Departamento de Electrónica, Telecomunicações e Informática

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# Técnicas com Múltiplas Antenas Distribuídas para Sistemas sem Fios

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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 Dr. Adão Silva, Professor Assistente Convidado do Departamento de Electrónica, Telecomunicações e Informática da Universidade de Aveiro

O trabalho apresentado tem o apoio do projecto Europeu CODIV FP7-ICT-2007.

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**Agradecimentos /**

**Acknowledgements**

Ao professor Adão Silva, pelo seu comportamento exemplar ao longo de todo o processo, estando sempre disponivel para esclarecer qualquer dúvida. A sua atenção revelou-se importante para que eu nunca me desviasse dos meus objectivos e conseguisse levar este trabalho até ao fim.

Aos meus colegas do IT Andreia Moço, Ludimar Guenda, e Mayra Alejandra, que se revelaram sempre prestáveis para me ajudar.

A todos os outros membros do IT que directa ou indirectamente acabaram por contribuir para o bom ambiente de trabalho.

Á minha familia, que sempre me apoiou.

**Palavras-chave**

Sistemas com *Relay*, MIMO virtual, Diversidade cooperativa, Diversidade espacial, Agregados de antenas, *Decode and Forward*, *Amplify and forward*, OFDM, Propagação multipercurso

**Resumo**

Transmissão cooperativa, em que uma fonte e um *relay* cooperam para enviar uma mensagem para o destino, pode proporcionar diversidade espacial contra o desvanecimento nas comunicações sem fios. O objectivo deste projecto é estudar a performance de um sistema de transmissão cooperativo com dois *relays* equipados com duas antenas, entre o transmissor e o utilizador. Considera-se que a estação base está equipada com duas antenas e o terminal móvel apenas com uma. O sistema cooperativo foi implementado de acordo com as especificações do LTE e avaliado em diversos cenários de propagação, considerando canais com diferentes Relação Sinal Ruído (SNR). Verificou-se que o desempenho do sistema proposto é melhor, quando comparado com o sistema não cooperativo, na maior parte dos cenários estudados.

**Key-words**

Relay systems, MIMO virtual, Cooperative Diversity, Spatial diversidade, aggregates of antennas, Decode and Forward, Amplify and forward, OFDM, Multipath propagation

**Abstract**

Cooperative transmission, in which a source and relay cooperate to sent a mensage to destination, can provide spatial diversity against fading in wirless telecomunications. The goal of this project is to study the perfomance of a cooperative tranmition systems with two relays equiped with two antennas, between transmitter and user. It is considered that the base station is equipped with two antennas and the mobile terminal with only one. The cooperative system was implemented according to the specifications of the LTE and evaluated at several propagation scenarios, considering channels with diferents Signal to Noise Ratio (SNR). It was found that the perfomance of the proposed system is better when compared with the non-cooperative ones, in most scenarios considered.

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# Glossary of Terms

| | |
|---|---|
| AC | Auxiliary Channel |
| ACI | Adjacent Channel Interference |
| ACPR | Adjacent Channel Power Ratio |
| AWGN | Additive White Gaussian Noise |
| ACF | Auto Correlation Function |
| AMPS | Advanced Mobile Phone System |
| BER | Bit Error Rate |
| BS | Base Station |
| CODIV | Cooperative Diversity |
| CSI | Channel State Information |
| CRC | Cyclic Redundancy Check |
| DBPSK | Differential Binary Phase Shift Keying |
| DQPSK | Differential Quadrature Phase Shift Keying |
| DFT | Discrete Fourier Transform |
| DF | Decode and Forward |
| DoT | Direction of Transmission |
| DoA | Direction of Arrival |
| DL | Direct Link |
| EVCM | Exported Variable Complexity Metric |
| EDGE | Enhanced Data Rates for GSM Evolution |
| FDD | Frequency Division Duplex |
| GSM | Global System for Mobile Communications |
| GPRS | General Packet Radio Service |
| HSCSD | High Speed Circuit-Switched Data |
| ISI | Inter-Symbol Interference |
| LAN | Local Area Network |
| LTE | Long Term Evolution |
| LOS | Line of Sight |

| | |
|---|---|
| MIMO | Multiple Input Multiple Output |
| MISO | Multiple Input Single Output |
| MAN | Metropolitan Area Networks |
| MCR | Maximum Ratio Combiner |
| ML | Maximum Likelihood |
| MMR | Mobile Multi Hop Relay |
| OFDM | Orthogonal Frequency-Division Multiplexing |
| OSTBC | Orthogonal Space Time Block Code |
| PAL | Phase Alternating Line |
| PDSCH | Physical Downlink Shared Channel |
| QAM | Quadrature Amplitude Modulation |
| QPSK | Quadrature Phase Shift Keying |
| QoS | Quality of Service |
| QSTBC | Quasi/Orthogonal Space Time Block Code |
| QSFBC | Quasi/Orthogonal Frequency Time Block Code |
| RB | Resource Block |
| RN | Relay Node |
| Rx | Receiver |
| RMS | Root Mean Square |
| SISO | Single Input Single Output |
| SNR | Signal-to-Noise Ratio |
| SSDT | Site Selection Diversity Transmition |
| STD | Selective Transmit Diversity |
| STTD | Space Time Transmit Diversity |
| STC | Space time Codes |
| SU | Subcarrier Unit |
| CPICH | Common Pilot Channels |
| S-CPICH | Secondary Common Pilot Channels |
| TS | Transport Stream |
| Tx | Transmitter |
| TSTD | Time Switched Transmit Diversity |
| TDD | Time Division Duplex |
| UHF | Ultra High Frequency |

| | |
|---|---|
| UMTS | Universal Mobile Telecommunication System |
| UT | User Terminal |
| UE | User Equipment |
| WCDMA | Wide-Band Code Division Multiple Access |
| WIMAX | Worldwide Interoperability for Microwave Access |
| VHF | Very High Frequency |
| ZF | Zero Forcing |
| 3GPP | 3rd Generation Partnership Project |

# Chapter 1

## Introduction

## 1.1 Introduction to broadband wireless

The evolution from 1G to 2G was a natural one, from analogue to a digital, with all the advantages it brings about. We are talking about a new system, but exactly the same service, voice. However, there were totally different approaches to this evolution in Europe and the US [1]: while in Europe the choice was made to go from a variety of incompatible analogue systems at 450 MHz and later at 900 MHz a single digital system, GSM (Global System for Mobile Communications). to finally have pan- European (nowadays quasi-global) roaming, in the US (United States) the reverse happened, with AMPS (Advanced Mobile Phone System) being superseded by a variety of incompatible systems at 800 h4Hz and GSM 1900, Omni point IS-661, PACS). While the evolution from 1G to 2G was necessarily revolutionary from the technology point of view, the idea was from the start, at least in Europe, to have 2G as an evolving platform GSM and the annual releases leading to the introduction of HSCSD (High Speed Circuit-Switched Data), GPRS (General Packet Radio Service) and EDGE (Enhanced Data rates for GSM Evolution). The evolution from 2G to 3G (also known as IMT- 2000) corresponds yet again to a adopting a new system(from GSM to UMTS (Universal Mobile Telecommunication System ) and its WCDMA (Universal Mobile Telecommunication System) and TD/CDMA modes, but most of all, to a change of focus from voice to Mobile Multimedia. Another first was the simultaneous support of several QoS (Quality of Service) classes in a single radio interface. The objective set from the start was to achieve global roaming, although that has become difficult with five different codes approved by ITU-R for IMT-2OOO (CDMA Direct Spread, CDMA Multi-Carrier, CDMA TDD (Time Division Duplex), TDMA Single Carrier, FD-MA). Once more, 3G was from the start conceived in Europe as an evolutionary platform, at the image of the very successful

GSM, with UMTS-Phase 2 planned to push data rates well above 2 Mbits into the 10 Mbits range.

In order for 4G to deserve that designation, to be a clear evolution from 3G, it has to bring about some clear changes. Besides new technologies in some areas where gaps have been identified (personal area networks and body LAN (Local Area Network) low power sensors, networked appliances and self-configuring ad hoc networks), we contend that what will define 4G is the ability to integrate all systems, and easy access to all services, all the time and everywhere. Also allowing for the integrated provision of personalized, enhanced services over the most efficiently preferred networks, depending on the user profile, on the type of data stream under consideration, and on the traffic load in the available networks. Furthermore, 4G will be designed to take into account multiple classes of terminals, adjusting content delivery to the terminal capabilities and the user profile. All four generations times and principal features are showed in fig.1 and fig2. Yet inside 4G, is relevant to talk about the most recent so-called 4G technologies: LTE (3GPP Long Term Evolutions, also called UMTS long term evolution) and WIMAX (Worldwide Interoperability for Microwave Access). This last one is a telecommunication protocol that provides fixed and fully mobile internet access. The current WIMAX revision provides up to 40 Mbits/s with the update expected to offer up to 1 Gbit/s fixed speeds. The first one, the LTE, in addiction to enabling fixed to mobile migrations of internet applications, also provide the capacity to support an explosion in demand for connectivity from a new generation of consumer devices. This new standard will ensure that UMTS remains competitive while giving users enhanced mobile Internet access. It is foreseen that the first commercial LTE networks could be in place by 2010, and LTE standardization is progressing as part of Release 8 from the 3GPP (3rd Generation Partnership Project).

LTE must provide downlink data rates of higher than 100 Mbps and uplink rates of higher than 50 Mbps. It must also significantly reduce the latency times for packet transmissions so users will not experience unacceptable delays. In short beyond the improvement in bit rates, LTE aims to provide a highly efficient, low-latency, packet-optimized radio access technology offering enhanced spectrum flexibility. Has also a support of variable bandwidth, simple protocol architecture and of course, compatibility and inter-working with earlier 3GPP releases.

Another significant feature of LTE is its high bandwidth up to 20 MHz. Because the usable bandwidth is scalable, LTE can also operate in the existing 5 MHz UMTS frequency bands, or in even smaller bands. Developers of LTE base stations and wireless devices must also account for a very short latency time; LTE has a transmission time interval of only 1 ms between data packets. LTE system can also employ multiple-input multiple-output (MIMO) antenna systems (Chapter 4).

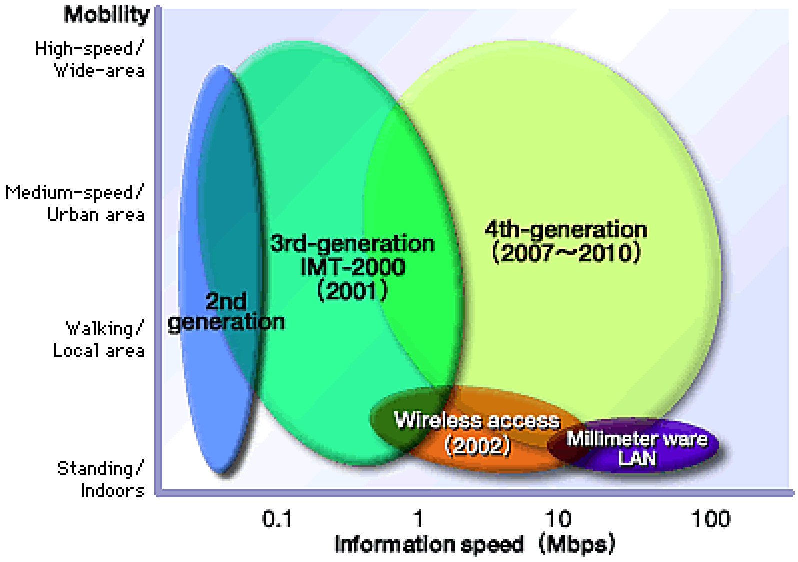



Fig. 1 Speed communications evolution

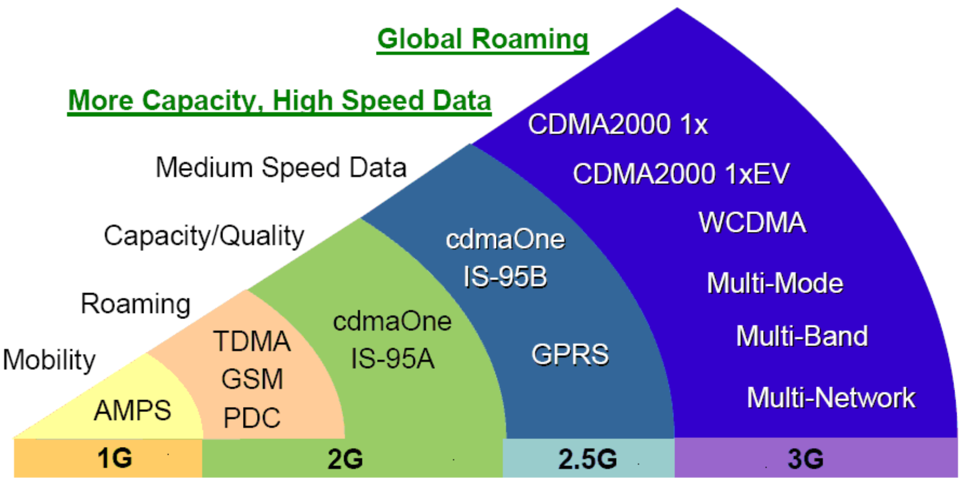



Fig. 2 Communication generations description

## 1.2 Motivation and Objectives

Owing to their ability to multiply wireless capacity, mitigate slow fading, and facilitate new adaptive communications beyond the limits of conventional single-antenna wireless systems, MIMO and Space Time Coding techniques have generated much research interest in recent years. Their adoption in cellular mobile radio, wireless LAN and wireless MAN (Metropolitan Area Networks) standards have also marked their increasing significance in commercial broadband wireless systems. The necessity of develop this techniques appear not only because the increase of users but also because of movement of a mobile user and movement of different objects along the path, which causes that the propagation path between the transmitter and the receiver may vary from LOS (simple line-of-sight) to very complex ones. Insufficient signal-to-noise ratio (SNR) at a certain instant of time may result in incorrect reception of the channel symbol possibly leading to an erroneously received bit. However, the problems inherent of these systems such as shadowing, significant correlation between channels in some environments,, significantly degrades the capacity gains promised by MIMO techniques.

Cooperative diversity is a promising solution for wireless systems to overcome the above limitations, improving system fairness, extending coverage and increasing capacity. In this thesis, it is implemented and assessed the performance of relay-assisted or cooperative schemes designed for downlink OFDM based systems, using distributed space-frequency block coding protocols. It is assumed the use of an antenna array at the BS (base station) and a single antenna at the UT (user terminal). At the RN relay node it is considered either single antenna or an antenna array. Also, it is assumed an half-duplex decode-and-forward relay communication protocol. This cooperative scheme it is an extension of the one proposed under the European CODIV (Cooperative Diversity) project [2], where all terminals were equipped with single antenna. The performance of these cooperative schemes is evaluated under realistic scenarios, considering typical pedestrian scenarios based on LTE specifications. The proposed schemes are also compared against the non-cooperative OFDM based systems. Numerical results show that the availability of antenna arrays at the relays significantly improves the cooperative systems performance, which outperform the non-cooperative ones in most studied scenarios.

## 1.3 Thesis Structure

The remainder of the thesis is organized as follows:

In chapter 2 we describe the most important skills of a communication channel. We start by making a comparison between a solid connection and a wireless connection, and then making focus in wireless channels. We describe the statistical models of fading and channel parameters as delay spread, coherence bandwidth and frequency selectivity.

Chapter 3 makes a description of a multicarrier system, with all transformation of a OFDM signal from the transmitter until the receiver. Also we make an introduction of a OFDM system applied to LTE.

In chapter 4, the focus is on building the foundation of MIMO, MISO and SIMO concepts that will be used extensively in cooperative communications and networking. Introduce the concept of multipath diversity, macro-diversity, received antenna diversity, transmit diversity and the concerned in maximum achievable space–time–frequency diversity available in broadband wireless communications. Finally we present the design of broadband space–frequency orthogonal and quasi-orthogonal codes.

In the fifth chapter we introduce the conceit of node and describe the simplest cooperative scheme, the single relay cooperative scheme. We give an over view of the most important transmition protocols, AF (Amplify and Forward) and DF (Decode and Forward) and we describe a multi node cooperative system. Also we give an overview in the world wile initiatives that are being developed.

In chapter 6, we present the proposed relay-assisted system based on distributed space-frequency block coding schemes. We describe detail of all calculations for all studied schemes and present the graphical results.

Finally, last chapter presents the conclusions of this work and outlines some future work perspectives.

# Chapter 2

# 2 Communication Channels

## 2.1 Wireless channels

There exists a large variety of channels, which may be divided into two groups. If a solid connection exists between transmitter and receiver, the channel is called a wired channel.

If this solid connection is missing, this connection is called a wireless channel. Wireless channels differ from wired channels, due to their unreliable behavior compared to wired channels. In wireless channels the state of the channel may change within a very short time span. This random and severe behavior of wireless channels turns communication over such channels into a difficult task. There are several different classifications regarding the wireless channels. Wireless channels may be distinguished by the propagation environment encountered.

The wireless channel puts fundamental limitations on the performance of wireless communication systems. The transmission path between the receiver and the transmitter can be altered from simple line-of-sight to one that is drastically obstructed by buildings, foliage and mountains. Even the speed of the mobile impacts how rapidly the signal level fades. Modeling the wireless channel has historically been one of the most difficult parts of the communication system design and is typically done in a statistical manner, based on measurements made specifically for a designated communication system or spectrum allocation.

Huygen's principle when there is an obstruction between the transmitter and the receiver antennas and as a result of this, there are secondary waves generated behind the obstructing body. Scattering arises when the incident wavelength is in the order of or larger than the dimension of the blocking object with non-regular shape, and causes the transmitting energy to be redirected in many directions. The relative importance of these three propagation mechanisms depends on the particular propagation scenario.

## 2.2 Statistical Models

In general, it is widely accepted that radio signals propagate according to the mechanisms of reflection, diffraction and scattering, which in general characterize the radio propagation by three nearly independent phenomena: path loss variance with distance, shadowing (slow fading) and multipath fading (fast fading) [3], fig.4. Multipath fading is introduced due to constructive and destructive combination of randomly delayed, reflected, scattered and diffracted signal components, while shadowing affects the link quality by slow variation of the mean level [4]. In many cases, multipath fading and shadowing occur simultaneously.

Wireless channels are subject to random fluctuations in received power arising from the multipath propagation. These short-term fading channels have been generally modeled as Rayleigh, Rice, Nakagami and Weibull [5].

Diversity techniques [6] are generally employed to mitigate the effects of short-term fading. Often, these channels are also subject to long-term fading or shadowing caused by the multiple scattering which take place in the channel. The shadowing is commonly modeled using a lognormal distribution [7].

## 2.3 Median Path Loss

Path loss models describe the signal attenuation between transmitter and receiver antenna as a function of the propagation distance and other parameters. Some models include many details of the terrain profile to estimate the signal attenuation, whereas others just consider carrier frequency and distance. The path loss given by the propagation models is the median value:

$$L_T = L + L_s, \tag{2.1}$$

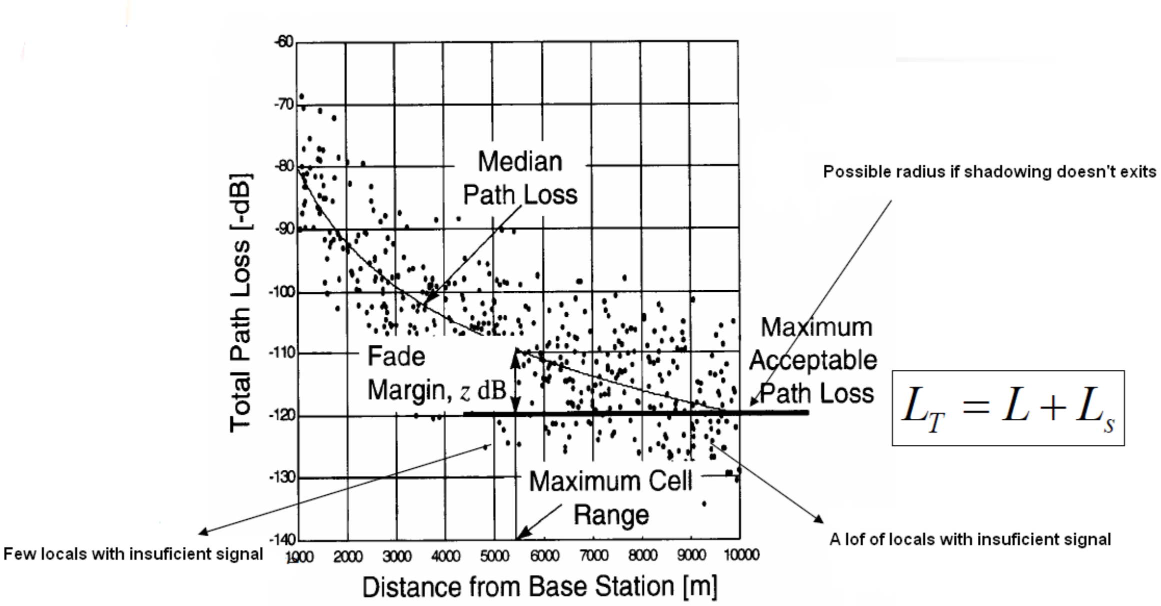



**Fig. 3 Path loss Variation**

Where $L_T$ represents the total path loss, $L$ the path loss caused by fast fading effect and $L_S$ the path loss for the slow fading, terms that we describe below. Fig.3 shows us a path loss variation along the space.

If we are talking about a small special scale the path loss is influenced by the fast fading effect if we assume a medium or bigger scale it is influenced by shadowing or slow fading such we will discuss in next topics.

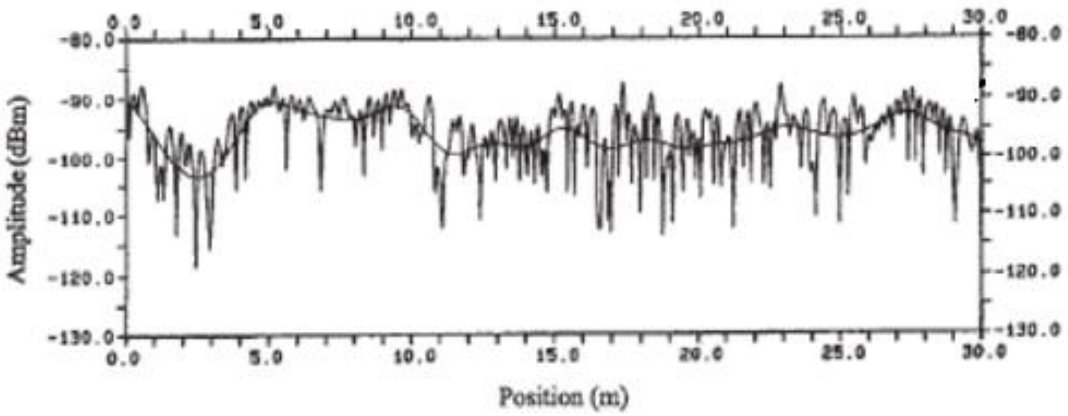



**Fig. 4 Slow and Fast fading space representation**

## 2.4 Fading

Most mobile communication systems are used in and around center of population, the transmitting antenna or BS are located on top of a tall building and tower or mountain and the mobile antenna or UT is well below the surrounding buildings. The Wireless communication phenomena are mainly due to scattering of electromagnetic waves from surfaces or diffraction over and around buildings. In fact this scatting of waves and diffractions create multiple propagation paths or multipath which have both slow and fast aspects. The received signal for narrowband excitation is found to exhibit three scales of spatial variation: Fast Fading, Slow Fading and Range Dependence, as well as temporal variation and polarization mixing.

As the UT moves along the street, rapid variations of the signal are found to occur over distances of about one-half the wavelength $\lambda = \frac{c}{f}$, where $c$ is the light speed and $f$ = 910 MHz. For example, in a few meters distance the signal can vary by 30 $dB$ or even in distances as small as $\frac{\lambda}{2}$, the signal is seen to vary by 20 $dB$ (as shown in fig. 4). In resume, small scale variations resume from the arrival of the signal at the MS along multiple ray paths due to reflection, diffraction and scattering.

The phenomenon of fast fading is represented by the rapid fluctuations of signal over small areas; when the signals arrive from all directions in the plane, fast fading will be observed for all directions of motion. The received signal is the sum of a number of signals reflected from local surfaces, and these signals sum in a constructive or destructive manner; it depends of the phase shift that for it way depends on the speed of motion, frequency of transmission and relative path lengths.

For other way, in response to the variation in the nearby buildings, there will be a change in the average about which the rapid fluctuations take place; this middle scale over which the signal varies, which is on the order of the buildings dimensions is known as Shadow Fading, Slow Fading or Log-Normal Fading.

We can also have a case of a Flat Fading. It happens when the wireless channel has constant gain and linear phase response over a bandwidth which is greater than the bandwidth if the transmitted signal, in other words, flat fading occurs when the bandwidth of the transmitted signal $B$ is smaller than the coherence bandwidth of the channel $B << B_m$ (fig.8) . The flat fading channel effect can be seen as a decrease of the signal-to noise ratio, and since the signal is narrow with respect to the channel bandwidth, the flat fading channels are also known as amplitude varying channels or narrowband channels.

## 2.5 Shadowing

Shadowing causes considerable variability about the mean power predicted by path loss, and it is caused by the different large obstacles found from position to angular position of the mobile in order to the issuer. Because of the physical size of the obstacles that produce shadowing, the scale of significant variation is hundreds wavelengths, and the shadow effect is roughly constant over many tens of wavelengths. However, if we average the signal strength around a circular path centered on the base, the familiar inverse cube or fourth power law reasserts itself. Typical obstacles are hills, large buildings, foliage, etc.. and the resulting excess loss is termed shadowing.

Shadowing causes the power to vary about the path loss value by a multiplicative factor that is usually considered as log normally distributed over than ensemble of typical locales (fig.5). An equivalent statement is that the power in dB contains a path loss term proportional to the log of distance, plus a Gaussian term with zero mean and standard deviation roughly 8 dB (for urban settings – rural topography exhibits a lower standard deviation)..

Causes also coverage holes, which may require fill-in by secondary transmitters. It makes cell boundaries less well defined than simple path loss calculations suggest. In consequence, the algorithms controlling handoff, as mobile progresses from one cell to the next, must be sophisticated, in order to avoid multiple handoffs and reversals of the handoff. Shadowing also shows up when we calculate cellular system capacity. We need an accurate model of the total co-channel interference from near and distant sources, and in this we must account for shadowing [8].

The density shadowing probability density probability function *f(x)* could be calculated this way:

$$f(x) = \frac{1}{\sigma_L\sqrt{2\pi}} e^{\frac{(x-\mu)^2}{2\sigma_L^2}}$$

(2.2)

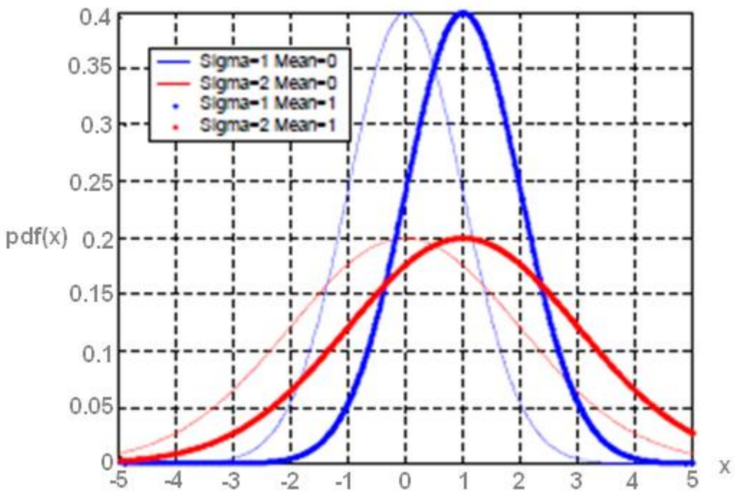



**Fig. 5 Probability density function pdf(x) to a couple values of sigma(μ) and mean(x)**

and,

$$p[\sigma_L > z] \;=\; Q\left(\frac{z}{\sigma_L}\right) \;=\; \frac{1}{2}erfc\left(\frac{z}{\sqrt{2}\sigma_L}\right) \tag{2.3}$$

where $\sigma_L$ denotes the location variability and,

$$erfc(x) \;=\; \left(\frac{z}{\sqrt{\pi}}\right)\int_x^{\infty} e^{-t^2}\,dt \tag{2.4}$$

### 2.5.1 Statistical Distribution for Fast Fading

Such we have seen before, a fixed transmitter in an environment with reflecting objects gives rise to a field formed by the superposition of a number of components traveling along different paths ("multipath"). Each component possesses in general different amplitude, phase, and direction of propagation. The resulting field exhibits spatial changes in amplitude coming from the different spatial rates of variation of the phases of the field components that add up to produce the total field. The resulting field is also referred to as "standing wave pattern." A mobile station traveling through that field will see the spatial field amplitude variations translated into time variations of the received signal amplitude denominated 'fast fading'. It means that fast fading occurs when the channel impulse response changes rapidly within the symbol duration, in other words, fast fading occurs when the coherence time of the channel $T_D$ is smaller than the symbol period of the transmitted signal $T_D \ll T$ (fig.8).

Received pass- band signal without noise after transmitted un-modulated carrier signal $cos(2\pi f_c t)$

$$x(t) = \sum_i \alpha_i(t)\, cos\left(2\pi f_c\big(t - \tau_i(t)\big)\right) = real\{\sum_i \alpha_i(t)\, e^{-j2\pi f_c \tau_i(t)} e^{-j2\pi f_c t}\} \tag{2.5}$$

Where $\alpha_i(t)$ is the time-varying attenuation factor of the $i^{th}$ propagation delay, $\tau_i(t)$ is the time-varying delay and $f_c$ is the carrier frequency. Assume that $\tau_i(t)$where $T_s$ is symbol time. And the equivalent baseband signal is done by,

$$h(t) = \sum_i \alpha_i(t)\, e^{-j2\pi f_c \tau_i(t)} \tag{2.6}$$

If the delays $\tau_i(t)$ change in a random manner and when the number of propagation path is large, central limit theorem applies and $h(t)$ can be modeled as a complex Gaussian process.

When the components of $h(t)$ are independent the probability density function of the amplitude $r = |h| = \alpha$ assumes Rayleigh distribution and the density probability function $f(r)$ is:

$$f(r) = \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} \text{ where } E\{r^2\} = 2\sigma^2 \text{ and } r \geq 0 \tag{2.7}$$

This represents the worst fading case because we do not consider having LOS, and it's because this is the most used signal model in wireless communications. The power is exponentially distributed and the phase is uniformly distributed and independent from the amplitude. This is the most used signal model in wireless communications.

In case the channel is complex Gaussian with non-zero mean (there is LOS), the envelope $r = |h|$ is Ricean distributed. Here we denote $h = \alpha e^{j\emptyset} + \nu e^{j\Theta}$ where $\alpha$ follows the Rayleigh distribution and $\nu > 0$ is a constant such that $\nu^2$ is the power of the mutually independent and uniformly distributed on $[-\pi, \pi]$.

The Rice density probability function *f(r)* is:

$$f(r) = \frac{r}{\sigma^2} e^{-\frac{r^2+\nu^2}{2\sigma^2}} I_0\left(\frac{r\nu}{\sigma^2}\right) \text{ and } r \geq 0 \tag{2.8}$$

Where $I_0$ is the modified Bessel function of order zero and $2\sigma^2 = E\{\alpha^2\}$. The Rice factor $K = \frac{\nu}{\sigma^2}$ is the relation between the power of the LOS component and the power of the Rayleigh component. When $K \rightarrow \infty$, no LOS component and Rayleigh=Ricean.

Weibull distribution represents another generalization of the Rayleigh distribution. When X and Y are zero-mean Gaussian variables, the envelope of $R = (X^2 + Y^2)^{\frac{1}{k}}$ is Rayleigh distributed. However, is the envelope is defined as $R = (X^2 + Y^2)^{\frac{1}{k}}$, the corresponding density probability function $f(r)$ with Weibull distributed:

$$f(r) = \frac{kr^{k-1}}{2\sigma^2} e^{-\frac{r^k}{2\sigma^2}} \tag{2.9}$$

where $E\{r^2\} = 2\sigma^2$.

The last distribution is Nakagami distribution. In this case we denote $h = re^{j\emptyset}$ where the angle *Φ* is uniformly distributed on$[-\pi, \pi]$. The variable $r$ and *Φ* are assumed to be mutually independent.

The Nakagami density probability function is:

$$f(r) = \frac{2}{\tau(k)}\left(\frac{k}{2\sigma^2}\right)^k r^{2k-1} e^{-\frac{kr^2}{2\sigma^2}} \text{and } r \geq 0 \tag{2.10}$$

Where $E\{r^2\} = 2\sigma^2, \tau(.)$ is the Gamma function and k $\geq \frac{1}{2}$ is the fading figure (degrees of freedom related to the number of added Gaussian random variables). If we have $k$ equal to 1, the Rayleigh model assume a particular case which calls Nakagami. Instantaneous receive power is Gamma distributed and it was developed empirically based on measurements.

### 2.5.2 Delay Spread, Coherence Bandwith and Frequency Selectivity

The function determined by the average power associated with each path is called the power delay profile of the multipath channel. Fig. 6 shows the power delay profile for a typical wireless channel. Several parameters are derived from the power delay profile or its spectral response (Fourier transform of the power delay profile), which are used to both characterize and classify different multipath channels:

- The channel delay spread is a measure of the multipath richness of a channel. In general, it can be interpreted as the difference between the time of arrival of the first significant multipath component (typically the line-of-sight component) and the time of arrival of the last multipath component. It is mostly used in the characterization of wireless channels. If the duration of the symbols used for signaling over the channel exceeds the delay spread, then the symbols will suffer from inter-symbol interference. Note that, in principle, there may be several signals arriving through much attenuated paths, which may not be measured due to sensitivity of the receiver. This makes the concept of delay spread tied to the sensitivity of the receiver.

  The delay spread can be characterized through different metrics, although the most common one is the root mean square (rms) delay spread:

  According to Goldsmith [9], let $Ac(\tau)$ be the power delay profile of a channel. Then, the mean delay of the channel is

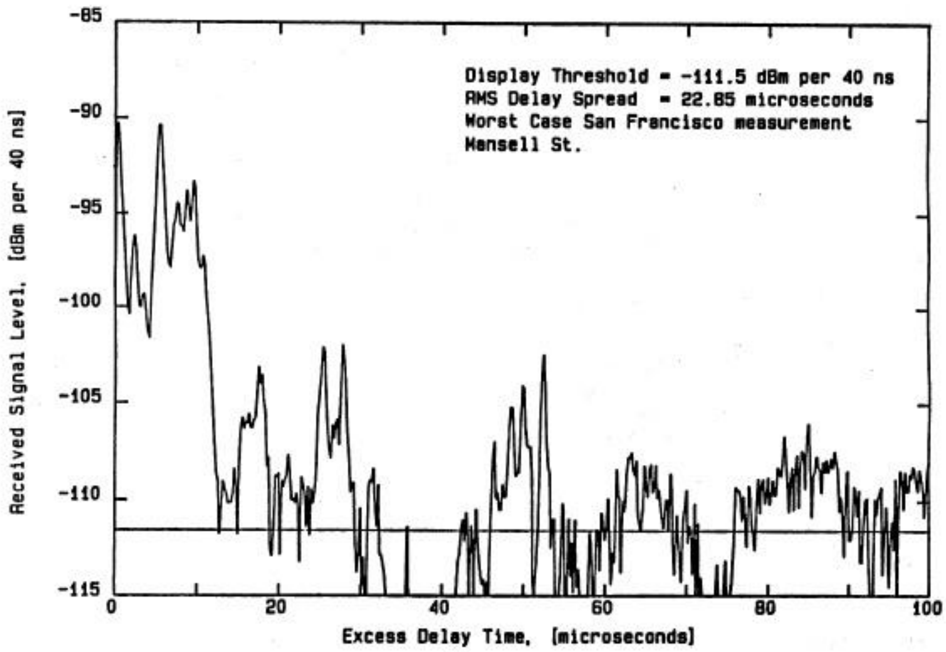
$$\bar{\tau} = \frac{\int_o^\infty \tau Ac(\tau)\, d\tau}{\int_o^\infty Ac(\tau)\, d\tau}$$

2.11

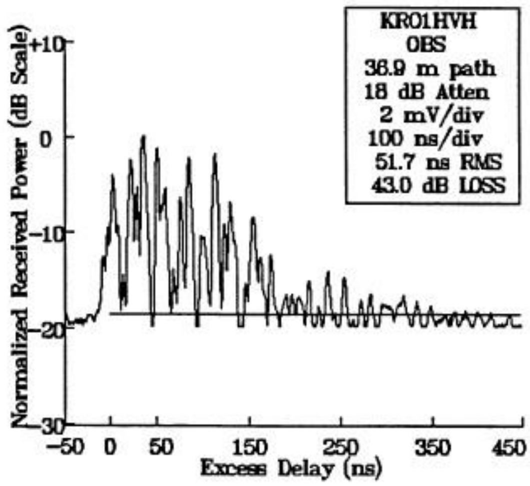
Thus, the rms delay spread is:

$$\tau_{rms} = \sqrt{\frac{\int_o^{\infty} (\tau - \bar{\tau})^2 Ac(\tau)\, d\tau}{\int_o^{\infty} Ac(\tau)\, d\tau}} \qquad 2.12$$

- The coherence bandwidth is the range of frequencies over which the amplitude of two spectral components of the channel response is correlated. The coherence bandwidth provides a measurement of the range of frequencies over which the channel shows a flat frequency response, in the sense that all the spectral components have approximately the same amplitude and a linear change of phase. This means that if the transmitted signal bandwidth is less than the channel coherence bandwidth, then all the spectral components of the signal will be affected by the same attenuation and by a linear change of phase. In this case, the channel is said to be a flat fading channel. In another way, since the signal sees a channel with flat frequency response, the channel is often called a narrowband channel. If on the contrary, the transmitted signal bandwidth is more than the channel coherence bandwidth, then the spectral components of the signal will be affected by different attenuations. In this case, the channel is said to be a frequency selective channel or a broadband channel.




(a)




(b)

**Fig. 6 The power delay profile of a typical wireless channel [58]**

### 2.5.3 Doppler Spread, Coherence Time and Time Selectivity

One key parameter in adaptation of mobile radio systems is the maximum Doppler spread. It provides information about the fading rate of the channel. Knowing Doppler spread in mobile communication systems can improve detection and help to optimize transmission at the physical layer as well as higher levels of the protocol stack.

Specifically, knowing Doppler spread can decrease unnecessary handoffs, adjust interleaving lengths to reduce reception delays, update rate of power control algorithms, etc. Because of multipath reflections, the channel impulse response of a wireless channel looks likes a series of pulses. Each propagation path is associated with a Doppler frequency that depends on the angle between it and at direction of the UT. The Doppler frequency is given by

$$f_{d,p,l} = \frac{v}{\lambda} cos\ (\varphi_{p,l}) \tag{2.13}$$

Where $v$ is the mobile velocity, $\lambda$ is the wavelength and $\varphi_{p,l}\ \epsilon\ [0; \pi]$ is the angle that each sub path makes with the direction of the UT; when it is 0º, it means that we have a wave coming from the opposite direction as the direction antenna, we obtain the maximum Doppler shift and it is done by

$$f_d = \frac{v}{\lambda} \tag{2.14}$$

When the signal bandwidth is much larger than the channel coherence bandwidth (bigger then Doppler spread), the channel is frequency selective. These channels present such impairments as inter-symbol interference, which deform the shape of the transmitted pulse, risking the introduction of detection errors at the receiver. In fig.7 we can see the frequency spectrum of a signal with the Doppler shift effect.

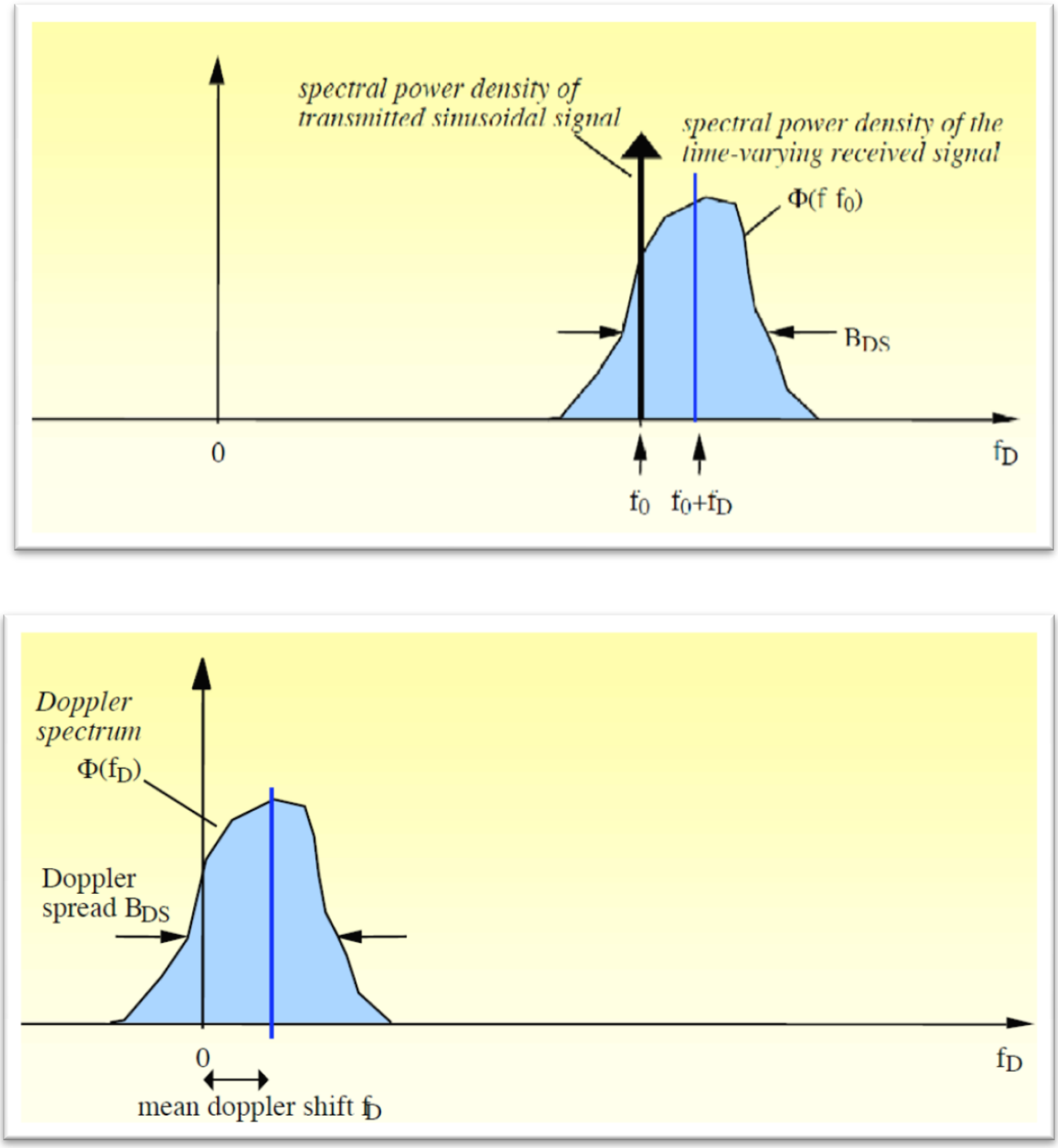



**Fig. 7 Frequency spectrum of the signal**

The coherence time of the channel $t_c$ is the duration over which the channel characteristics can be considered as time invariant [9]. It is of utmost importance for evaluating and adapting the techniques that try to exploit time diversity of the channel or to compensate its time selectivity [10]. Time diversity can be exploited if the separation between time slots carrying the same information exceeds the coherence time. For an OFDM transmission system, the coherence time has a special significance because the symbol duration should be chosen to be much smaller than $t_c$ to avoid ICI. In case the duration of the transmitted symbol exceeds the coherence time, we will have symbols interference. Although, there is no specific definition for the coherence time, it is known to be inversely proportional to the maximum Doppler frequency [11], i.e.,

$$t_c \approx \frac{1}{f_d} \tag{2.15}$$

The computation of the coherence time can be easily obtained from the ACF (auto Correlation Function) of the available channel estimates if a certain threshold is attained within an amount of time lags. Commonly, when the time elapsed for ACF to drop to half of its maximum zero lag value, this can be regarded as a measure of the coherence time. It is obvious that our proposed method will also allow a reliable estimate of the coherence time due to its immunity to noise and ISI perturbations.

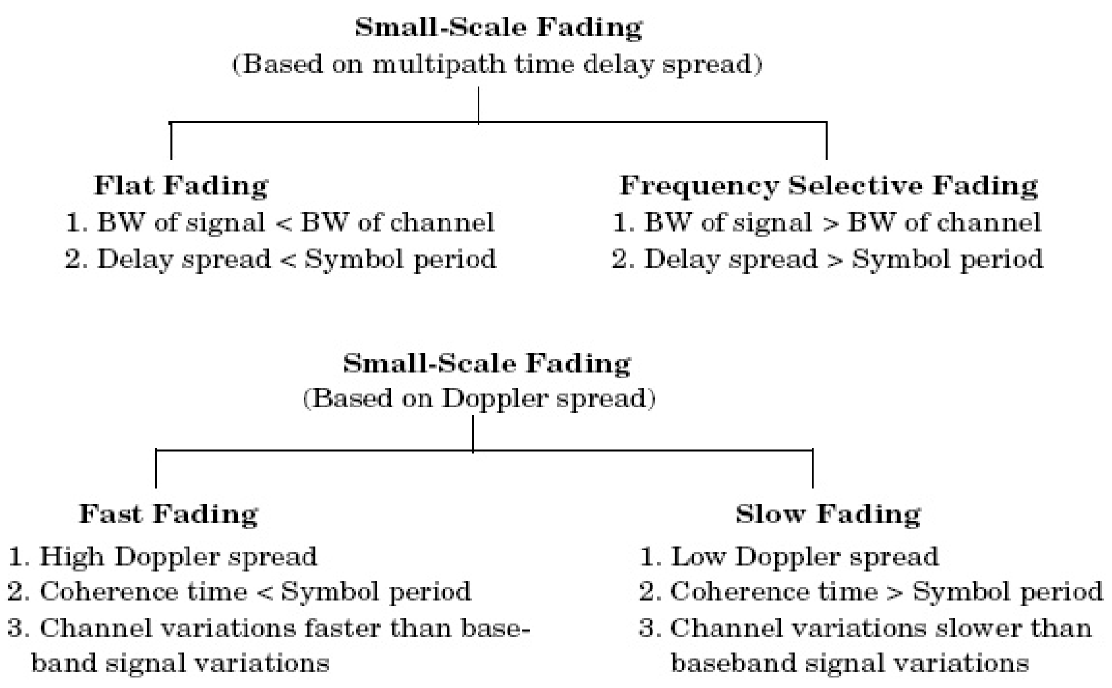



**Fig. 8 Fading summary scheme [12]**

# Chapter 3

## 3 Multicarrier Systems

As wireless communication evolves towards broadband systems to support high data rate applications, we need a technology that can efficiently handle frequency-selective fading. The OFDM system is widely used in this context. The pioneering work on OFDM was first started in the 60's [13] [14]. The key idea of OFDM is to divide the whole transmission band into a number of parallel sub channels (also called subcarriers) so that each sub channel is flat fading channel [15] [16]. In this case, channel equalization can be performed in all sub channels in parallel (fig.9), using simple one-tap equalizers, which have very small computational complexity.

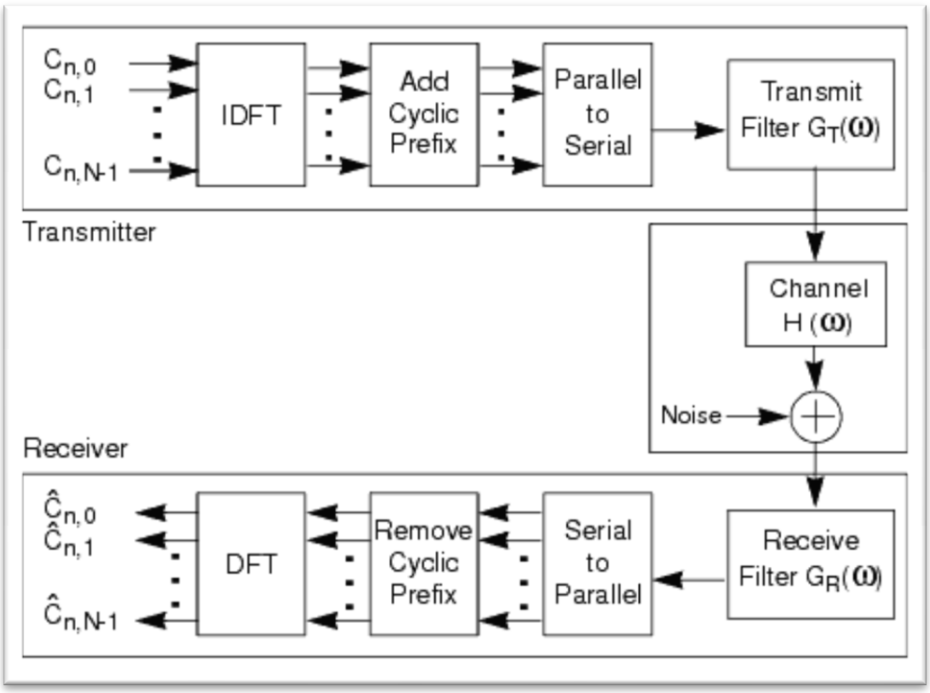



**Fig. 9 General Block diagram of an OFDM system [45]**

## 3.1 Basic Principles of OFDM

Such we discussed previously, when the signal bandwidth is much larger than then channel coherence bandwidth, the channel is frequency selective. We also explained that these channels present such impairments as inter-symbol interference, which deform the shape of the transmitted pulse, risking the introduction of detection errors at the receiver. This impairment can be addressed with different techniques. One of these techniques is multicarrier modulation.

The basic idea of multicarrier modulation is to divide the transmitted bit stream into many different sub-streams and send these over many different sub-channels, as shown in fig.9 and fig. 11. Typically the sub-channels are orthogonal under ideal propagation conditions. The data rate on each of the sub-channels is much less than the total data rate, and the corresponding sub-channel bandwidth is much less than the total system bandwidth. The number of sub-streams is chosen to ensure that each sub-channel has a bandwidth less than the coherence bandwidth of the channel, so the sub-channels experience relatively flat fading. Thus, the inter-symbol interference on each sub-channel is small. The sub-channels in multicarrier modulation need not be contiguous, so a large continuous block of spectrum is not needed for high rate multicarrier communications. Moreover, multicarrier modulation is efficiently implemented digitally. In this discrete implementation, called OFDM the ISI can be completely eliminated through the use of a cyclic prefix (fig. 12). A cyclic prefix is a repetition of the first section of a symbol that is appended to the end of the symbol.

The simplest form of multicarrier modulation divides the data stream into multiple sub-streams to be transmitted over different orthogonal sub-channels centered at different subcarrier frequencies. The number of sub-streams is chosen to make the symbol time on each sub-stream much greater than the delay spread of the channel or, equivalently, to make the sub-stream bandwidth less than the channel coherence bandwidth. This ensures that the sub-streams will not experience significant ISI. Consider a linearly modulated system with data rate $R$ and bandwidth $B$. The coherence bandwidth for the channel is assumed to be $B_C < B$, so the signal experiences frequency selective fading. The basic premise of multicarrier modulation is to break this wideband system into $N$ linearly modulated subsystems in parallel, each with sub-channel bandwidth $B_N = B/N$ and data rate $R_N \approx R/N$. For $N$ sufficiently large, the sub-channel bandwidth $B_N = B/N < B_C$, which ensures relatively flat fading on each sub-channel. This can also be seen in the time domain: the symbol time $T_s$ of the modulated signal in each sub-channel is proportional to the sub-channel bandwidth 1/BN. So $B_N$ « $B_C$ implies that $T_s$ » $\mathrm{T_M}$, where $\mathrm{T_M}$, denotes the delay spread of the channel. Thus, if $N$ is sufficiently large, the symbol time is much greater than the delay spread, so each sub-channel experiences little ISI degradation. In a multicarrier transmitter, the bit stream is divided into $N$ sub-streams via a serial-to-parallel converter. The nth sub-stream is linearly modulated (typically via QAM or PSK) relative to the subcarrier frequency $f_n$ and occupies bandwidth $B_N$. We assume coherent demodulation of the subcarriers so the subcarrier phase is neglected in our analysis. If we assume raised cosine pulses for $g(t)$ we get a symbol time $T_s = (1 + \beta)/B_N$ for each sub-stream, where $\beta$ is the roll off factor of the pulse shape. The resulting OFDM modulated symbols associated with all the sub-channels are summed together to form the transmitted signal, so for $0 \le t \le T_s$ we have

$$S\,(t) = \sum_{K=0}^{N-1} d_k \varphi_k(t) \; = \sum_{K=0}^{N-1} d_k e^{j2\pi f_k t}, \tag{3.1}$$

Where $f_k = f_0 + k\Delta f$ and $\Delta f = \frac{1}{T_s}$ and signals $\varphi_k(t)$ are defined as

$$\varphi_k(t) = \begin{cases} e^{j2\pi f_k t} & if\ 0 \le t \le T_s \\ 0 & else, \end{cases} \tag{3.2}$$

Form an orthonormal set that is used as the carrier signal of each subcarrier in this multicarrier modulation technique. Because these signals are truncated complex exponential, in frequency domain they are of the form $sin(x)/x$.

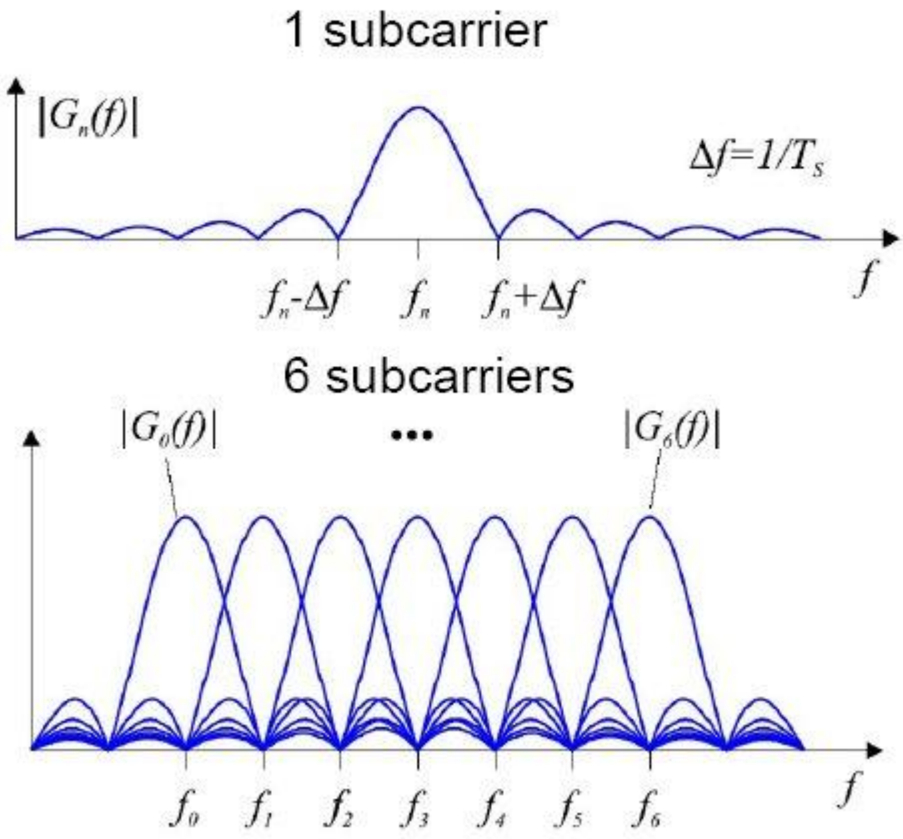



**Fig. 10 OFDM orthonormal set of 6 modulation signals in frequency domain [17]**

Then the OFDM symbol is sampled with a period $T_{SA} = \frac{T_S}{N}$, and the resulting sampled signal $s[n]$ is,

$$S[n] = s(nT_{SA}) = \sum_{K=0}^{N-1} d_k e^{j2\pi f_k nT_S/N},\ 0 \leq n \leq N-1. \qquad (3.3)$$

In the receiver, each sub-stream is passed through a narrowband filter (to remove the other sub-streams), demodulated, and combined via a parallel-to serial converter to form the original data stream. Note that the *i-th* sub-channel will be affected by flat fading corresponding to a channel gain $\alpha_i = H(f_i)$.

Although this simple type of multicarrier modulation is easy to understand, it has several significant shortcomings. First, in a realistic implementation, sub-channels will occupy a larger bandwidth than under ideal raised cosine pulse shaping because the pulse shape must be time limited. Let $\varepsilon/T_N$ denote the additional bandwidth required due to time limiting of these pulse shapes. The sub-channels must then be separated by $(1+\beta+\varepsilon)/T_s$, and since the multicarrier system has $N$ sub-channels, the bandwidth penalty for time limiting is $\varepsilon N/T_s$. In particular, the total required bandwidth for non overlapping sub-channels is,

$$B = \frac{N(1+\beta+\varepsilon)}{T_s} \tag{3.4}$$

Thus, this form of multicarrier modulation can be spectrally inefficient. Additionally, near ideal (and hence expensive) low-pass filters will be required to maintain the orthogonality of the subcarriers at the receiver. Perhaps most importantly, this scheme requires $N$ independent modulators and demodulators, which entails significant expense, size, and power consumption. It is possible a modulation method that allows subcarriers to overlap and removes the need for tight filtering, and a discrete implementation of multicarrier modulation, which eliminates the need for multiple modulators and demodulators.

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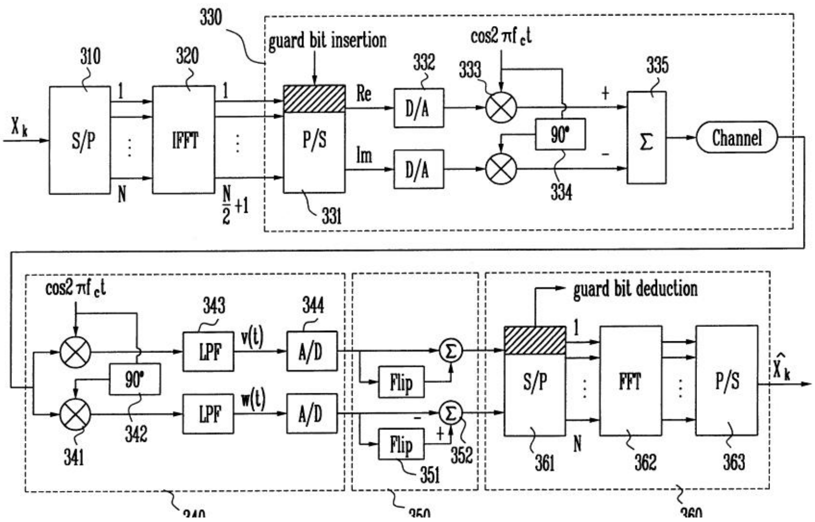


Fig. 11 Multicarrier transmitter and receiver scheme detailed [57]

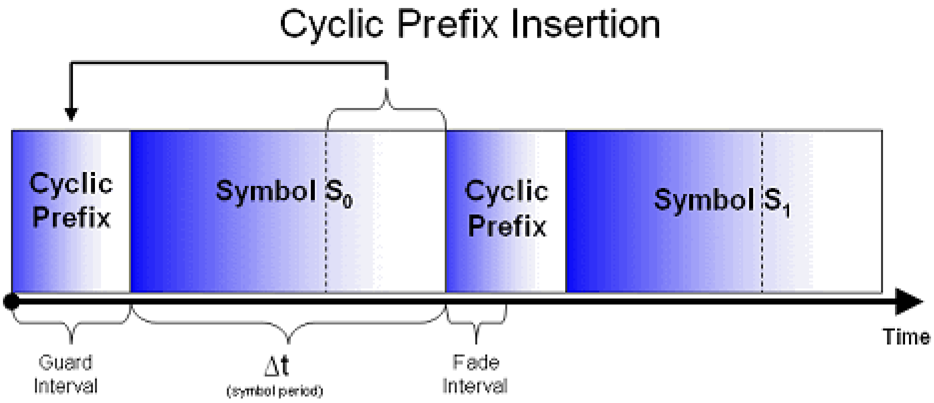


Fig. 12 A cyclic prefix illustration [55]

## 3.2 OFDM applied on LTE

The LTE downlink uses OFDMA, which is particularly robust when handling the varying propagation conditions seen in mobile radio. So the DL (Direct Link) transmission scheme used for LTE is based on conventional OFDM systems where the dedicated spectrum is divided into multiple orthogonal carriers to each other. Each of these sub-carriers is independently modulated using a low rate data stream. For a bandwidth of 5 MHz and according to the technical recommendation of 3GPP (TR 25.892 [18]), the OFDM concept is illustrated in fig. 13.

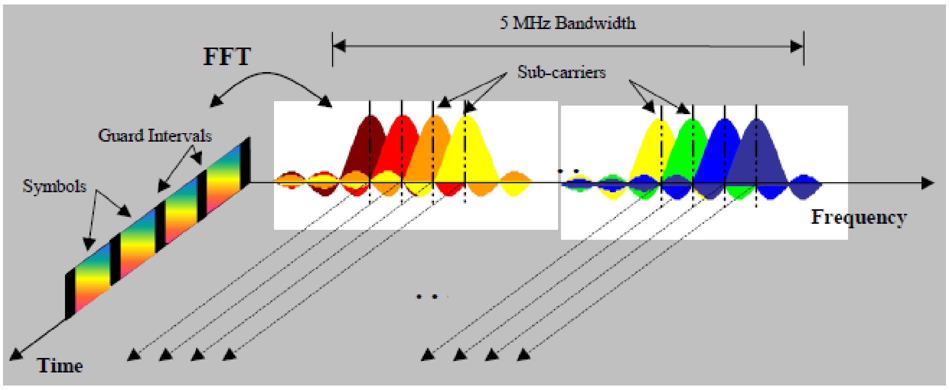


Fig. 13 General frequency-time representation of a OFDM symbol [19]

Fig 14 shows a radio frame which has duration of 10 ms and consists of 20 slots with slot duration of 0.5ms, numbered from 0 to 19. Two adjacent slots form one sub-frame of length 1 ms [20].

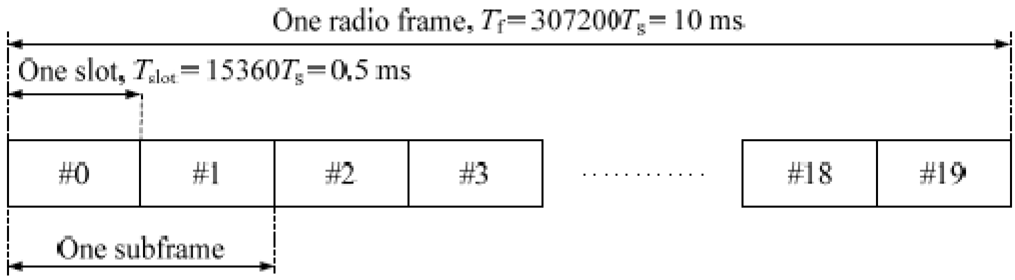


**Fig. 14 Frame structure of a LTE standard**

The structure of the downlink resource grid for the duration of one downlink slot is shown in fig. 15. The transmitted signal in each slot is described by a resource grid of $N_{RB}^{DL}N_{SC}^{RB}$ subcarriers with a spacing of Δf = 15 kHz and $N_{Symb}^{DL}$ OFDM symbols. The quantity $N_{RB}^{DL}$ depends on the downlink transmission bandwidth configured in the cell and shall fulfill the following condition: $6 \leq \mathrm{N}_{\mathrm{RB}}^{\mathrm{DL}} \leq 110$. Each element in the resource grid is called a resource element which is the smallest time-frequency unit for downlink transmission and is equal to one subcarrier per one OFDM symbol.

The downlink slot for the generic frame structure with normal cyclic prefix length contains 7 OFDM symbols. This translates into a cyclic prefix length of 5.2 μs for the first symbol and 4.7 μs for the remaining 6 symbols. Data is allocated to the UTs in terms of resource blocks. A physical resource block (RB) consists of 12 consecutive sub-carriers in the frequency domain for the Δf = 15 kHz case. In the time domain, a physical resource block consists of 7 consecutive OFDM symbols. Depending on the required data rate, each UT can be assigned one or more resource blocks in each transmission time interval of 1 ms (one sub-frame composed by two consecutive slots). The scheduling decision is done in the base station. The user data or payload is carried on the Physical Downlink Shared Channel (PDSCH). Downlink control signaling on the PDCCH is used to convey the scheduling decisions to individual UTs. The PDCCH is located in the first OFDM symbols of a slot.

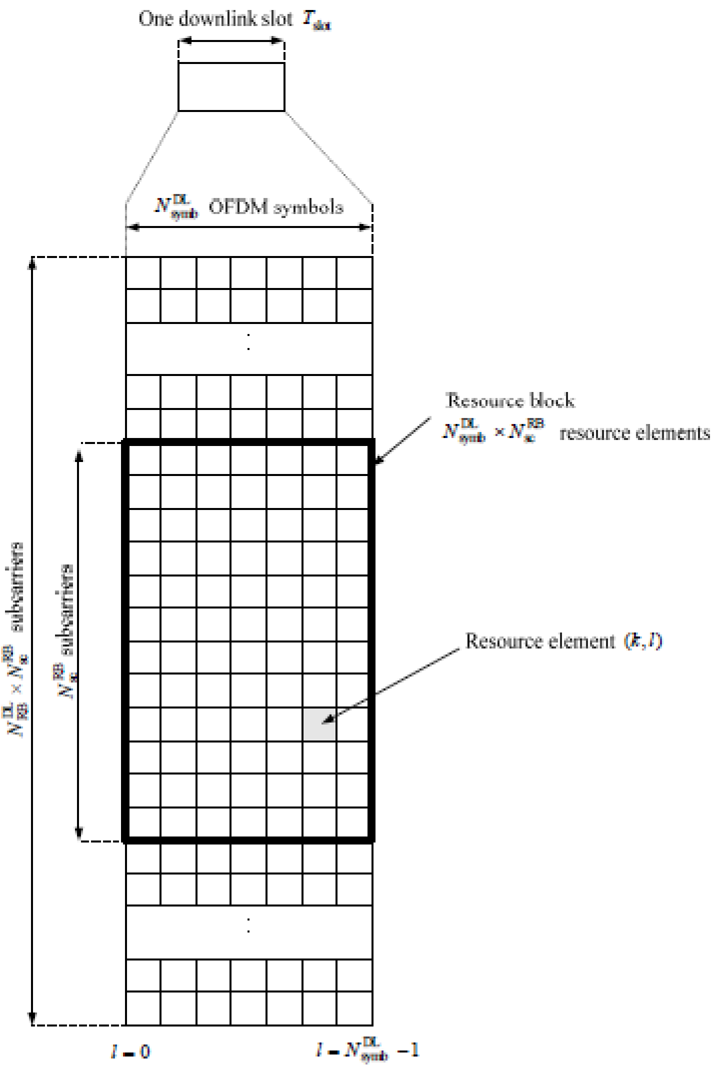


Fig. 15 Resource grid for downlink transmission

# Chapter 4

# 4 Multi-Antenna Techniques

To meet the requirements of future radio communication system, multiple antennas should be employed at BS and/or UT. Such system can exploit the spatial dimension of the mobile radio channel basically in three different ways: as an additional source of diversity, for spatial multiplexing, and for a spatial separation of users (precoding or beam-forming) techniques. In this chapter we will focus on the diversity techniques, namely the space-time/frequency block codes schemes. In chapter 6, we will extend these techniques to a relay-assisted based system.

## 4.1 Diversity

### 4.1.1 Multipath Diversity

In order to understand diversity, we first must understand multipath distortion. When a radio frequency (RF) signal is transmitted towards the receiver, the general behavior of the RF signal is to grow wider as it is transmitted further. On its way, the RF signal encounters objects that reflect, refract, diffract or interfere with the signal. When an RF signal is reflected off an object, multiple wave fronts are created. As a result of these new duplicate wave fronts, there are multiple wave fronts that reach the receiver.

Multipath propagation occurs when RF signals take different paths from a source to a destination. A part of the signal goes to the destination while another part bounces off an obstruction, then goes on to the destination. As a result, part of the signal encounters delay and travels a longer path to the destination. Multipath can be defined as the combination of the original signal plus the duplicate wave fronts that result from reflection of the waves off obstacles between the transmitter and the receiver.

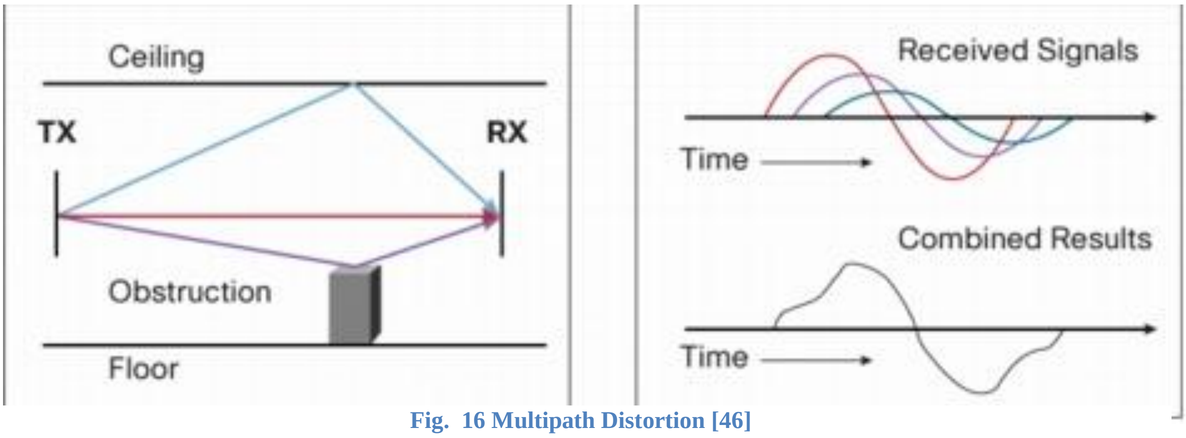


Fig. 16 Multipath Distortion [46]

Multipath distortion is a form of RF interference that occurs when a radio signal has more than one path between the receiver and the transmitter. Fig.16 shows an example of multipath distortion and his effect in final received signal results. This occurs in cells with metallic or other RF–reflective surfaces, such as furniture, walls, or coated glass. This phenomenon could results in a data corruption, signal nulling or increased or decreased signal amplitude. In a multipath environment, signal null points are located throughout the area. The distance an RF wave travels, how it bounces, and where the multipath null occurs are based on the wavelength of the frequency. As frequency changes, so does the length of the wave. Therefore, as frequency changes, so does the location of the multipath null.

### 4.1.2 Macro Diversity

Macro diversity creates antenna diversity by utilizing the network in a efficient manner. A signal transmitted by a mobile station in uplink propagates to multiple base stations, and since the channel coefficients to each base station are independent, the signal combined over all base stations enjoys diversity. On the other hand, due to limited bandwidth in the fixed network between the base stations, optimal signal combining (in the spirit of diversity antennas) is not feasible. Nevertheless, at least selection-type combining is possible, in the sense that it is sufficient to receive the transmitted signal correctly in at least one base station. In downlink, multiple copies of the same signal are transmitted from spatially separate source locations (the base stations), again to result in independent fading at the mobile station. The specification includes also a feedback-based macro diversity option, called SSDT (Site Selection Diversity Transmission). SSDT attempts to mitigate interference to other users in the system by more optimal power allocation across cells. Thus, it is essentially an antenna selection concept combined with trivial power allocation and improves both diversity and power efficiency provided that the feedback signaling is up-to-date. In SSDT cells are assigned a temporary identification (ID). The UT periodically informs the ID of a primary cell to the base stations using an uplink (feedback) signaling field. The dedicated channel in other cells (called non-primary cells) is turned off. The ID of the primary cell is signaled 1-5 times in 10 ms frame, depending on the selected signaling formats.

### 4.1.3 Time Diversity

It is quite common to find communication scenarios where the channel coherence time equals or exceeds several symbol transmission periods. This implies that two symbols transmitted with a separation in time longer than the coherence time will experience channel realizations that are highly uncorrelated and can be used to obtain diversity. The simplest way to achieve this is to form the two symbols by using a repetition coding scheme. Also, to guarantee that the repeated symbols will be transmitted over uncorrelated channel realization, an appropriate interleave is applied to the stream of symbols to be transmitted. At the receiver, the copies of the symbol will have to be combined together. As explained above, if the transmission of each symbol can be represented by an input–output expression of the form,

$$y_i = h_i\, x + n_i \tag{4.1}$$

where $x$ is the unit-energy transmitted symbol, $y_i$ is the symbol received over path $i$ , $n_i$ is the background noise modeled as a circularly symmetric Gaussian random variable with zero mean and variance $N_0$, and $h_i$ is the channel realization over path $i$ , assumed to be following a Rayleigh fading, optimal combining that maximizes the received SNR is achieved with a maximal ratio combiner. Recall that the SNR at the output of an MRC(Maximum Ratio Combiner) equals the sum of the SNRs of the branches at the input of the MRC. From the explanation above on Rayleigh fading, the MRC output SNR will have a Chi-squared distribution.

### 4.1.4 Frequency Diversity

Analogous to time diversity, in those wideband systems where the available bandwidth exceeds the channel coherence bandwidth, it is possible to realize diversity by using channels that are a partition of the available bandwidth and that are separated by more than the channel coherence bandwidth. Realizing frequency diversity as a partition of the whole system bandwidth into channels with smaller bandwidth and independent frequency response is perhaps the most intuitively natural approach. This approach is applicable in multicarrier systems, where transmission is implemented by dividing the wideband channel into non-overlapping narrowband sub-channels. The symbol used for transmission in each sub-channel has a transmission period long enough for the sub-channel to appear as a flat fading channel. Different sub-channels are used together to achieve frequency diversity by ensuring that each is separated in the frequency domain from the rest of the sub-channels in the transmission by more than the coherence bandwidth. In this way, the fading processes among the sub-channels will show a small cross-correlation. An example of these systems is that using OFDM.

### 4.1.5 Received Antenna Diversity

The number of receive antennas one wishes to deploy is typically an implementation issue. When multiple receive antennas are used we say that the receiver uses receive (Rx) antenna diversity. Rx diversity may be used in the base station to improve uplink capacity or coverage. Due to cost and space considerations multi-antenna reception is not popular in terminals. However, Rx diversity is one of the most efficient diversity techniques and likely to be used when performance or coverage improvements are desired.

### 4.1.6 Transmit Diversity

A significant effort has been devoted in 3GPP to develop efficient transmit diversity solutions to enhance downlink capacity. Transmit diversity methods provide space diversity for terminals with only one receive antenna, and improve the link performance while retaining the complexity at the base station. Typically, the transmitting antenna elements are relatively close to each other. In this case the delay profile is essentially the same for each transmitting element. The closed loop transmits (Tx) diversity solutions developed for the FDD (Frequency Division Duplex) mode support two transmit antennas. Both open-loop and closed-loop Tx diversity solutions are specified for UTRA FDD and TDD modes. The first open-loop concepts proposed in 3G standardization were based on Code Division Transmit Diversity (Orthogonal Transmit Diversity) and Time Switched Transmit Diversity [38]. TSTD (Time Switched Transmit Diversity) is applied in the WCDMA standard for certain common channels. In TSTD the transmitted signal hops across two transmit antennas, according to [21]. Eventually, also a more efficient STTD (Space Time Transmit Diversity) solution, based on a variant of the space–time block code developed by Alamouti, was adopted, where column 1 is transmitted from antenna 1 and column 2 from antenna 2.

The first feedback mode proposed to 3G systems was based on STD (selective transmit diversity), where only one additional feedback bit is used per feedback slot to select the desired transmit antenna. These contributions sparked the research on feedback modes, and a number of improvements were eventually suggested in 3G standardization. Currently, the WCDMA Release ’99 and Release 4 specifications include two closed-loop transmit diversity concepts. In both concepts co-phasing information, signaled using a fast feedback channel (of rate 1500 bps), is applied in selecting one of 4 or 16 possible beam weights, respectively. Both concepts approximate coherent transmission (channel-matched beam-forming) using different channel quantization and feedback signaling strategies.

## 4.2 Multiple antenna configurations

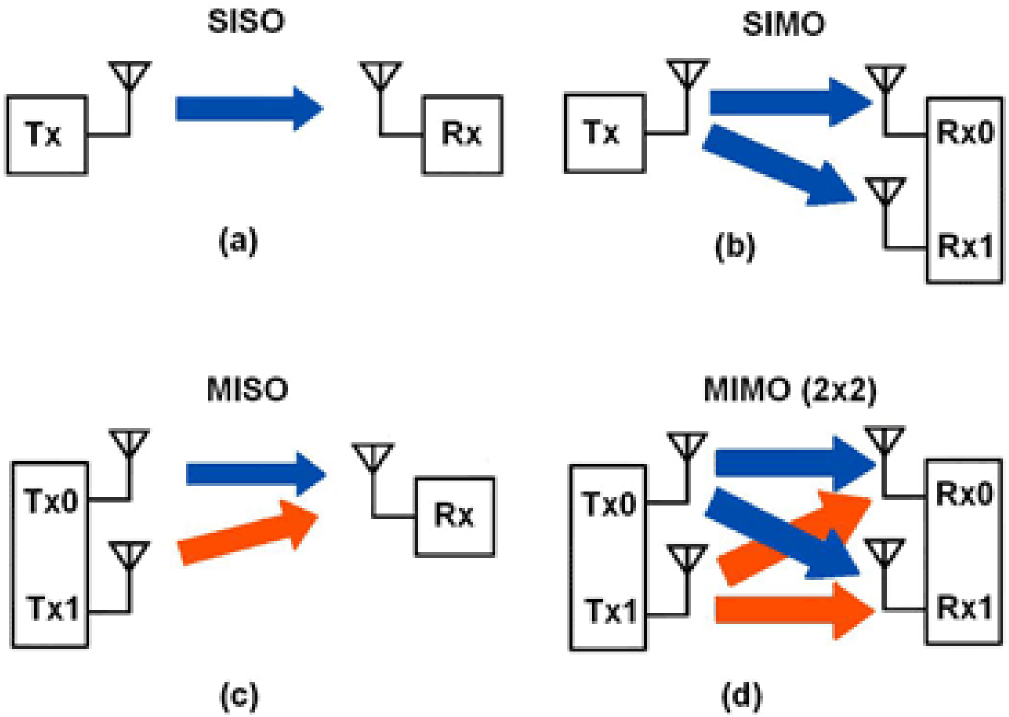


Fig. 17 Multiple antenna configurations

The more basic wireless transmission configuration is known as SISO (single input single output), fig17.a).. For a SISO system with a deterministic channel $h$, the received signal can be written only as $r = hs + n$, where $s$ is the transmitted symbol with symbol energy $Es$ and n is the zero mean AWGN (Additive Gaussian Noise) noise with power spectrum density $N_0$. However, in wireless systems, transmission failures occur mostly when the channel is deep fade, resulting in the so called communication outage. To overcome this effect, one of the necessity that appears was to exploit different kinds of diversity techniques in space, and this diversity can be exploited then the transmitter or the receiver have multiple antennas. Considering a flat-fading scenario, where no inter-symbol interference occurs over time, and assume that the antennas are located sufficiently apart from each other so that channel coefficients between different transmitter and receiver antennas are statistically independent. Spatial diversity gains can be achieved with either precoding at the transmitter or signal combining at the destination. In the following, we introduce these techniques for the tree different scenarios, namely, single-input multiple-output (SIMO) (fig. 17 b)), multiple-input single-output (MISO) (fig.17 c)), and multiple-input single-output (MIMO) (fig.17 d)). These techniques can also be extended to systems with diversity in other dimensions, such as the case of multipath or frequency selectivity.

When the receiver is equipped with multiple antennas, we can take advantages of spatial diversity at the receiver to enhance systems performance. Consider the system where a single-antenna transmitter sends data to a receiver with Nr antenna (SIMO). Let $x[n]$ be the symbol transmitted in the *n-th* symbol period and assume that $E[|x[n]|^2] = 1$. The signal received at *k-th* antenna of the receiver in *n-th* symbol period can be expressed as

$$\boldsymbol{y}_k[n] = \sqrt{P} h_k \boldsymbol{x}[n] + \boldsymbol{w}_k[n], \tag{4.2}$$

Where $P$ is the transmit power, $h_k$ is the channel coefficient observed by the *k-th* receive antenna, and $w_k[n] \sim \mathcal{CN}(0, \sigma_k^2)$ is the AWGN at the *k-th* antenna. The channel coefficient $h_k$ can be written in terms of its amplitude $|h_k|$ and phase $\Phi_k$ so thet, for $k = 1,2,..,Nr$,

$$h_k = |h_k| e^{j\Phi_k} \tag{4.3}$$

The SNR at the *k-th* antenna is then defined as

$$\Upsilon k \triangleq \frac{P|h_k|^2}{\sigma_k^2} \tag{4.4}$$

Suppose that the instantaneous channel state information (CSI), i.e., the state of channel coefficients $\{h_1, h_2, .., h_{N_r}\}$, is known at the receiver. Before performing signal detection, the receiver will linearly combine the received symbols $\boldsymbol{y}_1[n], \boldsymbol{y}_2[n], \dots, \boldsymbol{y}_{N_r}[n]$ with the respective weighting factors $\alpha_1, \alpha_2, .., \alpha_{N_r}$, to obtain the signal,

$$\boldsymbol{Z}[n] = \sum_{k=1}^{N_r} \boldsymbol{\alpha}_k \boldsymbol{y}_k[n] \tag{4.5}$$

When the transmitter is equipped with multiple antennas, the data symbols can be distributed among multiple transmit antennas to exploit spatial diversity at the transmitter. Here we consider a system with $N_t$ antennas at the transmitter and only a single antenna at the receiver. Let $\{ x[n]\}$ be the sequence of data to be transmitted and assume that the data sysmbols are i.i.d. over time with zero mean and unit variance. Depending on the specific transmit diversity scheme, the data is first pre-processed to form a sequence of transmit symbol vectors $\{s[n]\}$, where $\boldsymbol{s}[n] = [\boldsymbol{s}_1[n], \boldsymbol{s}_2[n], .., \boldsymbol{s}_{N_t}[n]]^T$ is the vector of symbols to be transmitted over the Nt antennas in the n-th symbol period. The transmitted symbols are assumed to satisfy the sum power constraint

$$\boldsymbol{E}[|s|^2] = \sum_{k=1}^{N_t} \boldsymbol{E}[|\boldsymbol{s}_k[n]|]^2 \leq 1 \tag{4.6}$$

The signal obtained at the receiver during the *n-th* symbol period is given by

$$\boldsymbol{y}[n] = \sum_{k=1}^{N_t} \sqrt{P}\boldsymbol{h}_k \boldsymbol{s}_{\boldsymbol{k}}[n] + \boldsymbol{w}[n] \tag{4.7 }$$

Where $P$ is the transmit power, $h_k \sim C(0, \sigma_h^2)$ is the channel coefficient between the $k$-$th$ transmit antenna and the receiver, and $\boldsymbol{w}[n]$ is the AWGN with zero mean and variance $\sigma_w^2$.

MIMO systems can be defined simply. Given an arbitrary wireless communication system, we consider a link for which the transmitting end as well as the receiving end is equipped with multiple antenna elements. Such a setup is illustrated in Fig. 1.d. The idea behind MIMO is that the signals on the Tx antennas at one end and the Rx antennas at the other end are “combined” in such a way that the quality (bit-error rate or BER) or the data rate (bits/sec) of the communication for each MIMO user will be improved. Such an advantage can be used to increase both the network’s quality of service and the operator’s revenues significantly. A core idea in MIMO systems are space–time signal processing in which time (the natural dimension of digital communication data) is complemented with the spatial dimension inherent in the use of multiple spatially distributed antennas. As such MIMO systems can be viewed as an extension of the so-called smart antennas, a popular technology using antenna arrays for improving wireless transmission dating back several decades. MIMO effectively takes advantage of random fading and when available, multipath delay spread], for multiplying transfer rates [22]. The prospect of many orders of magnitude improvement in wireless communication performance at no cost of extra spectrum (only hardware and complexity are added) is largely responsible for the success of MIMO as a topic for new research. This has prompted progress in areas as diverse as channel modeling, information theory and coding, signal processing, antenna design and multi antenna hardware cellular design, fixed or mobile.

## 4.3 Space-Time Block Coding Schemes

In this we present the main space-time block codes extended to the proposed relay-assisted based schemes.

### 4.3.1 Alamouti transmition

A very simple and effective coding scheme for two transmit antennas and a single receive antenna has been introduced by Alamouti [23]. The data sequence $x_1$; $x_2$* is sent over the first antenna and $x_2$ -$x_1$* is sent over the second antenna, where * denotes complex conjugation.

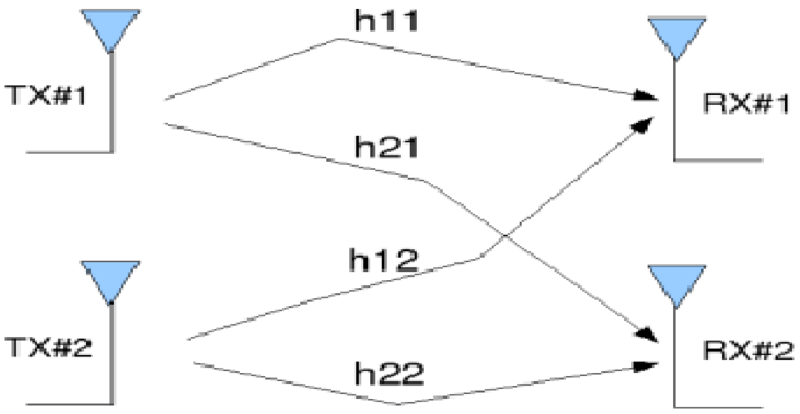


Fig. 18 . Alamouti 2x2 scheme [47]

**Transmitter:**

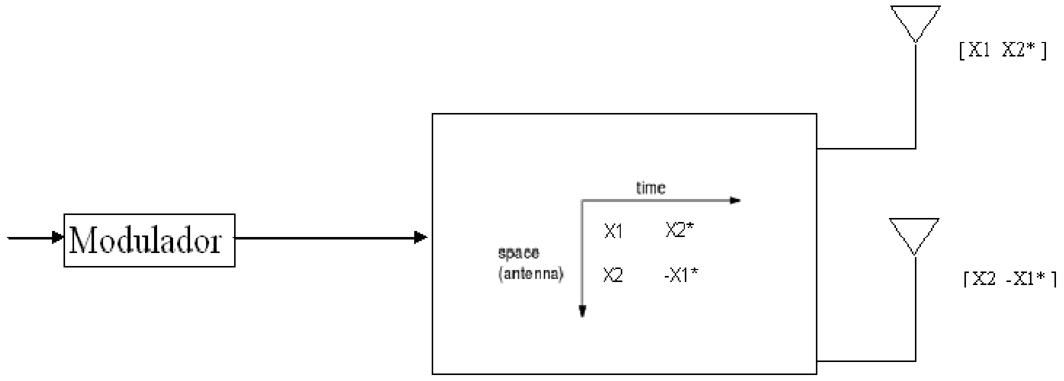


Fig. 19 Alamouti two antennas codification

The signal is modulated and then transmitted by the two conveying antennas encoded by the alamouty codification; The symbols are encoded two by two and transmitted in two different frequencies or two different time slots depending if the code is done in time or frequency. In fig.19 we show the signal transmitted in different times.

**Reception:**

At time *t*, the signal $Y_t^j$ received at antenna *j* is:

$$Y_t^j = \sum_{i=1}^{Nt} h_{i,j} * x_t^i + n_t^i \tag{4.8}$$

Where $h_{i,j}$ is the path gain from transmit antenna *i* to receive antenna *j*. $x_t^i$ is the signal transmitted by transmit antenna *i* and $n_t^i$ is a sample of AWGN. Assuming two transmit and receive antennas the signals at the receiver can be given by,

$$Y_1^1 = \left(\frac{1}{\sqrt{2}}\right) * \left(h_{1,1}x_1 - h_{2,1}x_2^*\right) + n_1^1 \tag{4.9}$$

$$Y_2^1 = \left(\frac{1}{\sqrt{2}}\right) * \left(h_{1,1}x_2 + h_{2,1}x_1^*\right) + n_2^1 \tag{4.10}$$

$$Y_1^2 = \left(\frac{1}{\sqrt{2}}\right) * \left(h_{1,2}x_1 - h_{2,2}x_2^*\right) + n_1^2 \tag{4.11}$$

$$Y_2^2 = \left(\frac{1}{\sqrt{2}}\right) * \left(h_{1,2}x_2 - h_{2,2}x_1^*\right) + n_2^2 \tag{4.12}$$

To recover the data we apply the follow formulates:

$$\mathrm{d}_1^1 = \left(\frac{1}{\sqrt{2}}\right) * \left(\mathrm{h}_{1,1}^*\mathrm{Y}_1^1 - \mathrm{h}_{2,1}\mathrm{Y}_2^{1*}\right) \tag{4.13}$$

$$\mathrm{d}_2^1 = \left(\frac{1}{\sqrt{2}}\right) * \left(-\mathrm{h}_{2,1}\mathrm{Y}_1^{1*} - \mathrm{h}_{1,1}^*\,\mathrm{Y}_2^1\right) \tag{4.14}$$

$$d_1^2=\left(\frac{1}{\sqrt{2}}\right)*\left(h_{1,2}Y_1^2 - h_{2,2}Y_2^2\right) \tag{4.15}$$

$$d_2^2=\left(\frac{1}{\sqrt{2}}\right)*\left(-h_{2,1}Y_1^{2*}+h_{1,1}Y_2^2\right) \tag{4.16}$$

### 4.3.2 Quasi-Orthogonal Space-Time Block Codes for Four Transmit Antennas

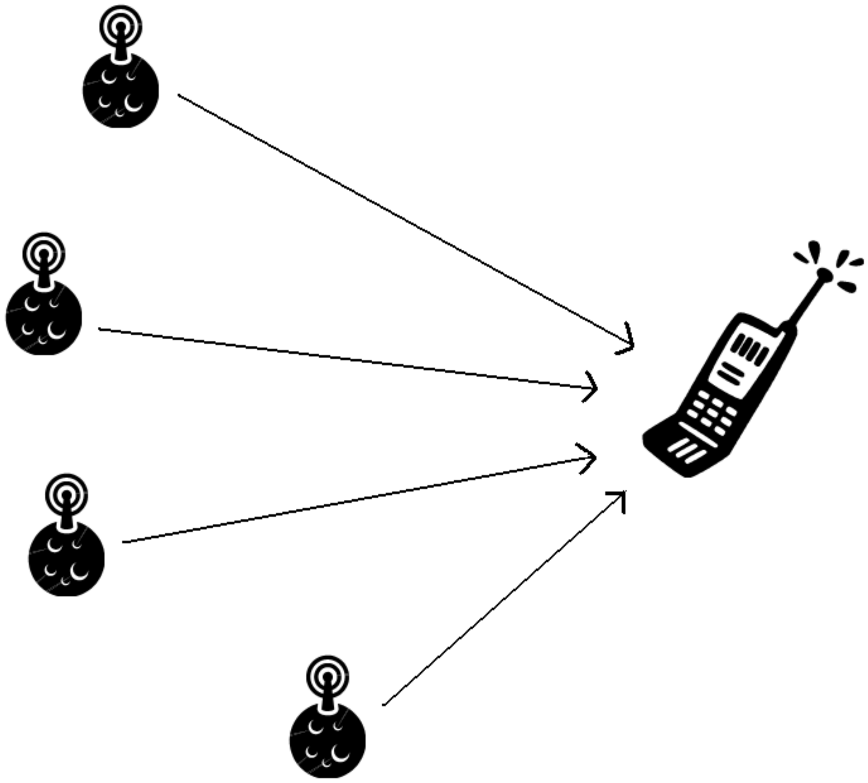

Fig. 20 Four antennas transmitting to a UT

The higher dimensional full rate schemes that obtain full transmit diversity exist only for binary modulation. Once complex-valued modulation is applied, OSTBCs (Orthogonal Space Time Block Codes) with full rate and full diversity do not exist. An alternative is to reduce the data rate. It is this data rate limitation that started the interest in QSTBCs to gain increased data rate while preserving most of the diversity advantage of OSTBCs.

Surprisingly, no researcher ever dared to define exactly what a quasi orthogonal code exactly is. The word quasi is not well defined in such context. A QSTBC of dimension $N \times N$ has an Exported Variable Complexity Metric (EVCM) that satisfies $H_v H_v^H = \sum_1^N |h_i|^2$ J with $J$ being a sparse matrix with ones on its main diagonal, having at least $N^2/2$ zero entries at off-diagonal positions and its remaining entries being bounded in magnitude by +1.

We will study the basis of four transmit antennas and one receive antenna although we like to point out that the statements we give are equivalently true for more antennas. However, explicit terms are often not as comprehensible as for the four antenna case and four antennas are more likely to be used in the near future than for example 8 or 16 transmit antennas.

So let's give an overview over the tree most well known, QSTBCs, the ABBA, the Jafarkhani and finally the code proposed by Papadias and Foschini.

**Jafarkhani:**

This code is also called Extended Alamouti code since it straightforwardly extends the concept of Alamouti:

$$\boldsymbol{S}_{EA} = \begin{bmatrix} \boldsymbol{S}_{12} & \boldsymbol{S}_{34} \\ \boldsymbol{S}_{34}^* & -\boldsymbol{S}_{12}^* \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ s_2^* & -s_1^* & s_4^* & -s_3^* \\ s_3^* & s_4^* & -s_1^* & -s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \tag{4.17}$$

where $\boldsymbol{S}_{12}, \boldsymbol{S}_{21}, \boldsymbol{S}_{34}$ and $\boldsymbol{S}_{43}$ correspond to (2x2) code blocks defined in alamouty 2 antennas codification.

The equivalent virtual channel matrix $\boldsymbol{H}_{EA}$ results in:

$$\boldsymbol{H}_{EA} = \begin{bmatrix} \boldsymbol{H}_{21} & \boldsymbol{H}_{34} \\ -\boldsymbol{H}_{34}^{*} & \boldsymbol{H}_{12}^{*} \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ -h_2^{*} & h_1^{*} & -h_4^{*} & h_3^{*} \\ -h_3^{*} & -h_4^{*} & h_1^{*} & h_2^{*} \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix}$$

(4.18)

This is virtual equivalent (4 x 4) channel matrix. In this way, a virtual, structured (4 x 4) MIMO channel corresponding to four transmit and four virtual receive antennas is obtained.

**The ABBA code proposed by Tirkkonen:**

$$\boldsymbol{S}_{ABBA} = \begin{bmatrix} \boldsymbol{S}_{12} & \boldsymbol{S}_{34} \\ \boldsymbol{S}_{34}^{*} & -\boldsymbol{S}_{12}^{*} \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ s_2^{*} & -s_1^{*} & s_4^{*} & -s_3^{*} \\ s_3 & s_4 & s_1 & s_2 \\ s_4^{*} & -s_3^{*} & s_2^{*} & -s_1^{*} \end{bmatrix}$$

(4.19)

The equivalent virtual channel matrix $\boldsymbol{H}_{ABBA}$ results in:

$$\boldsymbol{H}_{ABBA} = \begin{bmatrix} \boldsymbol{H}_{21} & \boldsymbol{H}_{34} \\ -\boldsymbol{H}_{34}^{*} & \boldsymbol{H}_{12}^{*} \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^{*} & -h_1^{*} & h_4^{*} & -h_3^{*} \\ h_3 & h_4 & h_1 & h_2 \\ h_4^{*} & -h_3^{*} & h_2^{*} & -h_1^{*} \end{bmatrix}$$

(4.20)

**Papadias and Foschini:**

This code that cannot be described by sub-blocks like the first two examples:

$$\boldsymbol{S}_{PF} = \begin{bmatrix} \boldsymbol{S}_{12} & \boldsymbol{S}_{34} \\ \boldsymbol{S}_{34}^{*} & -\boldsymbol{S}_{12}^{*} \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ s_2^{*} & -s_1^{*} & s_4^{*} & -s_3^{*} \\ s_3 & -s_4 & -s_1 & s_2 \\ s_4^{*} & s_3^{*} & -s_2^{*} & -s_1^{*} \end{bmatrix} \tag{4.21}$$

The equivalent virtual channel matrix $\boldsymbol{H}_{PF}$ results in:

$$\boldsymbol{H}_{PF} = \begin{bmatrix} \boldsymbol{H}_{21} & \boldsymbol{H}_{34} \\ -\boldsymbol{H}_{34}^{*} & \boldsymbol{H}_{12}^{*} \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ -h_2^{*} & h_1^{*} & -h_4^{*} & h_3^{*} \\ -h_3 & h_4 & h_1 & -h_2 \\ -h_4^{*} & -h_3^{*} & h_2^{*} & h_1^{*} \end{bmatrix} \tag{4.22}$$

An important characteristic of QSTBCs is their distinct equivalent, highly structured, virtual MIMO channel matrix $H_v$ . Let us consider an STBC defined in 4.23 by $S$, e.g SEA, and an 4×1 frequency flat MISO system. Then we obtain $\boldsymbol{r} = Sh + n$, where r denotes the vector of four temporally successive receive samples. The channel coefficients are denoted by $\boldsymbol{h} = [h_1, h_2, h_3, h_4]^T$ and $n$ is the modified noise vector similar to the Alamouti 2x1. Complex conjugating some rows of $S$ leads to the equivalent, highly structured, virtual MIMO channel matrix (EVCM). For example, changing the second and fourth element of the code ABBA or PF leads to a modified received signal vector y that can be written in the equivalent form

$$\boldsymbol{y} = \begin{bmatrix} r_1 \\ r_2^{*} \\ r_3 \\ r_4^{*} \end{bmatrix} = \boldsymbol{H}_v \boldsymbol{s} + \boldsymbol{n} = \boldsymbol{H}_v \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \boldsymbol{n}, \tag{4.23}$$

introducing $\boldsymbol{H}_v$ as the 4×1 highly structured EVCM. Defining h now as $\boldsymbol{h} = [h_1 h_2 h_3 h_4]^T$ we obtain the four received signal values:

$$r_1 = s_1 h_1 + s_2 h_2 + s_3 h_3 + s_4 h_4 + v_1$$

(4.24)

$$r_2 = s_2^* h_1 - s_1^* h_2 + s_4^* h_3 - s_3^* h_4 + v_2$$

(4.25)

$$r_3 = s_3^* h_1 + s_4^* h_2 - s_1^* h_3 - s_2^* h_4 + v_3$$

(4.26)

$$r_4 = s_4 h_1 - s_3 h_2 - s_2 h_3 + s_1 h_4 + v_4$$

(4.27)

Let's now continue concerned in model code used in this study, Extended Alamouti,

Applying a matched filter $\boldsymbol{H}^H$ at the receiver, the non-orthogonality of the EA code shows up in the Grammian matrix $\boldsymbol{G}_{EA}$:

$$\boldsymbol{G}_{EA} = \boldsymbol{H}_{EA}^H \boldsymbol{H}_{EA} = h^2 \begin{bmatrix} 1 & 0 & 0 & X_{EA} \\ 0 & 1 & -X_{EA} & 0 \\ 0 & -X_{EA} & 1 & 0 \\ X_{EA} & 0 & 0 & 1 \end{bmatrix} \tag{4.28}$$

Where $h^2 = h_1^2 + h_2^2 + h_3^2 + h_4^2$ is the channel gain and X is a channel dependent interference parameter:

$$X_{EA} = \frac{2Re \ (h_1 h_4^* + h_2 h_3^*)}{h^2} \tag{4.29}$$

It is well known that $\boldsymbol{G}_{EA}$ should approximate a scaled identity-matrix as far as possible to get full diversity and optimum BER performance. This means, $|X|$ should be as small as possible. In the case of $X = 0$, full orthogonality is obtained.

To receive and decode original data we will use two different methods, a Zero-Forcing (ZF) receiver and maximum-likelihood (ML) receiver. To simplify the transmission on the receiver side we apply a low-complexity ZF receiver. Due to its quasi orthogonality the results for this transmission system using a linear receiver differ from the optimum ML receiver performance. Nevertheless, the simple ZF detection approach exhibits very low computational complexity and also benefits from the fact that complex matrix inversion is not necessary in the case of QSTBCs. The ZF receiver algorithm leads to:

$$\boldsymbol{Y} = (\boldsymbol{H}_v^H \boldsymbol{H}_v)^{-1} r \ = \ \boldsymbol{s} \ + \ (\boldsymbol{H}_v^H \boldsymbol{H}_v)^{-1} \boldsymbol{H}_v^H \bar{v} \tag{4.30}$$

We also can see from (4.30) that the ML decision metric of this code can be written as the sum of two terms $f_1(s_1,s_4)+ f_2(s_2,s_3)$, where $f_1$ depends only on $s_1,s_4,$ and $f_2$ depends only on $s_2,s_3$. Thus, the minimization can be done separately on these two terms, i.e., symbol pairs $(s_1,s_4)$ and $(s_2,s_3)$ can be decoded separately, which leads to a fast ML decoding with the complexity close to the one required by the simple Alamouti scheme.

# Chapter 5

## 5 Cooperative Systems

After the idea of a cooperative system structure [24] [25] has been proposed for wireless communications, it has suffer a incredible development. The in-cell mobile units share the use of their antennas to assist each other with data transmission, whereby they form a virtual transmitting or receiving array system. Diversity gain, which is usually referred as cooperative diversity gain, can be achieved. There has been much inspired research activity on the cooperative communication systems [26] [27].

While multiple-antenna (MIMO) systems have been widely acknowledged, to the extent that certain transmit diversity methods (i.e., Alamouti signaling) and have been incorporated into wireless standards transmit diversity, it may not be practical for other scenarios. Specifically, due to size, cost, or hardware limitations, a wireless agent may not be able to support multiple transmit antennas. Examples include most handsets (size), or the nodes in a wireless sensor network (size, power). Is here that appears a new class of techniques known as cooperative communication. Cooperative diversity is a promising solution for wireless systems to overcome the above limitations, improving system fairness, extending coverage and increasing capacity. The basic idea is that either single-antenna mobiles in a multi-user scenario or dedicated fixed relays can share their antennas in a manner that creates a virtual system cooperate (fig.21).

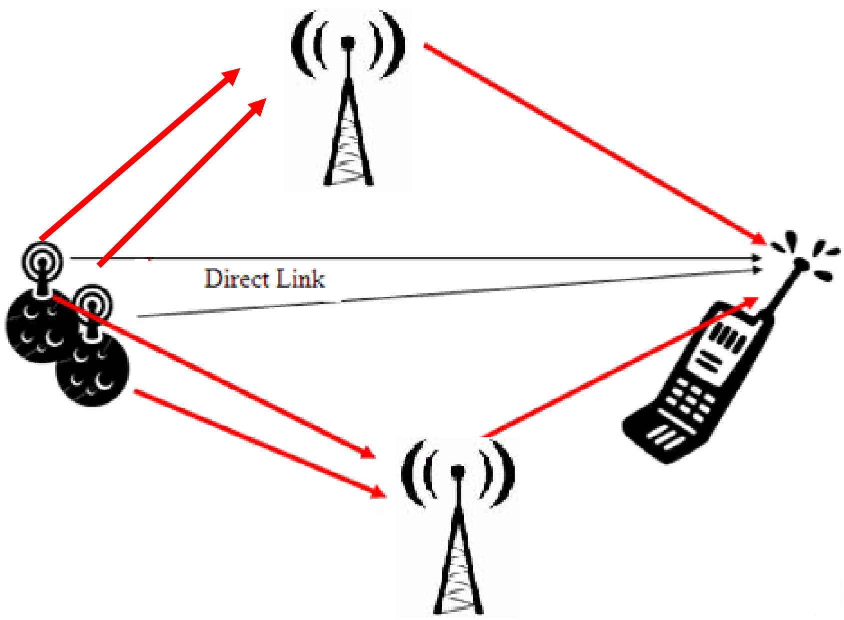


Fig. 21 Cooperative scheme

## 5.1 Cooperation strategies

Cooperative communications protocols can be generally categorized into fixed relaying schemes and adaptive relaying schemes. In fixed relaying, the channel resources are divided between the source and the relay in a fixed (deterministic) manner. The processing at the relay differs according to the employed protocol. In a fixed amplify-and-forward (AF) relaying protocol, the relay simply scales the received version and transmits an amplified version of it to the destination. Another possibility of processing at the relay node is for the relay to decode the received signal, re-encode it and then retransmit it to the receiver. This kind of relaying is termed a fixed decode-and-forward (DF) relaying protocol.

Another important aspect is the communication modes of operation. The communication can be done by simple operation, half-duplex operation or full-duplex operation, and there are a several differences between them. Duplex simply means you are able to send and receive data from the same device. So in half-duplex we can send and receive, but only one-way at a time, like when we use a walkie-talkie. For it way, full-duplex is able send and receive information at the same time.

### 5.1.1 Fixed amplify and forward relaying

Amplify-and-forward, first proposed by Laneman in [28] [29] is perhaps the most conceptually simple of the cooperative signaling methods. Each user in this method receives a noisy version of the signal transmitted by its partner. As the name implies, the user then amplifies and retransmits this noisy signal. The base station will combine the information sent by the user and partner and will make a final decision on the transmitted bit. Although the noise of the partner is amplified in this scheme, the base station still receives two independently faded versions of the signal and is thus able to make better decisions for the transmitted bits. A potential challenge in this scheme is that sampling, amplifying, and retransmitting analog values may be technologically non-trivial. Nevertheless, amplify-and-forward is a simple method that lends itself to analysis, and therefore has been very useful in furthering the understanding of cooperative communication systems.

### 5.1.2 Fixed decode and forward relaying

This method is perhaps closest to the idea of a traditional relay. In this scheme, a user attempts to detect the partner's bits, and then retransmits an estimate of the detected bits. An example of decode-and-forward signaling can be found in the work of Sendonaris in [30] which was actually one of the first works in the area of cooperative communication and has inspired much of the current activity in this area. This work presents a theoretical analysis and a simple CDMA implementation of detect-and-forward cooperative signaling. The following is an example of the proposed CDMA implementation. Each user has its own spreading code, denoted $C_1(t)$ and $C_2\ (t)$. The two user's data bits are denoted $b_i^{(n)}$, where $i\ =\ 1, 2$ are the user indices and $\boldsymbol{n}$ denotes the time index of information bits. The term $\hat{b}_i^{(n)}$ denotes the partner's hard-detected estimate of User it's bit. Factors $a_{i,j}$ denote signal amplitudes, and hence represent power allocation to various parts of the signaling.

Each signaling period consists of three bit intervals. Denoting the signal of User 1 by $X_1(t)$ and the signal of User 2 by $X_2(t)$,

$$X_1\,(t) = a_{11}\, b_1^{(1)} c_1(t)\,, a_{12}\, b_1^{(2)} c_1(t)\,, a_{13}\, b_1^{(2)} c_1(t) + a_{14}\, \hat{b}_2^{(2)} c_2(t) \tag{5.1}$$

$$X_2\,(t) = a_{21}\, b_2^{(1)} c_2(t)\,, a_{22}\, b_2^{(2)} c_2(t)\,, a_{23}\, \hat{b}_1^{(2)} c_1(t) + a_{24}\, b_2^{(2)} c_2(t) \tag{5.2}$$

In other words, in the first and second intervals, each user transmits its own bits. Each user then detects the other user's second bit, and in the third interval, both users transmit a linear combination of their own second bit and their estimate of the partner's second bit, each multiplied by the appropriate spreading code. The transmit powers for the first, second, and third intervals are variable, and by optimizing the relative transmit powers according to the conditions of the uplink channel and the inter-user channel, this method provides adaptability to channel conditions. The powers are allocated through the factors $a_{i,j}$ such that an average power constraint is maintained. Roughly speaking, whenever the inter-user channel is favorable, more power will be allocated to cooperation, whereas whenever the inter-user channel is not favorable, cooperation is reduced. With this signaling scheme, it is possible that the partner forwards an erroneous estimate of the user's bit, in which case cooperation can be detrimental to the eventual detection of the bits at the base station. In addition, the base station needs to know the probability of error for the inter-user channel for optimal (maximum-likelihood) detection. To avoid the problem of error propagation by the partner, Laneman proposed a hybrid detect-and-forward method where, at times when the fading channel has high instantaneous SNR, users detect and forward their partner's data, but when the channel has low SNR, the users revert to a non-cooperative mode. This is a variation of adapting the coefficients $a_{i,j}$ in the method of Sendonaris [31], and it has been shown to perform very well.

### 5.1.3 Other cooperation strategies

The two technique described below are the most common ones for fixed relaying, however there are other techniques, such as compress-and-forward and coded cooperation which are also used and we'll give an overview.

Coded cooperation [32] works by sending different portions of each user's code word via two independent fading paths. The basic idea is that each user tries to transmit incremental redundancy to its partner. Whenever that is not possible, the users automatically revert to a non-cooperative mode. The key to the efficiency of coded cooperation is that all this is managed automatically through code design, with no feedback between the users. The users divide their source data into blocks that are augmented with cyclic redundancy check (CRC) code.1In coded cooperation, each of the users' data is encoded into a codeword that is partitioned into two segments, containing $N_1$ bits and $N_2$ bits, respectively. It is easier to envision the process by a specific example: consider that the original codeword has $N_1$ +$N_2$ bits; puncturing this codeword down to $N_1$ bits, we obtain the first partition, which itself is a valid (weaker) codeword. The remaining $N_2$ bits in this example are the puncture bits. Of course, partitioning is also possible via other means, but this example serves to give an idea of the intuition behind coded cooperation.

Likewise, the data transmission period for each user is divided into two time segments of $N_1$ and $N_2$ bit intervals, respectively. We call these time intervals frames. For the first frame, each user transmits a code word consisting of the $N_1$-bit code partition. Each user also attempts to decode the transmission of its partner. If this attempt is successful (determined by checking the CRC code), in the second frame the user calculates and transmits the second code partition of its partner, containing $N_2$ code bits. Otherwise, the user transmits its own second partition, again containing $N_2$ bits. Thus, each user always transmits a total of $N = N_1 + N_2$ bits per source block over the two frames. We define the level of cooperation as $\frac{N_2}{N}$, the percentage of the total bits for each source block the user transmits for its partner.

Compress-and-forward differs from decode/amplify-and-forward because while in the later the relay transmits a copy of the received message, in compress-and-forward the relay transmits a quantized and compressed version of the received message. Therefore, the destination node will perform the reception functions by combining the received message from the source node and its quantized and compressed version from the relay node.

The quantization and compression process at the relay node is a process of source encoding, i.e., the representation of each possible received message as a sequence of symbols. For clarity and simplicity, let us assume that these symbols are binary digits (bits). At the destination node, an estimate of the quantized and compressed message is obtained by decoding the received sequence of bits. This decoding operation simply involves the mapping of the received bits into a set of values that estimate the transmitted message. This mapping process normally involves the introduction of distortion (associated to the quantization and compression process), which can be considered as a form of noise.

## 5.2 Cooperative communications with single relay

In traditional communication networks data transmission directly occurs between the transmitter and the receiver. No user solicits the assistance of another one. However, it has becomes to be insufficient to all quality and capacity of service required. So, in a general communication network, there are many intermediate nodes that are available to help. For example in wireless networks, when one node broadcasts its messages, all nearby nodes overhear this transmission. Processing and forwarding these messages to the intended destination, system performance, whether it be throughput, lifetime, or coverage area, can be improved. And from here was born the idea of cooperative communication. In information theory the idea of cooperation was first presented by van der Meulen (1971), which established the foundations of the relay channel (single relay)l. The relay channel is a three-terminal network, in which Terminal 1 (source) aims to transmit to Terminal 3 (destination) with the help of Terminal 2 (relay) (as show in fig.22). The aim is to attain the highest communication rate from Terminal 1 to Terminal 3. In general the intermediate relay enhances communication rates of the direct link from Terminal 1 to 3.

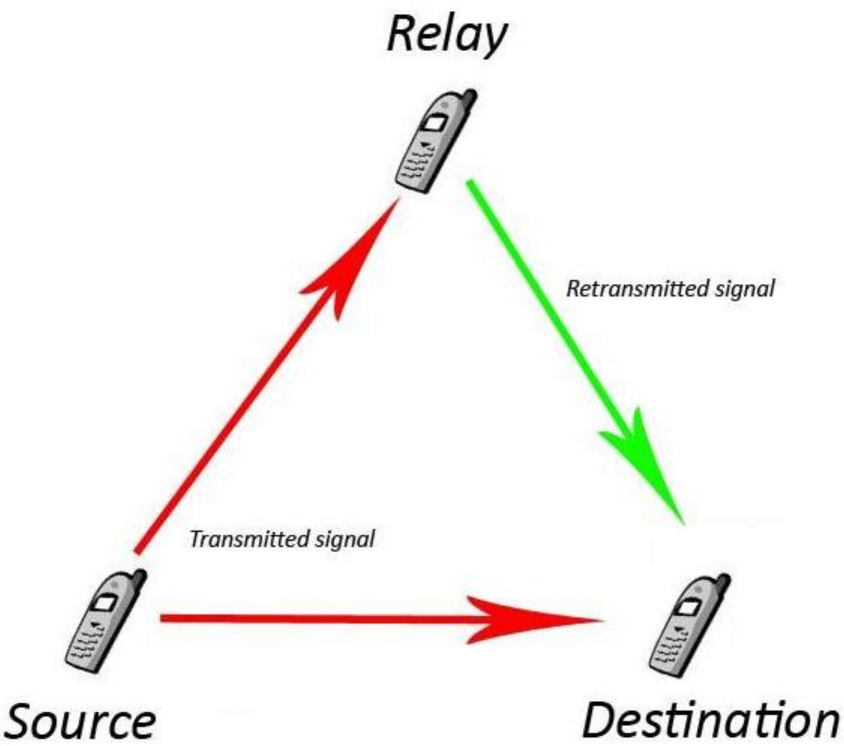


Fig. 22 One relay channel

With the introduction of the relay channel, independent paths between the user and the base station are generated. The relay channel can be thought of as an auxiliary channel to the direct channel between the source and destination.

In Phase 1, the source broadcasts its information to both the destination and the relay. The received signals $y_{s,d}$ and $y_{s,r}$ at the destination and the relay, respectively, can be written as

$$y_{s,d} = \sqrt{P_1} h_{s,d} x + \eta_{s,d}, \tag{5.3}$$

$$y_{s,t} = \sqrt{P_1} h_{s,t} x + \eta_{s,r}, \tag{5.4}$$

in which $P_1$ is the transmitted power at the source, $x$ is the transmitted information symbol, $\eta_{s,d}$ and $\eta_{s,r}$ are additive noise.

A key aspect of the cooperative communication process is the processing of the signal received from the source node done by the relay. These different processing schemes result in different cooperative communications protocol.

Fixed relaying has the advantage of easy implementation, but the disadvantage of low bandwidth efficiency. This is because half of the channel resources are allocated to the relay for transmission, which reduces the overall rate. This is true especially when the source–destination channel is not very bad, because in such a scenario a high percentage of the packets transmitted by the source to the destination could be received correctly by the destination and the relay's transmissions would be wasted. Adaptive relaying techniques, comprising selective and incremental relaying, try to overcome this problem.

In selective relaying, if the signal-to-noise ratio of the signal received at the relay exceeds a certain threshold, the relay performs decode-and-forward operation on the message. On the other hand, if the channel between the source and the relay has severe fading such that the signal-to-noise ratio is below the threshold, the relay idles. Moreover, if the source knows that the destination does not decode correctly, then the source may repeat to transmit the information to the destination or the relay may help forward information, which is termed as incremental relaying. In this case, a feedback channel from the destination to the source and the relay is necessary.

### 5.2.1 Decode and Forward protocols

In Phase 2 (according with the process described before), for a DF cooperation protocol, if the relay is able to decode the transmitted symbol correctly, then the relay forwards the decoded symbol with power $P_2$ to the destination. The received signal at the destination in Phase 2 in this case can be modeled as

$$y_{r,d} = \sqrt{P_2} h_{r,d} \bar{x} + \eta_{r,d}, \tag{5.5}$$

$h_{r,d}$ is the channel coefficient from the relay to the destination, and it is modeled as a zero-mean, complex Gaussian random variable with variance $\delta_{r,d.}$ $\bar{x}$ is the data decoded at the relay. No that if this data is correctly decoded ($\bar{x} = x$) this scheme can achieve a diversity order of 2. However, in the outage case when the relay fails to decode the data correctly, it cannot help the UT for the current cooperation round. In such case, the BS will get interference from the cooperative path and cooperation will not be beneficial. The noise term $\eta_{r,d}$ is also modeled as a zero-mean complex Gaussian random variable with variance $N_0$.

With knowledge of the channel coefficients $h_{s,d}$ and $h_{r,d}$, the destination detects the transmitted symbols by jointly combining the received signal $y_{s,d}$ from the source and $y_{r,d}$ from the relay. The combined signal at the MRC detector can be written as [33]

$$y = a_1 y_{s,d} + a_2 y_{r,d}, \tag{5.6}$$

in which the factors $a_1$ and $a_2$ are determined such that the SNR of the MRC output is maximized, and they can be specified as $a_1 = \sqrt{P_1} h_{s,d}^* / N_0$ and $a_2 = \sqrt{\tilde{P}_2} h_{r,d}^* / N_0$. Assume that the transmitted symbol $x$ has average energy 1, then the SNR of the MRC output is

$$\Upsilon = \frac{P_1 |h_{s,d}|^2 + P_2 |h_{r,d}|^2}{N_0} \tag{5.7}$$

### 5.2.2 Amplify and Forward protocols

For an AF cooperation protocol, in phase 2 the relay amplifies the received signal and forwards it to the destination with transmitted power $P_2$. The received signal at the destination in phase 2 is specified as

$$y_{r,d} = \frac{\sqrt{P_2}}{\sqrt{P_1 |h_{s,r}|^2 + N_0}} h_{r,d} y_{s,r} + \eta_{r,d} \tag{5.8}$$

where $h_{r,d}$ is the channel coefficient from the relay to the destination and $\eta_{r,d}$ is an additive noise. The received signal can also be written as

$$y_{r,d} = \frac{\sqrt{P_1 P_2}}{\sqrt{P_1 |h_{s,r}|^2 + N_0}} h_{r,d} h_{s,r} x + \eta'_{r,d}$$

(5.9)

Where

$$\eta'_{r,d} = \frac{\sqrt{P_2}}{\sqrt{P_1 |h_{s,r}|^2 + N_0}} h_{r,d} \eta_{s,r} + \eta_{r,d}$$

(5.10)

Assume that the noise terms $\eta_{s,r}$ and $\eta_{r,d}$are independent, then the equivalent noise $\eta'_{r,d}$ is a zero-mean, complex Gaussian random variable with variance

$$\left( \frac{P_2 |h_{r,d}|^2}{P_1 |h_{s,r}|^2 + N_0} + 1 \right) N_0$$

(5.11)

In both the DF and AF cooperation protocols, the channel coefficients $h_{s,d}$, $h_{s,r}$, and $h_{r,d}$ are assumed to be known at the receiver, but not at the transmitter. The destination jointly combines the received signal from the source in Phase 1 and that from the relay in phase 2, and detects the transmitted symbols by using MRC. In both protocols, we consider a total transmitted power P such as

$$P_1 + P_2 = P$$

(5.12)

## 5.3 Multi-node cooperative

The concept of multi-node cooperative communication appears as a development of a single relay cooperative technique, and therefore also has the objective of combat the signal fading due to multipath propagation. The cooperating nodes act as the relay channels for the source node. Cooperative diversity techniques, or equivalently virtual MIMO systems, constitute a new communication paradigm where numerous questions need to be answered. In [34] and [35], Laneman et al. proposed different cooperative diversity protocols and analyzed their performance in terms of outage behavior. The terms decode-and-forward and amplify and- forward have been introduced in these two works, as we have seen before when analyzing single relay performance. In DF, each relay receives and decodes the signal transmitted by the source, and then it forwards the decoded signal to the destination which combines all of these copies in a proper way. AF is a simpler technique, in which the relay amplifies the received signal and then forwards it to the destination. Although the noise is amplified along with the signal in this technique, we still gain spatial diversity by transmitting the signal over two spatially independent channels. Terminologies other than cooperative diversity are also used in the research community to refer to the same concept of achieving spatial diversity via forming virtual antennas. User cooperation diversity was introduced by Sendonaris et al. in [36]. In this two-part series of papers, the authors implemented a two-user CDMA cooperative system, where both users are active and use orthogonal codes to avoid multiple access interference. Another technique to achieve diversity that incorporates error-control-coding into cooperation is coded cooperation introduced by Hunter et al. in [37].

## 5.4 World Wile Initiatives

There are at the moment projects and groups whose purpose is the study of cooperative based systems.

This section has the purpose of introduce two word wide initiatives that are taking place in order to turn cooperative diversity results into practical and commercial standards; This two initiatives are IEEE 802.16j and CODIV (the work of this thesis was done in the scope of this European project).

The main differences between this two initiatives lies in the fact that CODIV assumes that relays can be mobile, whereas 802.16j assumes they are fixed, as show fig.21.

### 5.4.1 IEEE 802.16j Standard Main Characteristics

Based on the IEEE 802.16e standard, WiMAX has proposed a relay-based approach, i.e., IEEE 802.16j, to extend the service area of BSs and to improve the received signal strength (RSS) quality. IEEE 802.16j has the capability to achieve two significant advantages: low-cost of building IEEE 802.16 WiMAX networks and compatible with existing WiMAX standards.

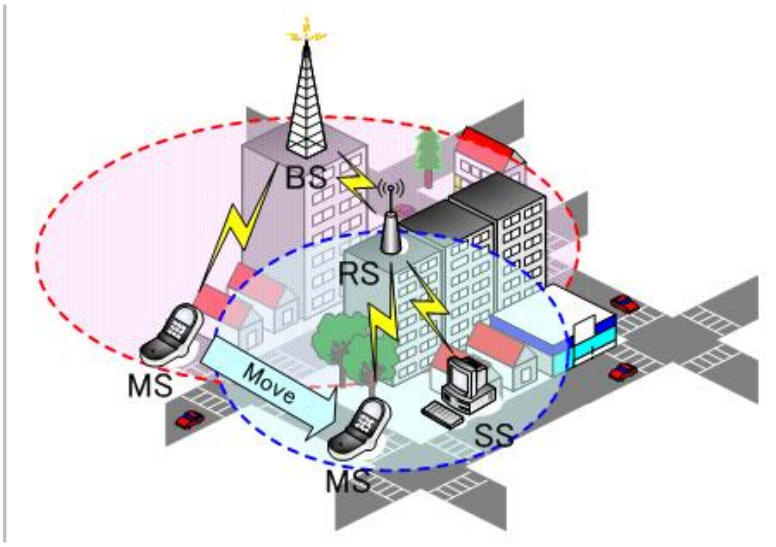


**Fig. 23 Main 802.16j proposed scenario**

The IEEE 802.16 working group has created on March 30, 2006 a \Relay Task Group" to develop a draft for the Mobile Multi-hop Relay (MMR) system (802.16j). As listed in the documents issued within this task group [38], the main motivations for multihop relaying were the following:

- Coverage/Range Extension
- Improved throughput
- In-building penetration
- Infrastructure offloading
- Improved frequency reuse

The proposed scenario for 802.16j is depicted in fig.23, from we can see the use of special diversity.

In sense of the standard to be developed has to provide backwards compatibility with 802.16.e, the architecture is not a mesh network but rather a tree. The relay stations are not allowed to forward the traffic between two UT nodes and all the traffic must go through the MMR-BS as in a conventional cellular architecture. An ad-hoc capable UT can be a source, destination, or a relay, although dedicated relay stations are considered in most of the documents issued up to now.

### 5.4.2 CODIV cooperation scenarios

The main goal of the CODIV is to investigate and develop technologies that would enhance the performance of wireless communication systems, employing the inherent diversity of radio channels (channel diversity), as well as the cooperation between users (multiuser diversity). In fact, one of the key points of the project is to exploit the cooperative reserves and capabilities of the user-equipments, although incentive policies might be needed to stimulate this cooperation. The expected improvement in the performance of radio systems pursued by CODIV technologies will come in terms of higher bit rates and spectrum efficiency, and decreased (enhanced) power consumption, as well as coverage extension and fairness. Therefore, a first step in the project should be to identify and define the scenarios where these technologies can be validated and evaluated, allowing the estimation of the benefits for the future wireless communications, on one hand, and enabling comparisons with other proposed solutions, on the other hand.

Also it is important to emphasize that the idea of CODIV project is not to develop a new standard rather to improve the performance of the existing wireless systems by means of the inclusion of particular algorithms exploiting cooperative diversity techniques. It is concerned in to avoid substantial modifications on those systems and keeping their basic characteristics unchanged. In this sense, the target radio systems addressed by CODIV from the start of the project have been WiMax (based on the IEEE 802.16 standard), and Long Term Evolution (LTE) promoted by the 3rd Generation Partnership Project (3GPP).

Two cooperative diversity strategies relevant for this project's purposes were identified: techniques based on the use of relays (dedicated or terminal-embedded) and multi-antennas techniques. Although both options could be implemented on any kind of scenario (e.g. indoor, outdoor, rural, urban, hot-spot, etc), a detailed description of the final evaluation scenario was not carried out.

The most important scenarios for CODIV purposes defined at single-cell level so far are the following:

- BS with at least two transceivers and simple (only one antenna) or complex UTs with the possibility to act as relaying-able terminal in idle periods in order to validate and assess the CODIV relaying techniques.

- BS with at least two transceivers, several dedicated relay nodes (in principle fixed RNs) with semi-duplex or full-duplex operation, and simple or complex UTs in order to validate and assess also the CODIV relaying techniques.

- BS with at least two transceivers and simple UTs without the possibility to act as relaying-able terminals in order to validate and assess the CODIV multi-user MIMO techniques.

However, any combination of these tree basic scenarios could be included in CODIV project.

# Chapter 6

# 6 Relay-Assisted Schemes Implemented

In this thesis was implemented and evaluated cooperative schemes designed for downlink OFDM based system as shown in fig. 24 and fig. 25, using distributed space-frequency block coding protocols. This cooperative scheme it is an extension of the one proposed under the European CODIV project [39], where all terminals were equipped with single antenna. The performance of these cooperative schemes is evaluated under realistic scenarios, considering typical pedestrian scenarios based on LTE specifications. The proposed schemes are also compared against the non-cooperative OFDM based systems.

## 6.1 System Model

In fig 24 scenario, is assumed the use of an antenna array of $M$ at the BS and a single antenna at the UT. At the RN we considered single antenna. In fig. 25 scenario, is also assumed $M$ antennas at the BS and single antenna at the UT, but $L$ antennas in each relay. For both cases is assumed a half-duplex decode-and-forward relay communication protocol.

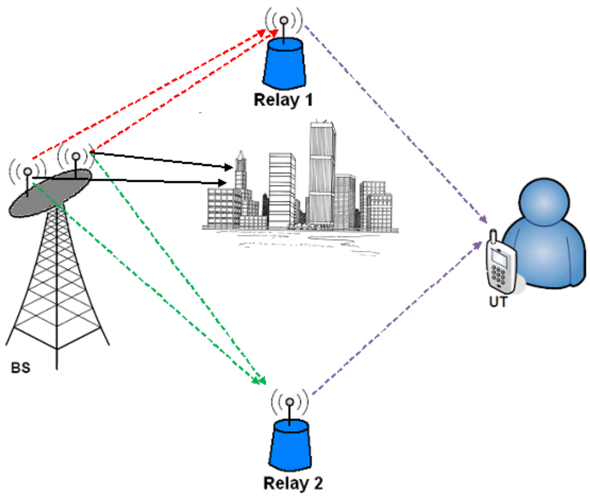


Fig. 24 Scheme 1 - Two relays with single antenna

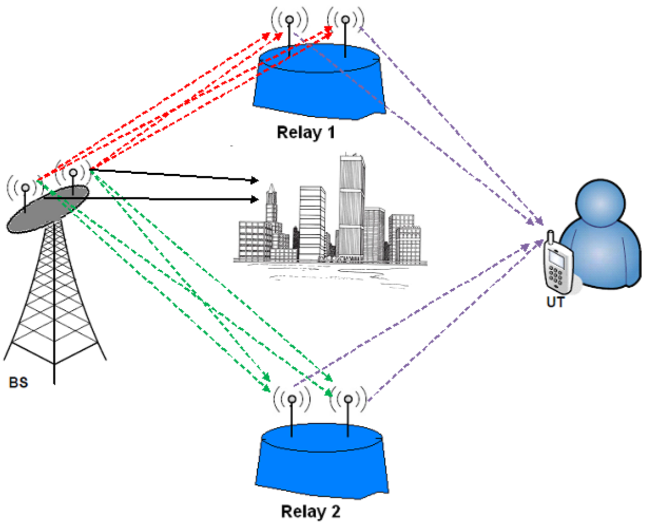


Fig. 25 Scheme 2 - Two relay with two antennas

Link Between BS and R1 (BS>R1)

Link Between BS and R2 (BS>R2)

Link Between Relays and UT (R>UT)

Direct link

From the above figures, four main links can be identified: BS>$R_1$, represented by the channels $h_{bs>r1}^{m,l,f}, m = 1{,}2\,, l = 1{,}2\,, f = 1, .. N_c.$ , i.e., the channel between the $m$ BS antenna, $l$ relay antennas on subcarrier $f$; BS>$R_2$ represented by the channels $h_{bs>r2}^{m,l,f}, m =$

$1,2\,, l = 1,2\,, f = 1,..N_c$.; $R_1$>UT represented by the channels $h^{l,f}_{r1>ut}, l = 1,2\,, f = 1,..N_c$ and $R_2$>UT represented by the channels $h^{l,f}_{r2>ut}, l = 1,2\,, f = 1,..N_c$. In this cooperative system the direct link it is not considered.

Since we consider half-duplex decode-and-forward relay protocol , the communication cycle for the studied cooperative scheme requires two phases:

- In the first one, the source, BS broadcasts its own data at full power to the relays, which do not transmit data during this stage. The data are space-frequency encoded before transmission by the BS.
- During the second phase, the relay node can help the source by forwarding the information, also at full power, to the destination node (UT), whereas the source is idle. If the relays are equipped with single antenna, the data to be transmitted to the UT are space-frequency encoded by using the Alamouti scheme implemented in a distributed manner. If the relays are equipped with 2-antenna array, the data to be transmitted to the UT are also space-frequency encoded but now by using the Quasi-Orthogonal codes discussed in chapter 4 implemented in a distributed fashion.

Let us define the matrix $\boldsymbol{H}^f_{bs>rn}, n = 1,2$ of size $MxL$ that represents the overall channel between the *BS* and each relay on subcarrier *f*. The channel $\boldsymbol{H}^f_{bs>rn}$ can be decomposed as the product of the fast fading $\overline{\boldsymbol{H}}^f_{bs>rn}$ and slow fading $\sqrt{\rho_{bs>rn}}$ components, i.e., $\boldsymbol{H}^f_{bs>rn} = \overline{\boldsymbol{H}}^f_{bs>rn}\sqrt{\rho_{bs>rn}}$ , where $\rho_{bs>rn}$ represents the long-term power gain between BS and relay $n$. The fast fading components of $\overline{\boldsymbol{H}}^f_{bs>rn}$ may exhibit correlation and we further decompose it as $\overline{\boldsymbol{H}}^f_{bs>rn} = \boldsymbol{R}^{1/2}_{Tx}\left(\overline{\boldsymbol{H}}^f_{bs>rn}\right)_{iid}\boldsymbol{R}^{1/2}_{Rx}$, where $\left(\overline{\boldsymbol{H}}^f_{bs>rn}\right)_{iid}$ contains the fast fading coefficients with i.i.d. $CN$ (0,1) entries and $\boldsymbol{R}_{Tx}$ and $\boldsymbol{R}_{Rx}$ are the normalized transmitter and receiver correlation matrixes defined in [40]. The antenna channels from the BS to relay $n$, may be correlated but the links seen from different BSs to a given relay are assumed to be uncorrelated as the relays are geographically separated. Note that for the case where the relays are equipped with single antenna $L$=1 and $\boldsymbol{R}_{Rx} = 1$. We also can define the overall channel between each relay and the UT as a vector

$h^f_{rn>ut}, n = 1{,}2$ of size $Lx1$ on subcarrier $f$. The channel $h^f_{rn>ut}$ can be decomposed as the product of the fast fading $\bar{h}^f_{rn>ut}$ and slow fading $\sqrt{\rho_{rn>ut}}$ components, i.e., $h^f_{rn>ut} = h^f_{rn>ut}\sqrt{\rho_{rn>ut}}$ , where $\rho_{rn>ut}$ represents the long-term power gain between relay $n$ and UT. The channel . $\bar{h}^f_{rn>ut}$ can be further decomposed by $\bar{h}^f_{rn>ut} = \boldsymbol{R}^{1/2}_{Tx}\left(\bar{h}^f_{rn>ut}\right)_{iid}$. The antenna channels from the relay $n$ to the UT, may be correlated but the links seen from different relays to the UT are assumed to be uncorrelated as said above the relays are geographically separated.

## 6.2 Relay-Assisted schemes

Throughout this work, we shall refer to these schemes as M×2L×1 coop., depending on the number of antennas at the BS (*M*) and relays (*L*).

### 6.2.1 Single Antenna Relay

Considering a single antenna relay (*L*=1), in the first phase of the communication cycle we consider a 2x1 Alamouti transmition between the BS and each relay.

The mapping scheme with two transmit antennas used in this work is shown in table I [43] , where $s_f$ is the data transmitted by the BS on subcarrier *f*. (.)* denotes complex conjugation.

| | **Antenna 1** | **Antenna 2** |
|---|---|---|
| **Subcarrier *f*** | $s_f$ | $-s^*_{f+1}$ |
| **Subcarrier *f+1*** | $s_f + 1$ | $s^*_f$ |

**Table I : SFBC Mapping Scheme.**

According to the communication cycle adopted in this work, the received signal during the first communication phase at the relay *n* on subcarrier *f* and *f*+1, is found to be,

$$y_n(f) = s_f h_{bs>rn}^{1,f} - s_{f+1}^* h_{bs>rn}^{2,f} + n_n^f$$

(6.1)

$$y_n(f+1) = s_{f+1} h_{bs>rn}^{1,f+1} - s_f^* h_{bs>rn}^{2,f+1} + n_n^{f+1}$$

(6.2)

Where $n_n^f$ is the Gaussian noise added in relay *n* on subcarrier *f*. The OFDM systems are usually designed so that the subcarrier separation is significantly lower than the coherence bandwidth of the channel. Therefore, the fading in two adjacent subcarriers can be considered flat, i.e., we can consider $h_{bs>rn}^{1,f}$ equal to $h_{bs>rn}^{1,f+1}$.

At the relays the signals are space-frequency decoded as,

$$\hat{s}_{n,f} = y_n(f)\left(h_{bs>rn}^{1,f}\right)^* + y_n(f+1)^* h_{bs>rn}^{2,f}$$

(6.3)

$$\hat{s}_{n,f+1} = -y_n(f)^* h_{bs>rn}^{2,f} + y_n(f+1)\left(h_{bs>rn}^{1,f}\right)^*$$

(6.4)

Where $\hat{s}_{n,f}$ is the soft data estimated at relay *n* on subcarrier *f*. Then, these estimates are hard decoded, which are represented now by $\bar{s}_{n,f}$. Note that if the data are successfully detected at the relays $\bar{s}_{n,f} = s_{n,f}, n = 1,2$. In the second phase of the cooperative protocol, the signals are spaced-frequency encoded in a distributed manner, i.e., now the antenna 1 is the relay 1 and the antenna 2 is the relay 2, as shown in table II.

| | **Relay 1** | **Relay 2** |
|---|---|---|
| **Subcarrier *f*** | $\bar{s}_{1,f}$ | $-\bar{s}_{2,f+1}^*$ |
| **Subcarrier *f*+1** | $\bar{s}_{1,f+1}$ | $\bar{s}_{2,f}^*$ |

**Table II : Distributed SFBC Mapping Scheme.**

Considering the above table, the received signals at the UT on subcarriers $f$ and $f+1$ are given by,

$$y_{ut}(f) = \bar{s}_{1,f} h_{r1>ut}^{f} - \bar{s}_{2,f+1}^{*} h_{r2>ut}^{f} + n_{ut}^{f} \tag{6.5}$$

$$y_{ut}(f+1) = \bar{s}_{1,f+1} h_{r1>ut}^{f} + \bar{s}_{2,f}^{*} h_{r2>ut}^{f} + n_{ut}^{f+1} \tag{6.6}$$

Where $n_{UT}^{f}$ is the AWGN added at the UT on subcarrier *f*. As for the first phase, at the UT the signals are space-frequency decoded as,

$$\hat{s}_f = y_{ut}(f)\left(h_{r1>ut}^{f}\right)^{*} + y_{ut}(f+1)^{*} h_{r2>ut}^{f} \tag{6.7}$$

$$\hat{s}_{f+1} = -y_{ut}(f)^{*} h_{r2>ut}^{f} + y_{ut}(f+1)\left(h_{r1>ut}^{f}\right)^{*} \tag{6.8}$$

As it can be seen from (6.5) - (6.8), in the outage case when the relays fail to decode the data correctly, i.e., $\bar{s}_{n,f} \neq s_{n,f}, n = 1,2$, the $\bar{s}_{1,f}$ sent by relay 1 will interfere with the data symbol $\bar{s}_{2,f}$ sent by the relay 2 and in this scenario the relays do not bring any systems improvement. For the particular case where $\bar{s}_{n,f} = s_{n,f}, n = 1,2$, i.e., the data sent by BS is successful decoded at the relays and (6.7) can be re-written as,

$$\hat{s}_f = \left(\left|h_{r1>ut}^{f}\right|^{2} + \left|h_{r2>ut}^{f}\right|^{2}\right) s_f + \left(h_{r1>ut}^{f}\right)^{*} n_{ut}^{f} + \left(h_{r2>ut}^{f}\right)\left(n_{ut}^{f+1}\right)^{*} \tag{6.9}$$

and a diversity order of 2 can be achieved. In practical systems this can be obtained when the link BS-to-RN has high quality.

### 6.2.2 Two Antenna Relay

Considering a two antenna relay ($L$=2), in the first phase of the communication cycle we consider a 2x2 Alamouti transmition between the BS and each relay.
The mapping scheme with two transmit antennas used in this work is shown in [43], where $s_f$ is the data transmitted by the BS on subcarrier $f$. (.)* denotes complex conjugation.(The same like in single antenna relay section).

According to the communication cycle adopted in this work, the received signal during the first communication phase at the relay $n$, antenna $l$ and on subcarriers $f$ and $f$+1, is found to be,

$$y_{n,l}(f) = s_f h_{bs>rn}^{1,l,f} - s_{f+1}^* h_{bs>rn}^{2,l,f} + n_n^f \tag{6.10}$$

$$y_{n,l}(f+1) = s_{f+1} h_{bs>rn}^{1,l,f} - s_f^* h_{bs>rn}^{2,l,f} + n_n^f \tag{6.11}$$

Where $n_n^f$ is the Gaussian noise added in relay $n$ on subcarrier $f$. Such in section 6.2.1 we consider the fading in two adjacent subcarriers can be considered flat, so $h_{bs>rn}^{1,l,f}$ is equal to $h_{bs>rn}^{1,l,f+1}$.

At the relay $n$ and antenna $m$ the signals is space-frequency decoded as,

$$\hat{s}_{n,l,f} = y_{n,l}(f)\left(h_{bs>rn}^{1,l,f}\right)^* + y_{n,l}(f+1)^* h_{bs>rn}^{2,l,f} \tag{6.12}$$

$$\hat{s}_{n,l,f+1} = -y_n(f)^* h_{bs>rn}^{2,l,f} + y_{n,l}(f+1)\left(h_{bs>rn}^{1,l,f}\right)^* \tag{6.13}$$

Where $\hat{s}_{n,m,f}$ is the soft data estimated at relay *n* in antenna *l* on subcarrier *f*. Then, we add the estimates data of two antennas in each relay, and we obtain the full data estimated at relay *n*,

$$\bar{s}_{n,f} = \hat{s}_{n,1,f} + \hat{s}_{n,2,f} \tag{6.14}$$

$$\bar{s}_{n,f+1} = \hat{s}_{n,1,f+1} + \hat{s}_{n,2,f+1} \tag{6.15}$$

These estimates are hard decoded, which are represented now by $\bar{s}_{n,f}$. Again, note that if the data are successfully detected at the relays $\bar{s}_{n,f} = s_{n,f}, n = 1,2$, and a diversity order of 4 can be achieved, because now we how four channels between BS and RN. In the second phase of the cooperative protocol, the signals are spaced-frequency encoded in a distributed manner, but we cannot use an orthogonal code because we have a total of four antennas to transmit. Therefore, we use a Quasi-Orthogonal Space Frequency Block Code (QSFBC). Particularly, we consider a 4x1 Extended Alamouti transmition between the RN and UT. The mapping scheme with four transmit antennas used in this work is shown in table III, where $\bar{s}_{n,f}$ is the data transmitted by the relay *n* on subcarrier *f*.

| | **Relay 1 Antenna 1** | **Relay 1 Antenna 2** | **Relay 2 Antenna 1** | **Relay 2 Antenna 2** |
|---|---|---|---|---|
| **Subcarrier *f*** | $\bar{s}_{1,f}$ | $\bar{s}^*_{1,f+1}$ | $\bar{s}^*_{2,f+2}$ | $\bar{s}_{2,f+3}$ |
| **Subcarrier *f*+1** | $\bar{s}_{1,f+1}$ | $-\bar{s}^*_{1,f}$ | $\bar{s}^*_{2,f+3}$ | $-\bar{s}_{2,f+2}$ |
| **Subcarrier *f*+3** | $\bar{s}_{1,f+2}$ | $\bar{s}^*_{1,f+3}$ | $-\bar{s}^*_{2,f}$ | $-\bar{s}_{2,f+1}$ |
| **Subcarrier *f*+4** | $\bar{s}_{1,f+3}$ | $-\bar{s}^*_{1,f+2}$ | $-\bar{s}^*_{2,f+1}$ | $\bar{s}_{2,f}$ |

**Table III : Distributed SFBC Mapping Scheme.**

Now we apply two different receive algorithms to decode and demodulate data in UT, ZF and ML. Let's analyze first in ZF receiver. We start to consider the following vectors,

$$\boldsymbol{n} = \begin{bmatrix} n_{ut}^{f} & n_{ut}^{f+1} & n_{ut}^{f+2} & n_{ut}^{f+3} \end{bmatrix}^{H} \tag{6.16}$$

$$\boldsymbol{h} = \begin{bmatrix} h_{rn>ut}^{1,1,f} & h_{rn>ut}^{1,2,f} & h_{rn>ut}^{2,1,f} & h_{rn>ut}^{2,2,f} \end{bmatrix}^{H} \tag{6.17}$$

In which $n_{ut}^{f}$ in the Gaussian noise added at the UT in frequency *f*, and $h_{rn>ut}^{n,l,f}$ is the link channel from relay *n* and antenna *l* to UT. For the same reasons like we considerate before, we treat the fading in four adjacent subcarriers like flat, and $h_{rn>ut}^{n,l,f} = h_{rn>ut}^{n,l,f+1} = h_{rn>ut}^{n,l,f+2} = h_{rn>ut}^{n,l,f+3}$.

Now, relating vector $\boldsymbol{n}$, $\boldsymbol{h}$, and matrix $\boldsymbol{S}$ we get,

$$\boldsymbol{y} = \boldsymbol{S} * \boldsymbol{h} + \boldsymbol{n} \tag{6.18}$$

Where $\boldsymbol{S}$ is a matrix with the original data symbols sent by BS as in the table III. $\boldsymbol{y}$ will be manipulated giving place to

$$\ddot{\boldsymbol{y}} = \begin{bmatrix} y_{[1]} & y_{[2]}^{*} & y_{[3]}^{*} & y_{[4]} \end{bmatrix} \tag{6.19}$$

Where $y_{[p]}$is the *p* element of the vector $\boldsymbol{y}$. We also create a special channel matrix **H** like in (4.18), composed by the RN->UT channels,

$$\boldsymbol{H}_{EA} = \begin{bmatrix} h_{1,1,f} & h_{1,2,f} & h_{2,1,f} & h_{2,2,f} \\ -h_{1,2,f}^{*} & h_{1,1,f}^{*} & -h_{2,2,f}^{*} & h_{2,1,f}^{*} \\ -h_{2,1,f}^{*} & -h_{2,2,f}^{*} & h_{1,1,f}^{*} & h_{1,2,f}^{*} \\ h_{2,2,f} & -h_{2,1,f} & -h_{1,2,f} & h_{1,1,f} \end{bmatrix} \tag{6.20}$$

Taking in consideration (6.19) and (6.20), (6.18) can be re-write as,

$$\ddot{\boldsymbol{y}} = \boldsymbol{s} * \boldsymbol{H}_{EA} + \boldsymbol{n} \tag{6.21}$$

Where now is $\boldsymbol{s}$ is a vector with the four data symbol, assuming that $\bar{s}_{1,f} = \bar{s}_{2,f} = \bar{s}_f \,\forall f$. Basically, it is assumed that the data symbols decoded at the relays are the same. Then the received signal is equalized by $\boldsymbol{H_{EA}^H}$ , and the first estimate is given by,

$$\tilde{\boldsymbol{s}}_{ut} = \boldsymbol{H}_{EA}{}^{H} * \ddot{\boldsymbol{y}} = h^2 \begin{bmatrix} 1 & 0 & 0 & X_{EA} \\ 0 & 1 & -X_{EA} & 0 \\ 0 & -X_{EA} & 1 & 0 \\ X_{EA} & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} \bar{s}_f \\ \bar{s}_{f+1} \\ \bar{s}_{f+2} \\ \bar{s}_{+3f} \end{bmatrix} + \boldsymbol{H}_{EA}{}^{H} * \boldsymbol{n} = \begin{bmatrix} \tilde{s}_{1,ut,zf} \\ \tilde{s}_{2,ut,zf} \\ \tilde{s}_{3,ut,zf} \\ \tilde{s}_{4,ut,zf} \end{bmatrix} \tag{6.22}$$

Where $X_{EA}$ and $h_{EA}$ are given by,

$$h_{EA} = |h_{1,1,f}|^2 + |h_{1,2,f}|^2 + |h_{2,1,f}|^2 + |h_{2,2,f}|^2 \tag{6.23}$$

$$X_{EA} = \frac{2RE(h_{1,1,f}h_{2,2,f} - h_{1,2,f}h_{2,1,f})}{h_{EA}^2} \tag{6.24}$$

Where $RE(.)$ denotes the real parte. From (6.22), and contrarily to the Alamouti scheme, each symbol interfere with other, e.g., symbol $\bar{s}_f$ interferes with $\bar{s}_{+3f}$.Two separate the data symbols two strategies are considered: one based on ZF criterion and other based on ML

Let's analyze first in ZF receiver. For this case to separate the data symbols, the signal given by (6.22) should be multiplied by $(\boldsymbol{G}_{EA})^{-1}$,i.e., $\tilde{\boldsymbol{s}}_{ut,zf} = (\boldsymbol{G}_{EA})^{-1} * \tilde{\boldsymbol{s}}_{ut}$ where $\boldsymbol{G}_{EA}$ is,

$$\boldsymbol{G}_{EA} = \begin{bmatrix} 1 & 0 & 0 & X_{EA} \\ 0 & 1 & -X_{EA} & 0 \\ 0 & -X_{EA} & 1 & 0 \\ X_{EA} & 0 & 0 & 1 \end{bmatrix} \tag{6.25}$$

Where $()^{H}$ and $()^{-1}$ denotes the transposed and inverse respectively, and $\tilde{\boldsymbol{s}}_{ut,zf}$ is a vector that contains data received and demodulated in UT by ZF receiver. If $\tilde{s}_{n,ut,zf} = s_f, n = 1,2,3,4$ the data was successful transmitted through all path. In practical systems this can be obtained when the links BS-to-RN and RN-to-UT have high quality

Now let's see the case of ML receiver,

From (6.22) we get,

$$\tilde{\boldsymbol{s}}_{ut,f} = |h_{EA}|^2 s_f + |h_{EA}|^2 X_{EA} s_{f+3} + noise \tag{6.26}$$

$$\tilde{\boldsymbol{s}}_{ut,f+1} = |h_{EA}|^2 s_{f+1} - |h_{EA}|^2 X_{EA} s_{f+2} + noise \tag{6.27}$$

$$\tilde{\boldsymbol{s}}_{ut,f+2} = |h_{EA}|^2 s_{f+2} - |h_{EA}|^2 X_{EA} s_{f+1} + noise \tag{6.28}$$

$$\tilde{\boldsymbol{s}}_{ut,f+3} = |h_{EA}|^2 s_{f+3} - |h_{EA}|^2 X_{EA} s_f + noise \tag{6.29}$$

So we see that $\tilde{\boldsymbol{s}}_{ut,f}$ and $\tilde{\boldsymbol{s}}_{ut,f+3}$ depends on $s_f$ and $s_{f+3}$, and $\tilde{\boldsymbol{s}}_{ut,f+1}$and$\tilde{\boldsymbol{s}}_{ut,f+2}$ depends on $s_{f+1}$ and $s_{f+2}$. Thus each par can be decoded separately with approximately the same complexity as the Alamouti scheme.

The rule to detect the pars$s_f, s_{f+3}$ can be given by,

$$\begin{aligned}(s_f, s_{f+3}) = \arg \underbrace{min}_{(s_f, s_{f+3})} \Big( & \big| |h_{EA}|^2 s_f + |h_{EA}|^2 X_{EA} s_{f+3} - \tilde{\boldsymbol{s}}_{ut,f} \big|^2 \\ & + \big| |h_{EA}|^2 s_{f+3} - |h_{EA}|^2 X_{EA} s_f - \tilde{\boldsymbol{s}}_{ut,f+3} \big|^2 \Big)\end{aligned} \tag{6.30}$$

## 6.3 Numerical Results

In order to evaluate the performance of the presented relay-assisted schemes we considered a typical pedestrian scenario, based on LTE specifications. The main simulation settings are summarized in Table IV. With the aim of having the same spectral efficiency between both cooperative and non-cooperative schemes, we used 2 different modulation: for the relay-assisted schemes we used QPSK and for the non-cooperative ones we used BPSK., i.e., in both systems one information bits are transmitted per data symbol interval.

<table>
<tr><td rowspan="1">General<br>Signal Definitions</td><td colspan="2">• FFT size: 1024;<br>• number of available carriers: 1024;<br>• sampling frequency: 15.36 MHz;<br>• useful symbol duration: 66.66 μs;<br>• cyclic prefix length: 5.208 μs;<br>• overall OFDM symbol duration: 71.87 μs;<br>• sub-carrier separation: 15 kHz;<br>• number of OFDM symbols per block: 12</td></tr>
<tr><td>Nodes' Antennas<br>Array Size</td><td colspan="2">• UT: 1 Rx antenna<br>• RN: 1 ou 2 Tx/Rx antennas<br>• BS: 2 Tx antennas</td></tr>
<tr><td>Channel Model</td><td colspan="2">• ITU pedestrian model B</td></tr>
<tr><td>Channel Equalization</td><td colspan="2">• Alamouti MRC<br>• Quasi-Orthogonal ZF<br>• Quasi-Orthogonal ML</td></tr>
<tr><td rowspan="3">Modulation and<br>coding schemes</td><td>Constellation</td><td></td></tr>
<tr><td>BPSK</td><td>For the non-cooperative schemes</td></tr>
<tr><td>QPSK</td><td>For the cooperative schemes</td></tr>
</table>

**Table IV: Main Simulation Parameters**

| **Scenario** | **$E_b/N$ Relations with $E_b/N_0$** | **Observations** |
|---|---|---|
| I | $E_b/N_R = E_b/N_{UT} = E_b/N_0$ | All links exhibit similar quality |
| II | $E_b/N_R = E_b/N_0$+10dB<br>$E_b/N_{UT} = E_b/N_0$ | BS has a good connection to the relays |
| III | $E_b/N_{UT} = E_b/N_0$+10dB<br>$E_b/N_R = E_b/N_0$ | Relays have good connection to the UT |
| IV | $E_b/N_R = E_b/N_{UT} = E_b/N_0$ +10dB | All cooperative links have better quality than direct link |

**Table V: Downlink Simulation Scenarios**

| **Link** | **$E_b/N$** |
|---|---|
| BS-to-RN | $E_b/ N_R$ |
| RN-to-UT | $E_b/N_{UT}$ |
| BS-to-UT | $E_b/N_0$ |

**Table VI : Eb/N Notations**

In order to assess proposed cooperative schemes in DL communication, we consider the different scenarios presented in Table V. The first one assumes that all links have the same quality (scenario I), i.e., $E_b/N_0= E_b/N_R= E_b/N_2$. In the second one, the quality of the BS-to-RN link is made 10 dB higher than RN-to-UT (and direct link), i.e., $E_b/N_R= E_b/N_2$+10dB. In the third scenario, the quality of the RN-to-UT link is made 10 dB higher than the BS-to-RN link, i.e., $E_b/N_2= E_b/N_R$+10 dB. In fourth scenario, the quality of BS-to RN and RN-to-UT links is both made 10dB higher than direct link, i.e., $E_b/N_2= E_b/N_R = E_b/N_0$ +10 dB. This fourth scenario is more realistic, since the relay links are assumed to have better quality.

Concerning the MIMO model two approaches are considered: in the first one we assume that the distance between antenna elements of each BS and relays (L=2) is far apart to assume uncorrelated channels, i.e. $\mathbf{R}_{Tx} = \mathbf{R}_{Rx} = \boldsymbol{I}$. In the second, the average angle of departure (DoA) set to $50^0$, standard deviation of DoA set to $8^0$, angle of arrival (AoA) set to $67.5^{0,}$ standard deviation of AoA set to $68^0$,and antenna spacing in the BS and relays set to half of wavelength. In order to simplify the channel model we used the same correlation matrix for all the taps.

The schemes considered in our evaluations are presented next:

- Non-cooperative SISO, 2x1 Alamouti and 4x1 ZF/ML Quasi-Orthogonal discussed in chapter 4. These schemes serve as reference.
- Cooperative 2x2x1 (relays equipped with single antenna), using distributed Alamouti space-frequency coding as discussed in section 6.2.1.
- Cooperative 2x4x1 (relays equipped with two antenna array), using distributed Quasi-Orthogonal coding as discussed in section 6.2.1. For this case the results were obtained considering the ML detector, since it has better performance than ZF with approximately the same complexity.

The results of the relay-assisted and non-cooperative schemes are presented in terms of the average BER as function of $E_b/N_0$, where $E_b$ is the received energy per bit and $N_0/2$ the bilateral power spectral density of the noise added at the BS in the direct link. We present the BER according various SNR values and with and without correlation in antennas system. Note that in all cases when we present the quasi-orthogonal cases we present the ML receiver performance instead of ZF. It’s because ML has a better performance such proves fig.26 and fig.27.

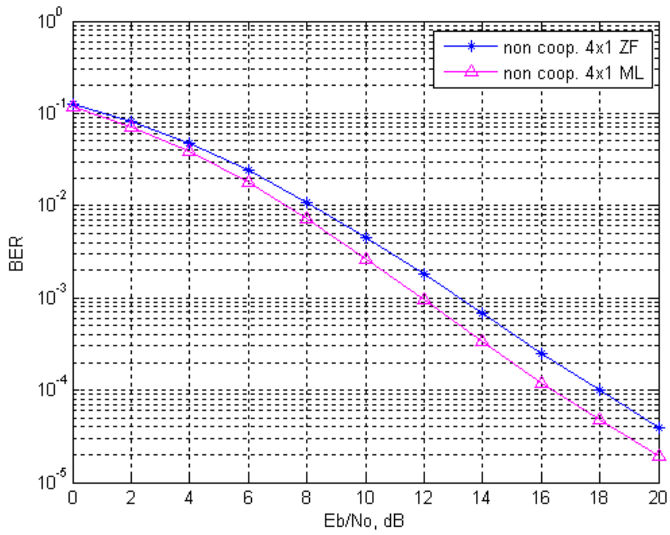


**Fig. 26 Performance comparison between ML and ZF algorithms for non-correlated antennas**

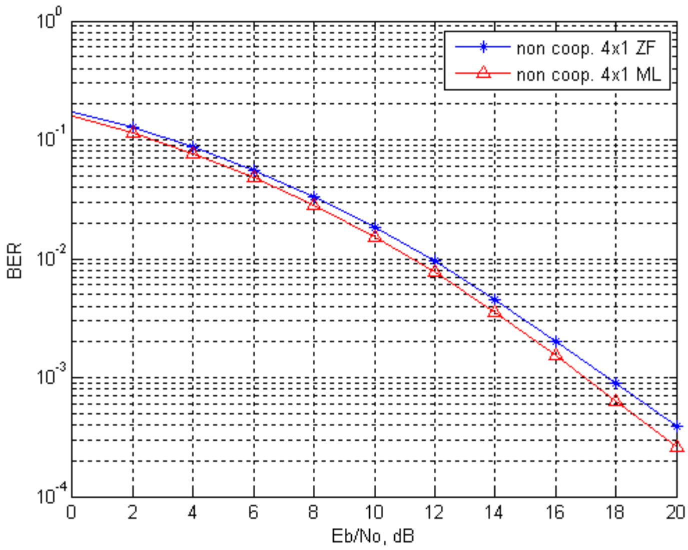


**Fig. 27 Performance comparison between ML and ZF algorithms for correlated antennas**

## Results with non-correlation in antennas:

So first we present the cases in which there is no correlation in antennas, and for non-cooperative schemes. From fig.28 we can see that SISO is obvious worse than another else because it represents the most simplest transmition scheme without any diversity. Comparing 2x1 and 4x1, we see than the use of four antennas array instead of two increases the system performance in terms of bit detection such proves fig.28. For

Alamouti a diversity order of 2 can be achieved while for 4x1 is higher than 2 but less than 4.

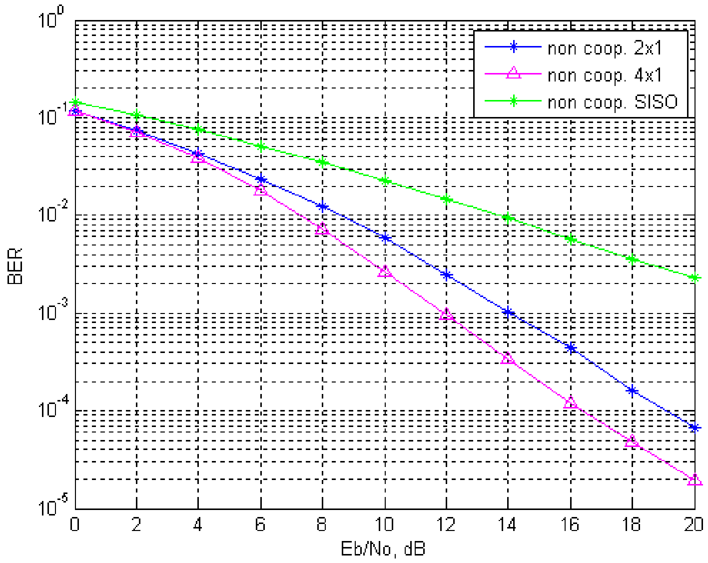


**Fig. 28 Performance comparison of non-cooperative schemes**

Fig. 29 presents the results for scenario I. From this figure we can observe that the non coop. 2x1 outperforms the correspondent coop. 2x2x1. This is because in the outage case when the relay fails to decode the data correctly, it cannot help the UT for the current cooperation round. In such case, the UT will get interference from the cooperative path and cooperation will not be beneficial. Comparing the non coop. 2x1 with the coop. 2x4x1, we can see that the performance is approximately the same. In this case, since the relay collect the signal by two antennas, the number of data symbols decoded with errors at the relays is less than for the case where the RS are equipped with single antenna. These results show the benefits of the relays be equipped with an antenna array

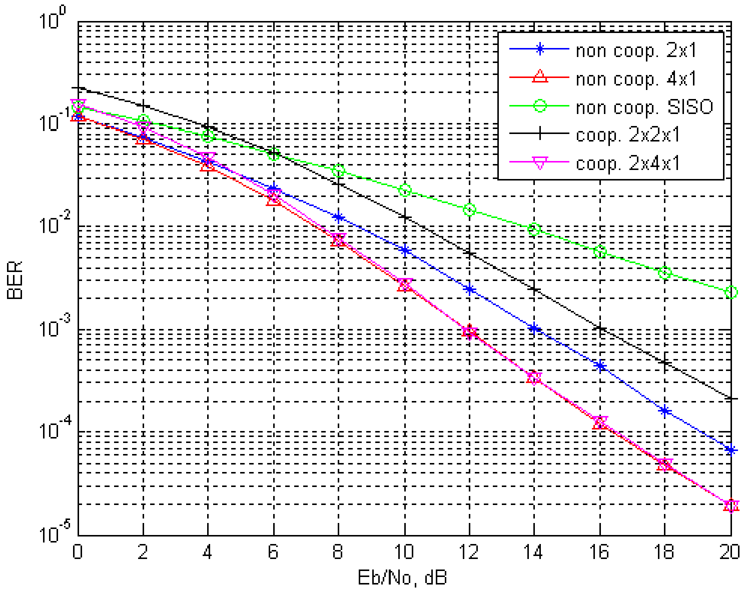


Fig. 29 Performance comparison of single and multiple antenna RA schemes in scenario I

Fig. 30 depicts the results for scenario II. The results show that for this scenario the performance of the non-coop. schemes is approximately the same of the correspondent coop. ones. This fact is explained by the fact that the quality of the BS->RS links are good quality and thus almost data symbols are successfully detected at the relays. Since in the second phase the quality of the RS->UT links are the same as the direct link, and is assumed that the antennas at the BS are uncorrelated, the performance is similar to the non-coop. approaches.

In fig.31 we present the results for scenario III. Thus results show that performance of coop.2x2x1 in almost the same like the correspondent non coop. 2x1. It's because transmition of data from relays to UT is very good taking into account the quality of RS->UT links, so we will only have errors in symbols detection at relays which will be the same like in non coop. 2x1 because we have one antenna relays. For its part, the 2x4x1 scheme outperforms all non coop.; Since the relay collect the signal by two antennas, the number of data symbols decoded with errors at the relays is low, and plus the high quality of RN->UT links make with the cooperation scheme have a very good performance.

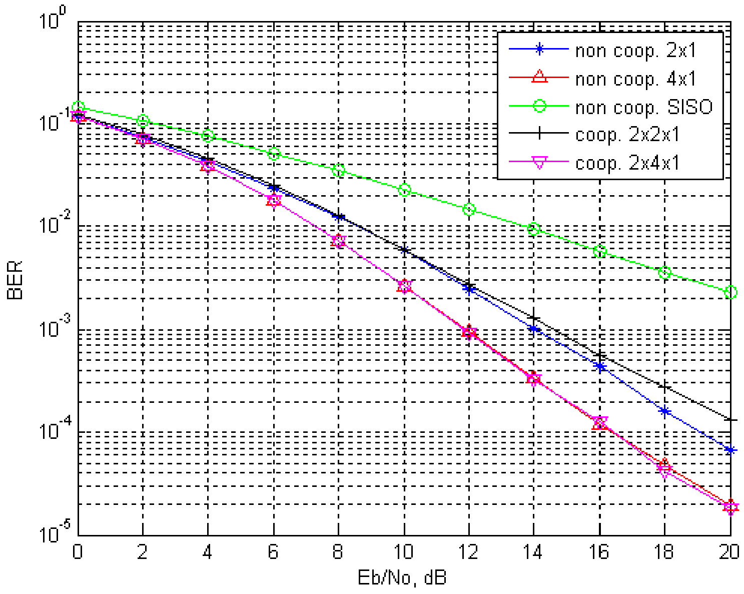


**Fig. 30 Performance comparison of single and multiple antenna RA schemes in scenario II**

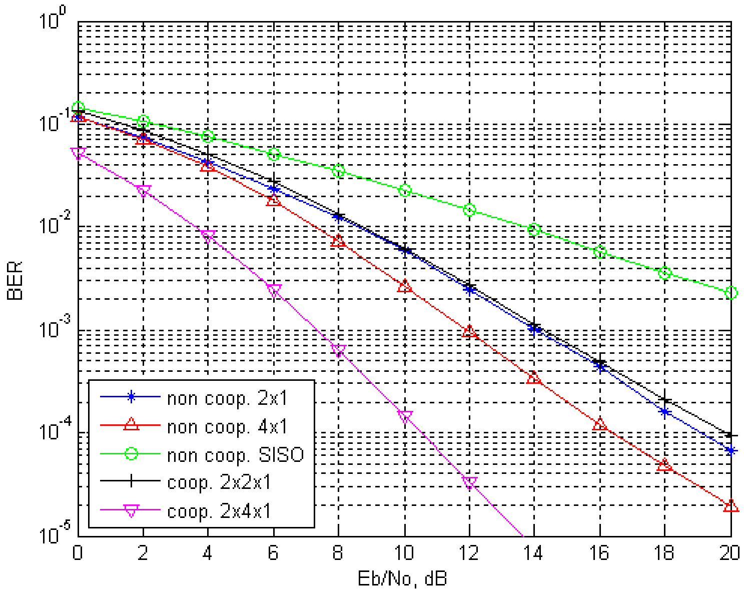


**Fig. 31 Performance comparison of single and multiple antenna RA schemes in scenario III**

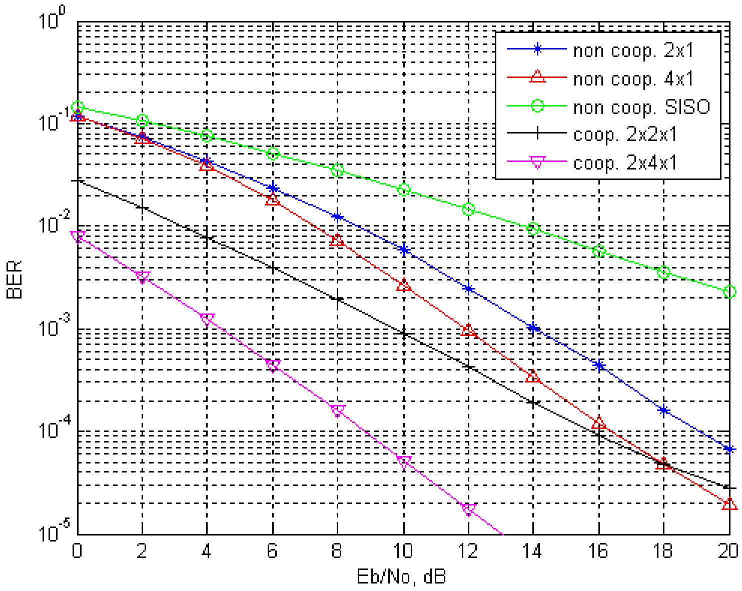


**Fig. 32 Performance comparison of single and multiple antenna RA schemes in scenario IV**

In fig. 32 we show the schemes performance under the definitions of scenario IV. The choice of this scenario derives from the fact that, in most real situations, the cooperative link has higher transmission quality conditions than the direct link. From this figure we can observe that the performance of both RA schemes is improved in comparison with the previous scenarios and that now, all cooperative schemes outperform non coop. ones. This is due to the fact that in the case that the links between BS, relays and UT are highly reliable, most information is successfully detected at the RN and next at the UT. Also note that between RA schemes, the 2x4x1 has a better performance than the 2x2x1. Once more, it shows the benefits of the relays be equipped with an antenna array.

**<u>Results with correlation in antennas:</u>**

For antennas with correlation scenarios, we will see a degrading relating to non-correlated ones. The correlation between the antennas will increase the probability of to have symbols interference. So for the curves will have a displace up, showing that system performance had got worse.

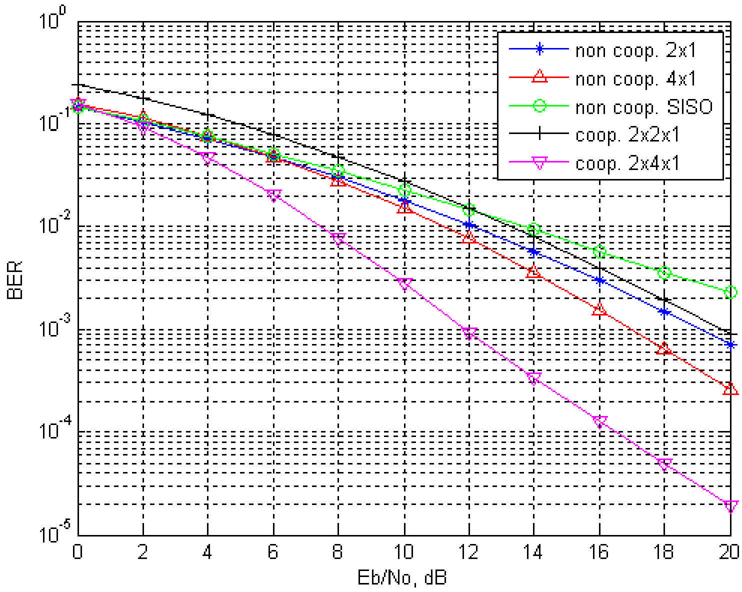


**Fig. 33 Performance comparison of single and multiple antenna RA schemes in scenario I**

Fig. 33 summarizes performance results for RA and non-cooperative schemes for scenario I. From this figure we can see that non coop. schemes are close to the SISO and close to one another. It's because the correlation degrades the gain of diversity. Also we see that coop. 2x2x1 is worse than SISO for some SNR values; the amount of decoded symbols errors caused by the correlation antennas in cooperative stages will be high, and cooperation will not improve the system behavior. However, and oppositely of what happened in scenario I for non-correlated antennas, coop. 2x4x1 scheme outperforms the non coop. ones. The explanation lies in the fact that relays are spatial uncorrelated. Although the correlation in BS and relays antennas degrades the system performance, the special diversity of the relays outperforms all of other schemes.

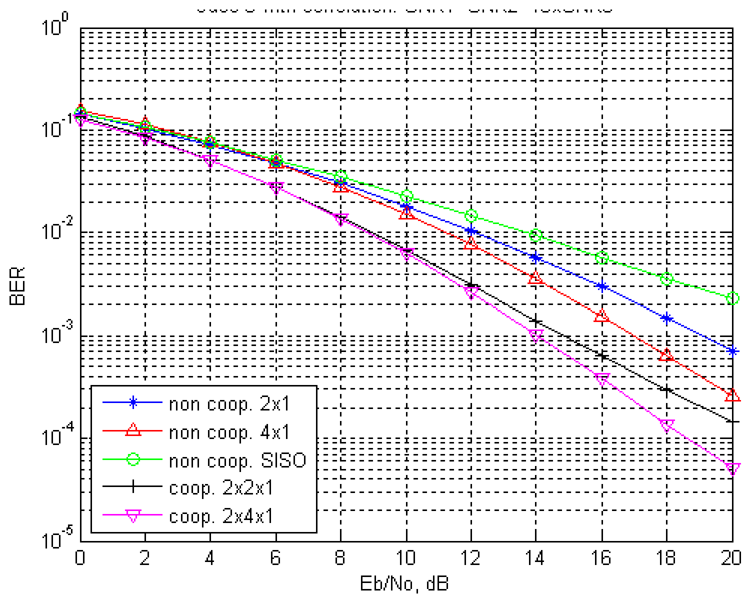


**Fig. 34 Performance comparison of single and multiple antenna RA schemes in scenario II**

Fig. 34 show the results of scenario II. We see that coop 2x2x1 had a improvement compared with previous scenario and the curve came close to coop. 2x4x1. Since the BS-to-UT link is improved, the hard-decoding process that takes place at the relay is more effective and less likely to produce decoding errors, and because relays have only one antenna, this scheme will not suffer the interference of correlation, contrarily that what happens in coop. 2x4x1 and also a space diversity gain in relays is achieved. In 4x2x1 scenario we cannot see a big improvement. It's explained by the fact that although the correlation in BS antennas, the MIMO transmition scheme between BS and RN can already provide a good transmition quality of symbols to relays.

In fig.35 we see the results for scenario III. The high quality of RN->UT links make with the amount of errors in hard-decoding of symbols occurs mostly in relays. In case of coop. 2x2x1 it makes with the perfomance coincides with the perfomance of non-coop 2x1 such proves the results. For coop. 2x4x1 the situation is different because, such we said in comments before, perhaps the errors caused by antennas interference, the MIMO transmition between BS and relays with $L$=2 provides a good symbols transmition to relays; It's makes with this cooperative scheme already have a good quality for this scenario.

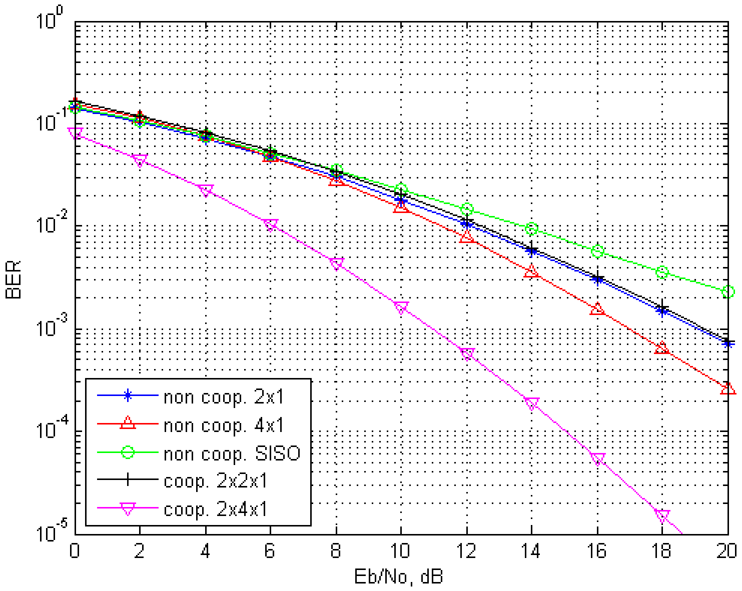


**Fig. 35 Performance comparison of single and multiple antenna RA schemes in scenario III**

Fig. 36 shows results for the case that the cooperative links have higher quality than the direct link. It suggests that for this scenario, all cooperative transmition has a better performance than the non-cooperatives. Also we can observe that the systems whose relay is equipped with an antenna array outperform the ones that use single antenna relays. One important repair to do is that for cooperative transmition, the curves are very similar with the analogous scenario with non-correlation in antennas. It suggests that correlation does not influence much the systems performance if we ensure that links quality is high.

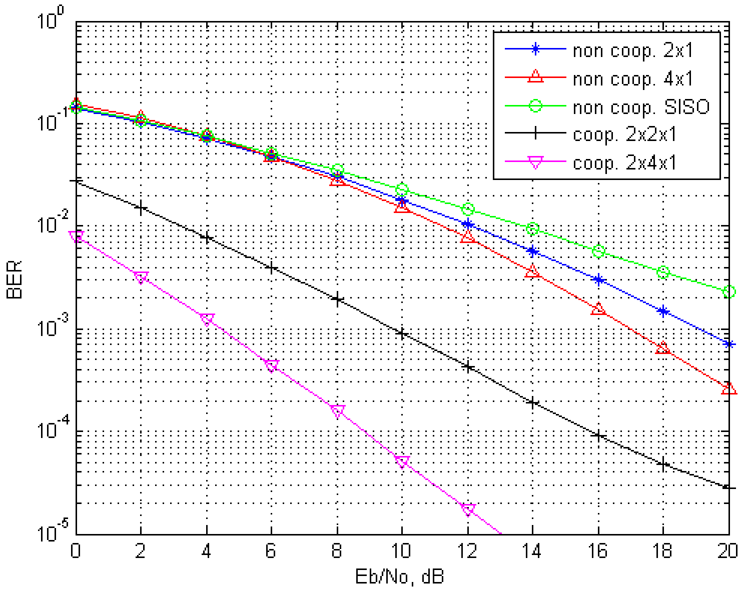


**Fig. 36 Performance comparison of single and multiple antenna RA schemes in scenario IV**

The aforementioned results suggest that the RA schemes are severely limited by the relative quality of the cooperative links, specially the BS-to-RN link. The antennas correlation has also importance in transmition quality, but as much as the link quality. Thus, one success points of the cooperative schemes lies upon the choice of the relay. Ideally, it should be chosen so that the BS-to-RN and RN-to-UT links are highly reliable and better that direct link.

# Chapter 7

## 7 Conclusions

The target of this dissertation was to study, implement and evaluate the performance of a cooperative diversity scheme for mobile communication systems, using two relays and two antennas for each relay.

To compare and evaluate performances, we proposed and evaluated single and multiple antenna relay-assisted schemes designed for Down-Link OFDM based systems, using decode and forward relay-assisted protocols. The schemes were evaluated under realistic scenarios based on LTE specifications and compared against the non-cooperative SISO, MISO and MIMO systems.

Regarding the impact of the links quality, and concerning in non-correlated antennas scenarios, numerical results show that the cooperative schemes do not have proved interesting in the case that cooperative links have the same quality like direct ones. In the case we have just BS->RN with high quality we verify that system performance is the same for both cooperative and non-cooperative schemes, and the non-cooperative wins because of low complexity. Scenario III and IV however, presents a good improvement in cooperative systems performance, outperforming clearly the non-cooperative ones.

Relating of correlated antennas the results show a different performance. In opposite of what happen in non-correlation antennas, for scenario I, the cooperative 2x4x1 outperforms such the non-cooperative such the coop. 2x2x1. The same happens in scenario III and it's because the transmition of data between BS and RN is done using a MIMO configuration $L=2$. This situation leads us to consider the importance of the availability of antenna arrays. For scenario II and IV clearly cooperative schemes outperforms the non-cooperatives, and this difference of performance quality is accentuated in IV, because of high quality of all links in cooperative path.

Concerned in the availability of antenna arrays at the relays, we can conclude that for all scenarios cooperative 2x4x1 outperform cooperative 2x2x1. It lead us to conclude that cooperative communication has better quality as bigger is the relay antennas array (at least until four array size).

Another interesting conclusion is that the correlation antennas parameter greatly influences the performance of the systems in different scenarios except in scenario IV, were performance for with and without correlation in antennas do not vary to much in the cooperative transmissions.

## Extensions and continuing work

While the results of this thesis presented good solutions to improve wireless communications quality, there are many more issues that remain to be addressed. Projects like CODIV and IEEE 802.16j are already trying to develop innovative scenarios.

- In this thesis, the mobile terminal was assumed to have only one antenna and not an antenna array. But it can also happen that the UT is a laptop or even a PDA. In this case, the availability of antenna arrays is more likely, would be interesting to study this scenarios performance.
- In this theses only DF protocol was used at the relays, would be interesting to compare the performance with other cooperative protocols, e.g.,AF.
- It was also assumed perfect channel information at both relays and UT to perform channel equalization. In practice the channels are estimated with some errors. It will be interesting assess the proposed cooperative schemes under imperfect channel estimates.
- Also could be pertinent try to design and simulate transmission chains where the relay is equipped with an antenna array with more than 2 antennas. In addition, could also be interesting focus in a more realistic scenario in which BS, RN and UT are all in movement. These scenarios are the ones that the IEEE 802.16j working group is exploring.

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