimize $\Theta(k)$ in order to F_k. For each k symbol this results in

$$F_k = \frac{H_k^*}{\beta + |H_k|^2} \tag{4.20}$$

where the β is the inverse of the SNR and is given by

$$\beta = \frac{\sigma_N^2}{\sigma_S^2} \tag{4.21}$$

and it is composed by the noise variance is given by

$$\sigma_N^2 = \frac{E[|N_k|^2]}{2} \tag{4.22}$$

and the variance of the symbol

$$\sigma_S^2 = \frac{E[|S_k|^2]}{2} \tag{4.23}$$

If we have R receiving antennas then we have a vector $F_k^{(r)}$ where $(r = 1, ..., R)$ and if we admit that the channel has equal noise level conditions for every branch then $\sigma_N^{(1)} = \sigma_N^{(2)} = ... = \sigma_N^{(R)}$ we have

$$F_k^{(r)} = \frac{H_k^{(r)*}}{\beta + \sum_{r'=1}^{R} |H_k^{(r')}|^2}. \tag{4.24}$$

If we optimize the FDE receiver using the MMSE equalizer it won't attempt to fully invert the channel when it has a deep fading because the noise dependent term β avoids the noise enhancements when the frequency response has very low values, therefore it won't boost the noise and we get better performance. If we increase the number of receiving antennas, the probability of having deep notches simultaneously at every path diminishes considerably. After combining the outputs of every diversity branch we are able to compensate the channel effects with better efficiency. Therefore if we increase the diversity order by adding more antennas we reduce equalization requirements because the overall channel gets closer to a frequency flat channel and that means that the ZF equalizer will give the same results as the MMSE equalizer using less complexity [42].

Chapter 5

Iterative Frequency-Domain Equalizer for LTE

Time-Domain Decision Feedback Equalizer (TD-DFE) significantly outperform linear equalizers in dispersive channels due to cancelling the ISI though a feedback filter [43]. The major drawback of TD-DFE equalizers is that complexity increases if the channel is severely dispersive because more signal processing is needed. To achieve a good compromise between the performance of TD-DFE and simplicity of Frequency Domain Equalization, an hybrid time-frequency Iterative Block Decision Feedback Equalizer (IB-DFE) was proposed [44] with the forward filter in the frequency domain and the feedback filter in the time domain. Despite of achieving better performance, it also suffers from error propagation. Finally a IB-DFE where both the feedback and forward parts implemented in the frequency domain was proposed for SC transmission [45] and later extended to diversity scenarios , i.e., for systems employing multiple antennas at both terminals.

Contrarily to the 3G cellular systems, the LTE supports the use of MIMO technology. For the uplink and for the single antenna UEs the cell level maximum data rate can be increased by the allocation of some UEs with the same frequency resources. Thus the transmission in the BS is treated like a MIMO transmission as shown in Fig. 5.1, and the data stream separated with MIMO receiver processing. This kind of "virtual" or "Multi-user" MIMO is supported in LTE Release 8. However, since the signals associated to different users are affected by different propagation channels the use of linear frequency domain multi-user detection is not the most efficient, mainly for high time dispersive channels [11]. A more sophisticated detectors are required. Therefore, in this work we extend the IB-DFE, designed in frequency domain for single carriers systems, to the LTE virtual MIMO scenario to efficiently separate the UEs that share the same frequency resource.

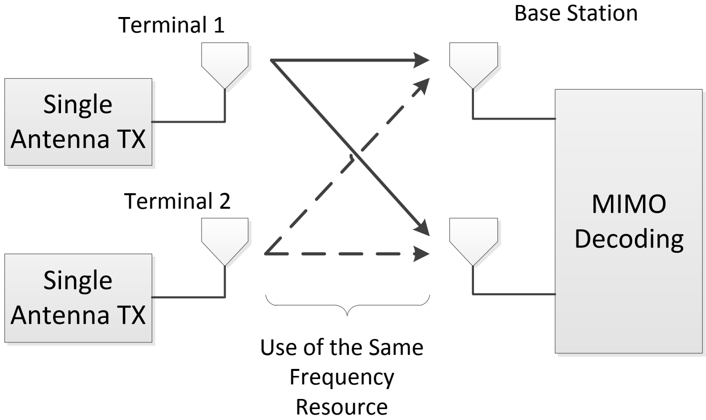

Figure 5.1: Multi User MIMO Principle [28]

5.1 System Characterization

Lets consider a MIMO system where at the transmitter side we have different users (UEs) with one transmitting antenna each and at reception we have a BS with R receiving antennas, all sharing the same frequency band. In Section 5.3, the SC-FDMA simulation platform implemented will be presented in detail. It is assumed that the transmitting users $\{p = 1, 2, ..., P\}$ transmit data blocks with the same size $\{s_{n,p}; n = 0, 1...N-1\}$ and similar date rate where $s_{n,p}$ is a symbol mapped using a known constellation like QPSK or 16-QAM. In the frequency domain, the data block transmitted by the pth user is $\{S_{k,p}; k = 0, 1...N-1\}$, that is the DFT operation of the transmitted data block by the same UE. The system model considered is represented in the Fig. 5.2.

The received signal at each rth antenna of the BS in the frequency domain, and assuming that the CP is long enough, is

$$Y_k^{(r)} = \sum_{p=1}^{P} S_{k,p} H_{k,p}^{eq(r)} + N_k^{(r)} \tag{5.1}$$

,

with $H_{k,p}^{eq(r)} = \beta_p H_{k,p}^{(r)}$ which represents the overall channel frequency response between the pth UE and the rth antenna of the BS at the kth subcarrier and β_p is the long-term channel power gain between the user p and BS.

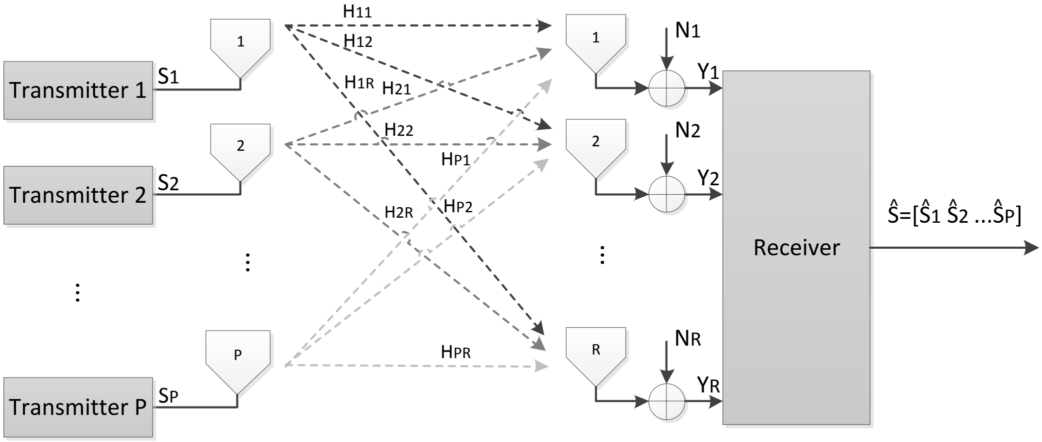

Figure 5.2: System Characterization

Considering the matrix notation, 5.2 can be re-written as

$$\mathbf{Y}_k = \mathbf{H}_k^T \mathbf{S}_k + \mathbf{N}_k \tag{5.2}$$

$\mathbf{Y}_k$ is a R column vector where each line corresponds to the received antenna signal at the BS and is represented by:

$$\mathbf{Y}_k = \begin{bmatrix} Y_k^{(1)} \\ Y_k^{(2)} \\ \vdots \\ Y_k^{(R)} \end{bmatrix}, \tag{5.3}$$

$\mathbf{S}_k$ is a P column vector where each element represents the transmitted data in the frequency domain, for each user at the subcarrier k,

$$\mathbf{S}_k = \begin{bmatrix} S_k^{(1)} \\ S_k^{(2)} \\ \vdots \\ S_k^{(P)} \end{bmatrix}, \tag{5.4}$$

$\mathbf{N}_k$ is a R column vector containing the noise samples received at each antenna,

$$\mathbf{N}_k = \begin{bmatrix} N_k^{(1)} \\ N_k^{(2)} \\ \vdots \\ N_k^{(R)} \end{bmatrix}, \tag{5.5}$$

and $\mathbf{H}_k$ is a PxR matrix containing the channel frequency response, defined by

$$\mathbf{H}_k = \begin{bmatrix} H_{k,1}^{eq(1)} & H_{k,1}^{eq(2)} & \dots & H_{k,1}^{eq(R)} \\ H_{k,2}^{eq(1)} & H_{k,2}^{eq(2)} & \dots & H_{k,2}^{eq(R)} \\ \vdots & \vdots & \ddots & \vdots \\ H_{k,p}^{eq(1)} & H_{k,p}^{eq(2)} & \dots & H_{k,p}^{eq(R)} \end{bmatrix}. \tag{5.6}$$

Now lets consider an IB-DFE receiver where each iteration consists of P detection stages, one for each user. When the user P is being detected, the residual ISI of that user and the interference of the previously detected users is cancelled. The structure used for the detection of the pth user at the ith iteration is represented in the Fig. 5.3 below:

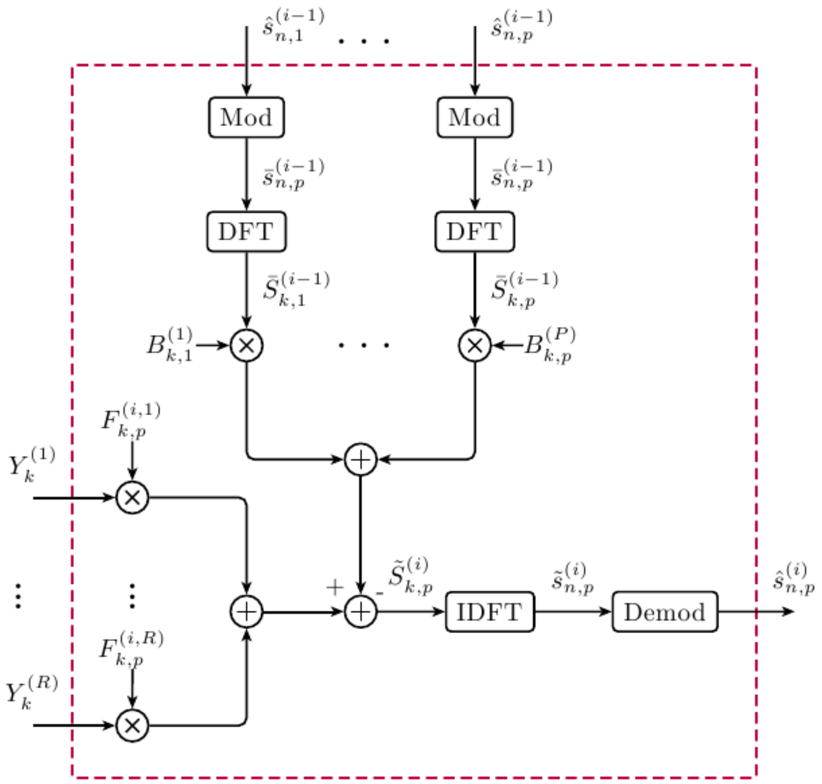

Figure 5.3: Detail of the IB-DFE for the pth user

The forward filter $\{F_{k,p}; k = 0, ..., N-1\}$ is designed to minimize both the ISI and interference that couldn't be removed by the feedback filter $\{B^{p'}_{k,p}; p' = 0, ..., P\}$, due to decision errors in the previous decision steps. To obtain the estimate of the transmitted block by the pth user the symbols represented in the time domain after the IDFT $\{\tilde{s}_{n,p}; n = 0, ..., N-1\}$, are passed through a hard-decision device. The estimated symbols after the hard decision are represented by $\{\hat{s}_{n,p}; n = 0, ..., N-1\}$.

To separate the users, at each iteration all UEs are simultaneously detected in parallel at each iteration forming a Parallel Interference Cancellation (PIC) receiver, as shown in the Fig. 5.4. Note that in the first iteration the system is reduced to a linear FDE since there is no feedback information.

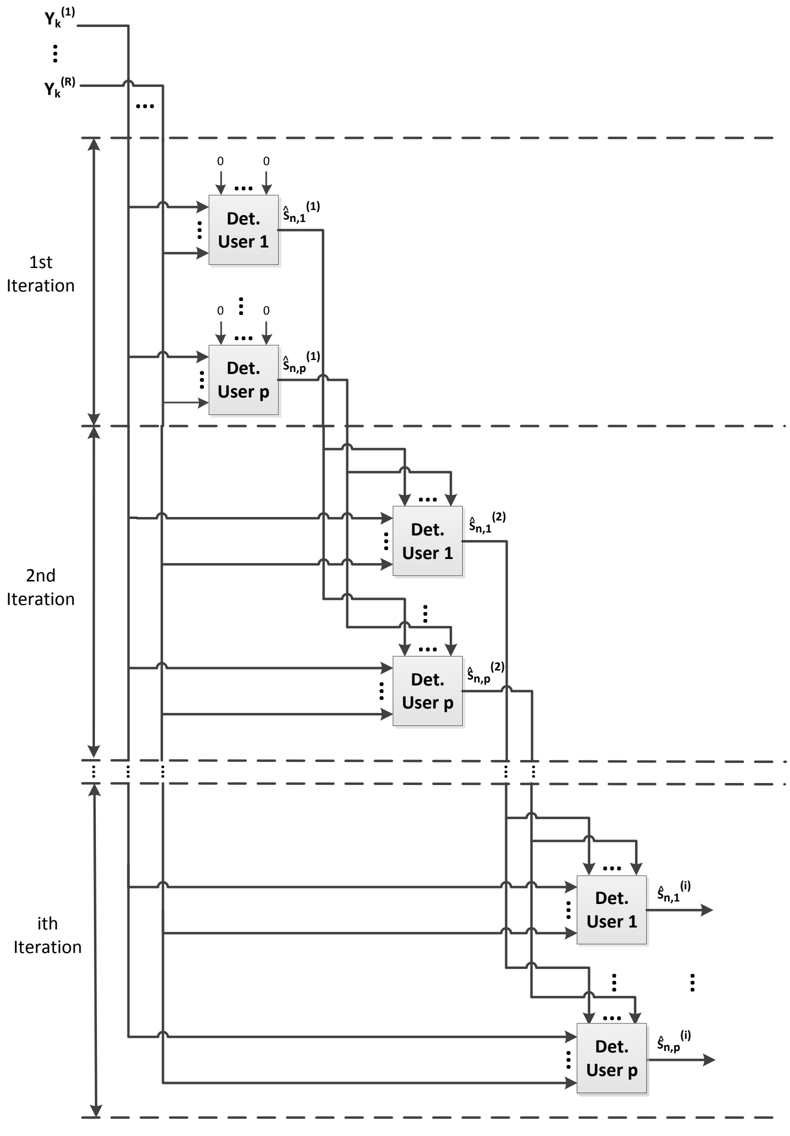

Figure 5.4: PIC IB-DFE Receiver

From Fig. 5.4 we can see that the result samples $\tilde{S}_k$ for each iteration are given by

$$\tilde{S}_{k,p}^{(i)} = \sum_{r=1}^{R} F_k^{(i,r)} Y_k^{(r)} - \sum_{p=1}^{P} B_k^{(i,p)} \bar{S}_{k,p}^{(i-1)} \tag{5.7}$$

In the matrix format the previous equation is given by,

$$\tilde{\mathbf{S}}_k^{(i)} = \mathbf{F}_k^{(i)T} \mathbf{Y}_k - \mathbf{B}_k^{(i)T} \bar{\mathbf{S}}_k^{(i-1)} \tag{5.8}$$

where $\bar{\mathbf{S}}_k$ is the vector with the soft decisions of $\mathbf{S}_k$ from the previous iteration, $\mathbf{F}_k$ is a RxP matrix containing the forward coefficients, defined by

$$\mathbf{F}_k^{(i)} = \begin{bmatrix} F_{k,1}^{(i,1)} & F_{k,2}^{(i,1)} & \dots & F_{k,p}^{(i,1)} \\ F_{k,1}^{(i,2)} & F_{k,2}^{(i,2)} & \dots & F_{k,p}^{(i,2)} \\ \vdots & \vdots & \ddots & \vdots \\ F_{k,1}^{(i,R)} & F_{k,2}^{(i,R)} & \dots & F_{k,p}^{(i,R)} \end{bmatrix}, \tag{5.9}$$

and $\mathbf{B}_k^{(i)}$ the PxP feedback coefficients matrix given by

$$\mathbf{B}_k^{(i)} = \begin{bmatrix} B_{k,1}^{(i,1)} & \dots & B_{k,p}^{(i,1)} \\ \vdots & & \vdots \\ B_{k,1}^{(i,P)} & \dots & B_{k,p}^{(i,P)} \end{bmatrix}. \tag{5.10}$$

Due to decision errors, we have $\bar{s}_{n,p} \neq s_{n,p}$ and consequently $\bar{S}_{k,p} \neq S_{k,p}$. To simplify the computation of the receiver coefficients, it is assumed that [42]

$$\bar{\mathbf{S}}_k^{(i-1)} = \mathbf{P}^{(i-1)} \mathbf{S}_k + \mathbf{\Delta}_k = diag(\rho_1, \rho_2 .., \rho_p) \mathbf{S}_k + \mathbf{\Delta}_k \tag{5.11}$$

where $\mathbf{\Delta}_k$ is a mean zero error vector uncorrelated with $\mathbf{P}^{(i-1)}$ representing the feedback noise, and ρ_p is the correlation coefficient of the pth user and its given by:

$$\rho_p = \frac{E[\bar{s}_{n,p} s_{n,p}^*]}{E[|s_{n,p}|^2]} = \frac{E[\bar{S}_{k,p} S_{k,p}^*]}{E[|S_{k,p}|^2]}, \tag{5.12}$$

and represents a measure of the reliability of the estimates associated to the ith iteration that can be obtained as described in [42].

The forward and backward matrices, $\mathbf{F}_k$ and $\mathbf{B}_k$, respectively are chosen to maximize the SINR for each user, at a particular iteration, given by

$$\mathbf{F}_k^{(i)} = \left[\mathbf{H}_k^H(\mathbf{I}_p - \mathbf{P}^{(i-1)^2})\mathbf{H}_k + \sigma^2\mathbf{I}_R\right]^{-1}\mathbf{H}_k^H\mathbf{Q} \tag{5.13}$$

and

$$\mathbf{B}_k^{(i)} = \mathbf{H}_k\mathbf{F}_k^{(i)} - \mathbf{I}_p \tag{5.14}$$

Where σ^2 is the noise variance and $\mathbf{Q} = diag(Q_1, ..., Q_p)$ represents the normalization PxP diagonal matrix to ensure that

$$\frac{Q_p}{N}\sum_{k=0}^{N-1}\sum_{r=1}^{R} F_{k,p}^{(r,i)} H_{k,p}^{eq(r)} = 1 \tag{5.15}$$

Replacing (5.2) and (5.11) in (5.8) we get the following expression,

$$\tilde{\mathbf{S}}_k^{(i)} = \mathbf{F}_k^{(i)T}\mathbf{H}_k^T\mathbf{S}_k - \mathbf{B}_k^{(i)T}\mathbf{P}^{(i-1)}\mathbf{S}_k - \mathbf{B}_k^{(i)T}\boldsymbol{\Delta}_k + \mathbf{F}_k^{(i)T}\mathbf{N}_k \tag{5.16}$$

and replacing 5.14 in the previous eq. we have,

$$\tilde{\mathbf{S}}_k^{(i)} = \underbrace{\mathbf{I}_p\mathbf{S}_k}_{\text{DS}} + \underbrace{\left(\mathbf{F}_k^{(i)T}\mathbf{H}_k^T - \mathbf{I}_p - (\mathbf{F}_k^{(i)T}\mathbf{H}_k^T - \mathbf{I}_p)\mathbf{P}^{(i-1)}\right)\mathbf{S}_k}_{\text{ISI + MAI}} - \underbrace{(\mathbf{F}_k^{(i)T}\mathbf{H}_k^T - \mathbf{I}_p)\boldsymbol{\Delta}_k}_{\text{Feedback Noise}} + \underbrace{\mathbf{F}_k^{(i)T}\mathbf{N}_k}_{\text{Channel Noise}} \tag{5.17}$$

Note that as the number of iterations increases the matrix $\mathbf{P}^{(i-1)}$ tends to an identity matrix, i.e., $\mathbf{P}^{(i-1)} \rightarrow \mathbf{I}_p$, and thus the (ISI + MAI) term tends to zero.

In the first iteration $\mathbf{B}_k = 0$ and $\mathbf{P}_k = 0$, therefore the system is reduced to the linear multi user detector FDE.

If we consider a single antenna UE the forward and feedback filters given by

$$F_k^{(r,i)} = \frac{H_k^{eq(r)}Q}{\sigma^2 + \sum_{r'=1}^{R}\left|H_k^{eq(r')}\right|^2} \tag{5.18}$$

and

$$B_k^{(i)} = \sum_{r'=1}^{R} F_k^{(r',i)} H_k^{eq(r')} - 1 \tag{5.19}$$

where Q is selected to ensure that

$$\frac{1}{N} \sum_{k=0}^{N-1} \sum_{r=1}^{R} F_k^{(r,i)} H_k^{eq(r)} = 1 \tag{5.20}$$

5.2 Simulation Platform

In this section we describe the main blocks of the SC-FDMA platform implemented (see Figs. 5.5 and 5.7). The implemented platform allows to configure the parameters presented in table 5.1, where N is the size of the FFT, L the number of blocks per OFDM symbol and $Nc = NL$.

Transmitter Blocks

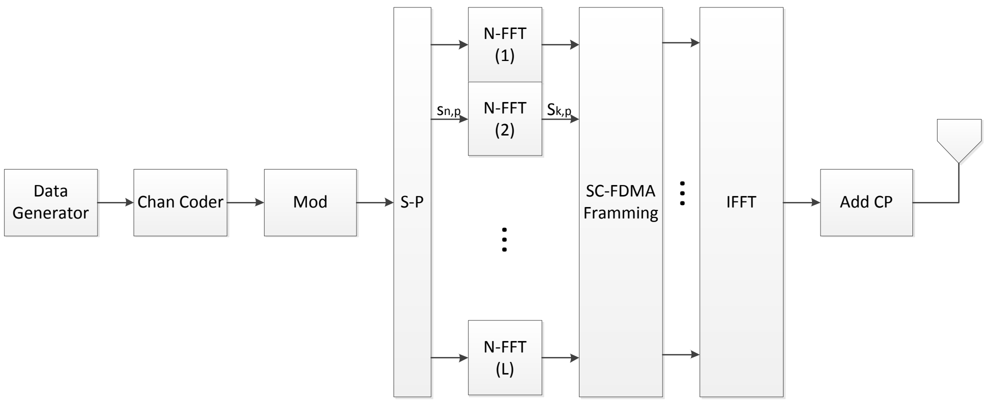

Figure 5.5: Transmitter Blocks for a generic user p

- **Data Generator** - generates a random vector of binary data where it's length is defined by the $N\_Codeword = 1536$.
- **Channel Coder** - performs the Convolutional Turbo Code (CTC) channel coding.
- **Data Modulation** - maps a given binary data vector into constellation symbols. QPSK, 16-QAM and 64-QAM are supported.

- **N-FFT** - depending on the size of the FFT given by the N value, it splits the data vector into L blocks with N size and then performs the FFT for each of those blocks individually
- **SC-FDMA Framming** - this block maps the FFT block into subcarriers which are allocated into SC-FDMA frames as depicted in Fig. 5.6. It selects the interleaved mode off (adjacent mapping) or on (interleaved mapping). This block also performs the IFFT and adds the CP.

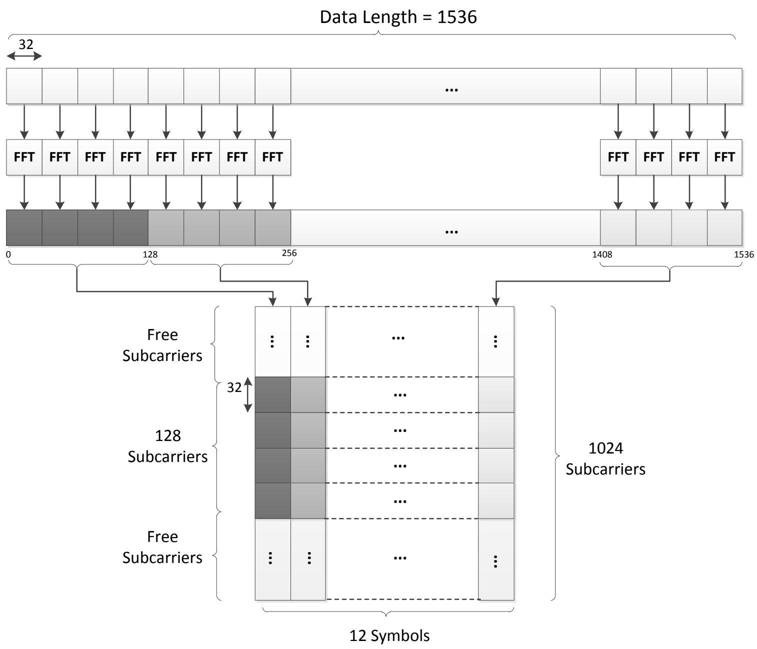

Figure 5.6: SC-FDMA Framing considering a FFT size of 32

Receiver Blocks

- **Multi-user Equalizer Block** - it performs the IB-DFE in frequency domain described in the previous section.
- **SC-FDMA De-framming** - de-maps the subcarriers into a vector and de-interleaves them if they were previously interleaved.
- **IFFT** - splits the vector after the de-framming block into blocks with N size and performs the IFFT of those blocks converting them back to the time domain
- **Data Demodulation** - converts the symbols that were mapped by a chosen constellation into a stream of bits.
- **Channel Decoder** - reverse the process of Coder's block

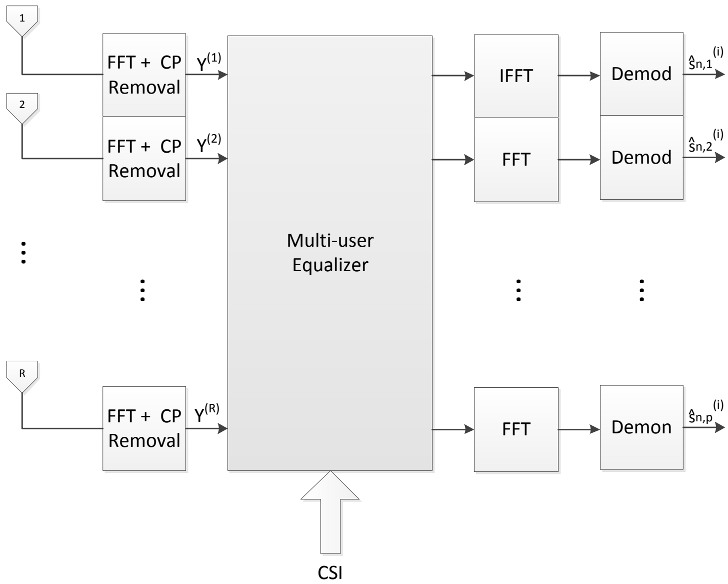

Figure 5.7: Receiver Blocks of the BS

The simulation requires the introduction of the following parameters:

Table 5.1: Input Parameters of the simulation

Parameter	**Variable**	**Options**
Modulation	m	QPSK, 16-QAM or 64-QAM
Number of UEs	P	1-4
Number of Receiving antennas	R	1-4
Coder	Coder	Off or CTC
FFT Size	N	16,32, 64 or 128
Interleaver	Interleaver_on_off	On or Off

5.3 Results

In this section we present the performance results of the implemented iterative multi-user FDE. The main parameters used in the simulations are based on LTE standard and are presented in table 5.2. To get those results we considered an uncorrelated Rayleigh fading channel model. Finally it was assumed that there is perfect synchronization and channel estimation. The results are presented in terms of average BER as function of Eb/N0.

Table 5.2: Chosen Parameters for the Simulation

Parameter	**Value**
FFT Size	1024
Number of Available Subcarriers	128
Sampling Frequency	15.36 Mhz
Useful Symbol Duration	66.67μs
Cyclic prefix Duration	5.21μs
Total Symbol Duration	71.88μs
Subcarrier Separation	15 Khz
Frame Size	12 OFDM Symbols
Modulation	QPSK
Channel Coder	CTC
Code Rate	1/2 - 1536/3072
Number of Transmitting Antennas per UE	1
Number of Receiving Antennas	1, 2 or 4

Also, for the decoder a Max Log MAP algorithm with 8 iterations was used. For the simulations all UEs transmits with a single antenna. We considered several scenarios:

1. **SISO System**

   In this first scenario we assumed a single antenna UE transmitting to a single antenna BS, i.e., we have a SISO scenario. We compare the performance results of the linear frequency domain ZF equalizer (presented in 4.16) and the implemented IB-DFE scheme for 1, 2 and 3 iterations as shown in Fig. 5.8. Note that the first iteration correspond to the linear frequency domain MMSE equalizer (see 4.24). Also, we present the curve for the Matched Filter (MF) for comparison, in this scheme the ISI is zero and thus a diversity order of N can be achieved. In this single user scenario the performance is basically affected by the ISI and the noise. From the results we can observe that the MMSE (IB-DFE with 1 iteration) outperforms the linear ZF as expected, since the noise in enhanced namely in the frequency deep notches. Increasing the number of iteration the implemented algorithm tends to MF. In can be seen that the penalty against the MF is approximately 6dB for 1 iteration and 2 dB for 3 iterations (considering a BER=1.0e^{-3}). For the next simulations we omit the ZF equalizer.

2. **SIMO 1x2 System**

   In this scenario also consider a single antenna UE but the BS is equipped with 2 antennas. The results for this scenario are depicted in Fig. 5.9. Since there is no MAI and the diversity order is 2, in the first iteration we get better performance gain than the **SISO System** (for example for BER=1.0e^{-3} the penalty against the MF is just 2.3dB instead of 6dBs in the **SISO System**). For the second iteration there is improvement of almost 1dB in relation to the first iteration but a negligible improvement into the third iteration.

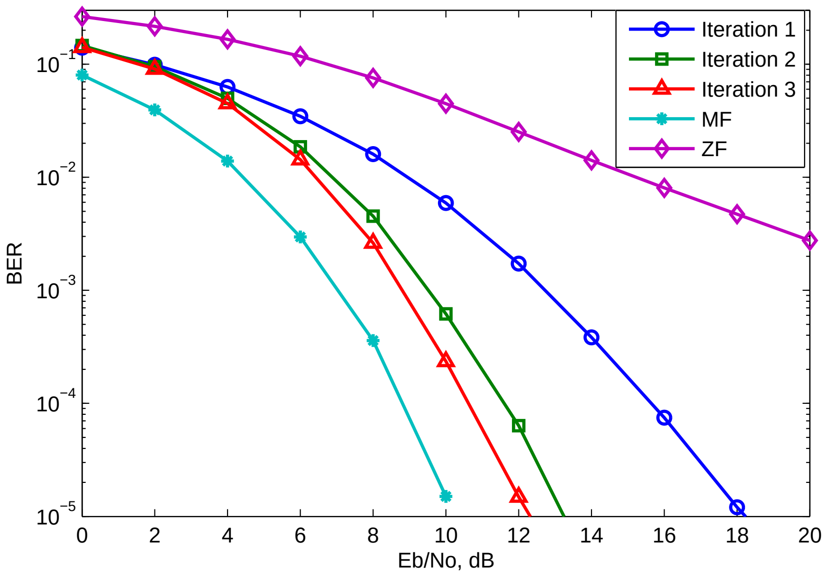

Figure 5.8: Performance evaluation of the implemented Multi-user Equalizer for SISO System

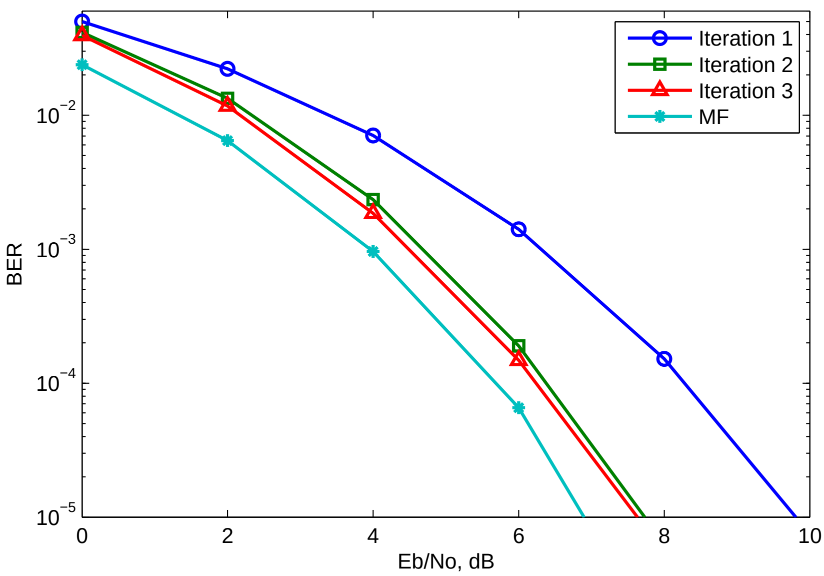

Figure 5.9: Performance evaluation of the implemented Multi-user Equalizer for 1x2 SIMO System

3. **Virtual MIMO 2x2 System**

   Now it is presented a scenario where 2 single antenna UEs transmit to a BS with 2 antennas, making up a virtual MIMO 2x2 scheme where the results are shown in Fig. 5.10. Unlike the previous scenarios, in this one we get not only ISI but also MAI. Considering a BER=$1.0e^{-3}$ the first iteration has a penalty against the MF of approximately 8dB and there is a improvement of 5dB for the second iteration, which means the proposed algorithm removes both ISI and MAI efficiently. From the third iteration, the performance gain is negligible. We can see a penalty of less than 1dB against the MF.

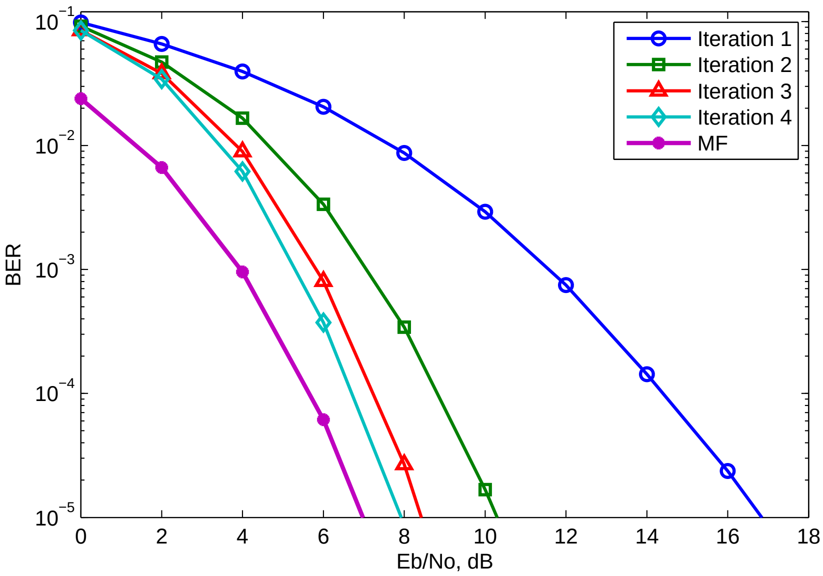

Figure 5.10: Performance evaluation of the implemented Multi-user Equalizer for Virtual MIMO 2x2 System

4. **Virtual MIMO 2x4 System**

   In this muti-user scenario we have UEs transmitting to 4 receiving antennas of the BSs. Since we have a diversity order higher than the **Virtual MIMO 2x2 System**, the signal at the receiver is less affected by ISI and MAI. In comparison to the **Virtual MIMO 2x2 System** in the first iteration we will have less 5dB of penalty against the MF (8dB against 3dB respectively for BER=$1.0e^{-3}$, see Figs. 5.10 and 5.11). The next iterations are almost coincident and very close to the MF with less than 1dB of penalty, meaning that there is no need to use more than 2 iterations for this scenario.

5. **Virtual MIMO 4x4 System**

   Finally it is presented a scenario where 4 UEs are transmitting to a BS with 4 receiving antennas. Clearly, this is the scenario with high level of MAI. From the Fig. 5.12 we can observe a penalty of approximately 1dB against the **Virtual MIMO 2x2 System** for the first iteration and a for BER=$1.0e^{-3}$. The second iteration removed a lot of interference with less 6dB of penalty against the MF than the first iteration due to the cancellation of the extra MAI while the third iteration has just 1dB of penalty for the same BER. We can see that with 4 iterations and for high Eb/N0 the performance is close to the one obtained with MF

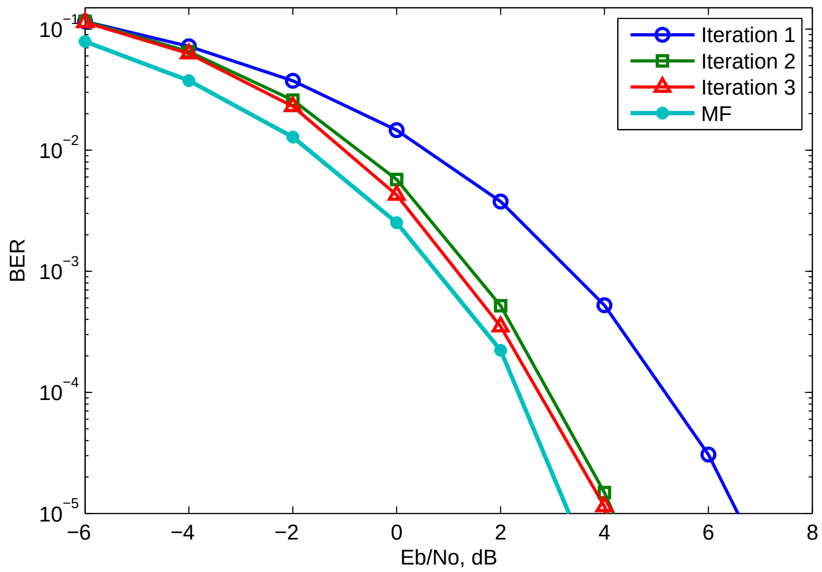

Figure 5.11: Performance evaluation of the implemented Multi-user Equalizer for Virtual MIMO 2x4 System

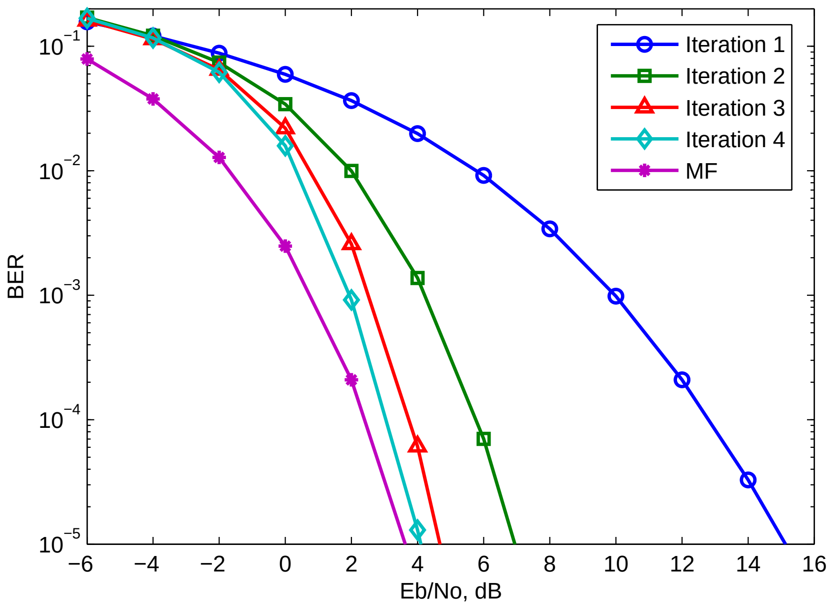

Figure 5.12: Performance evaluation of the implemented Multi-user Equalizer for Virtual MIMO 4x4 System

5.3.1 Impact on the Near-Far Effect

The near-far effect occurs when multiple UEs are transmitting to a BS and they are differently spaced from it. The closer UEs will transmit with higher power and thus making the detection harder or impossible of the users which are more farther to the BS whose their signal is weaker.

The results of the next figures were obtained for the virtual MIMO 2x2 System. It is assumed that the second user has higher power (3dB, 6dB and 9dB) than the first one. The curves of Figs 5.13, 5.14 and 5.15 are refereed to the user 1. The aim is to assess the impact that the users with higher power has on the weaker users.

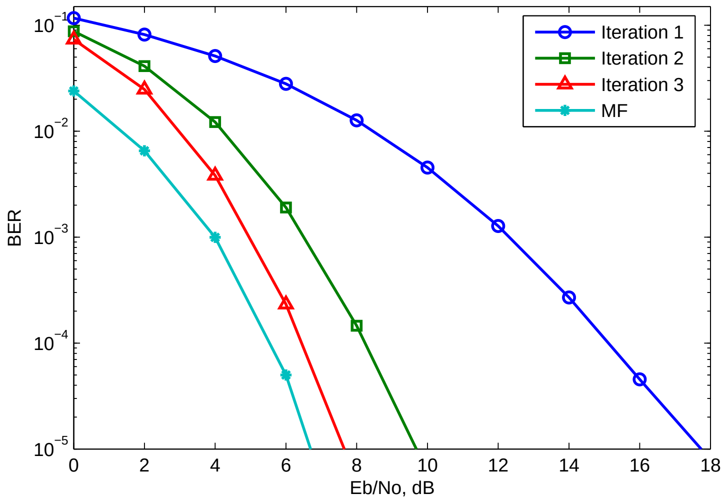

Figure 5.13: Impact on user 1 when the user 2 has 3dB higher power than the first user

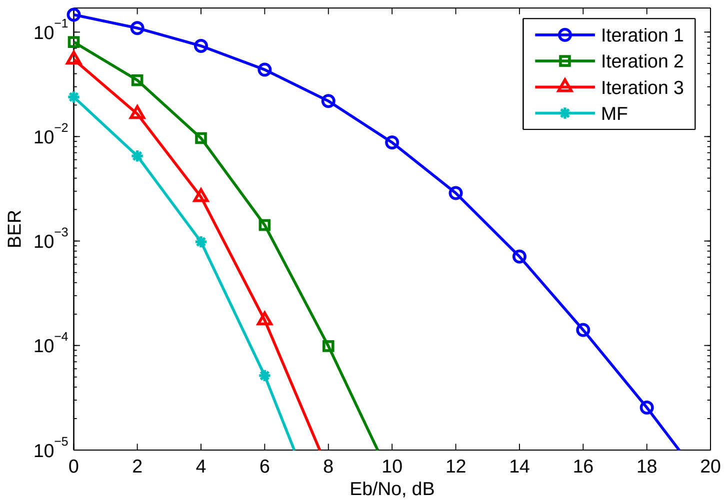

Figure 5.14: Impact on user 1 when the user 2 has 6dB higher power than the first user

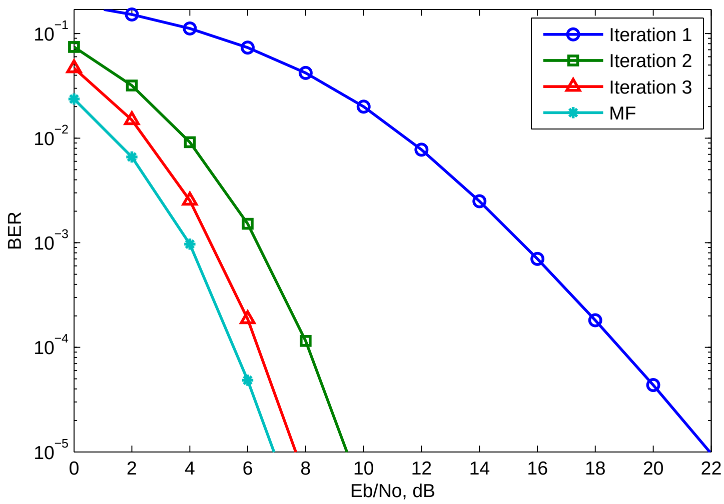

Figure 5.15: Impact on user 1 when the user 2 has 9dB higher power than the first user

As expected, with the increase of power transmitted by the user 2 the performance of the user 1 degrades more. For the first iteration, when BER=$1.0e^{-4}$ the penalty against the MF is 9dB, 11dB and 13dB for the cases where the user 2 transmits with more 3dB, 6dB and 9dB, respectively. For the second and third iterations the penalty against the MF between the simulations is the same for all the 3 simulations (for example, 2.5dB of penalty for BER=$1.0e^{-4}$ in the second iteration) that means that the implemented system also removes the near-far effect caused by the difference of power between the user 1 and user 2.

5.3.2 Simulation with Channel Coding

The results of the Fig. 5.16 were obtained for the virtual MIMO 2x2 System. The channel was coded using a CTC. For the first iteration, when the BER=$1.0e^{-3}$ there is penalty of approximately 1.25 dB which is less 6.75 dB than the **Virtual MIMO 2x2 System** without channel coding. For the second iteration the penalty was just 0.75dB and there was a negligible performance gain for the third iteration.

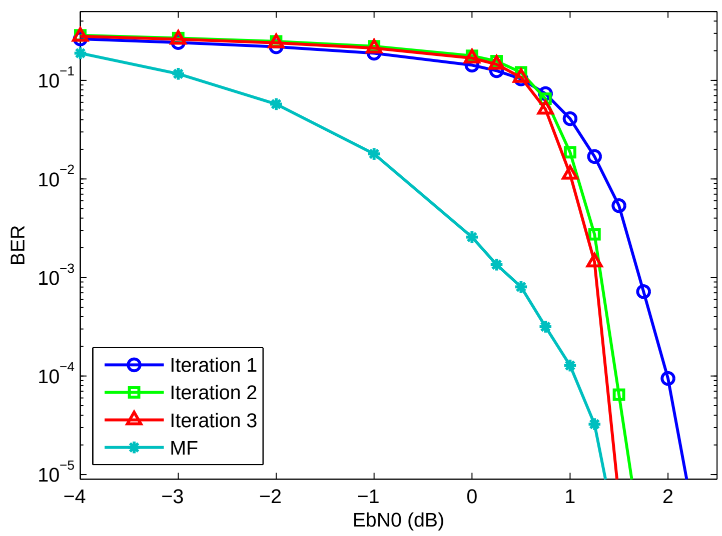

Figure 5.16: Virtual MIMO 2x2 with channel coding - BER

Chapter 6

Conclusion and Future Work

Wireless communications plays an important role in our society and they are in constant progress to adjust to the increasing number of users and services. Several new techniques and technologies have been implemented along the years. In terms of cellular communications major advances where divided by the so called generations, the first one characterized by using analog voice communications, the second had digital voice communications, the third had multimedia data services and now with the introduction of 4G the concept "Anywhere, anything, anytime" gains a complete new meaning with high data rates, great cell coverage and high mobility for the end user. LTE is not the last technology of cellular systems, LTE Advanced is already being standardized for implementation.

In this thesis we started with a quick overview of the evolution of cellular communications before introducing LTE. In the next chapter it was introduced the specifications, requirements and the network aquiculture of this cellular technology, just before introducing in chapter 3 the fundamental radio access network changes, namely the introduction of OFDM versus the 3G's CDMA that were essential to provide high spectral efficiency and high data rates. In this chapter it was also described the multiple access schemes used in LTE for downlink (OFDMA) and uplink (SC-FDMA) that were different due to the autonomy of UEs while sharing most of the architecture. In chapter 4 it was introduced the concept of multiple antennas with MIMO systems and the importance of MIMO schemes to increase spatial diversity and data rates, referring some techniques used based on block codes in time and frequency domain for the transmitter and frequency domain equalizers for the receivers.

In Chapter 5, we derived an iterative multi-user frequency domain equalizer for the "virtual" MIMO SC-FDMA scenario in the LTE standard. A flexible simulation platform was implemented based on LTE specifications. In that platforms several parameters can be configured such as: number of UEs, number of BS's antennas, number of subcarriers, modulation, channel coding, etc.

The results have shown that the proposed iterative equalizer efficiently removes both ISI and MAI even in scenarios with near-far effect. In the most studied scenarios only a few number of iterations (2-4) were enough to obtain the performance obtained with the MF. Since the algorithm requires a reduced number of iterations the complexity is not significant and thus can be used in practical systems. Finally, by using CTC channel coding we get major improvements in the performance of the IB-DFE.

6.1 Future Work

Some future work can be done,

- In this work it was assumed single antenna UEs. It would be interesting to extend for the case where the UEs are equipped with an antenna array.
- In this work we considered hard decisions in the feedback loop but it would be interesting to compare the performance considering soft decisions.
- It was assumed perfect channel information at the BSs, it would be interesting to assess the algorithm under imperfect channel information.

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