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Nonlinear Amplifier Distortion in Cooperative OFDM Systems.

The thesis has been submitted for official examination for the degree of Master of Science in Espoo, Finland on February 22nd, 2008.

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TEKNILLINEN KORKEAKOULU	Diplomityön tiivistelmä

Tekijä:	Jose Victor del Razo Sarmina
Työn nimi:	Epälineaarinen vahvistussärö yhteistoiminnallisissa OFDM-järjestelmissä
Päivämäärä:	22. helmikuuta 2008 **Sivumäärä:** 115
Tiedekunta:	Sähkö- ja tietoliikennetekniikan tiedekunta
Professuuri:	Signaalinkäsittelytekniikka
Työn valvoja:	Professori Risto Wichman
Työn ohjaaja:	TkT Fernando Gregorio

OFDM (Orthogonal frequency division multiplexing) on lupaava langattoman tietoliikenteen teknologia johtuen sen hyvästä suorituskyvystä monitieympäristössä. Yhteistoiminnallisen tiedonvälityksen tekniikka on nykyisin jatkuvan tutkimuksen kohteena. Se hyödyntää muiden päätteiden antenneja virtuaalisen moniantennijärjestelmän luomiseen mahdollistaen moniantennijärjestelmille ominaisia kapasiteettihyötyjä.

Tässä diplomityössä tutkitaan epälineaarista vahvistussäröä, kun näitä molempia tekniikoita käytetään yhdessä. Ensimmäiset kappaleet käsittelevät OFDM-järjestelmien ja epälineaaristen OFDM-järjestelmien särön sekä yhteistoiminnallisen tiedonvälityksen taustoja. Yhteistoiminnallisten OFDM-järjestelmien suorituskykyä mitataan simulaatioiden avulla epälineaarisen särön vaikuttaessa. Suorituskykyä mitataan bittivirhesuhteena käyttäen epäyhteistoiminnallista ja lineaarista yhteistoiminnallista järjestelmää vertailukohteena. Lisäksi särötermi myös analysoidaan. Systeemimalli sisältää epälineaarisen vahvistuksen välittimessä, jota mallinnetaan elektronisella tehovahvistimella.

Lopuksi esitellään ja testataan tekniikka järjestelmän suorituskyvyn parantamiseen optimoimalla maksimisuhdeyhdistintä. Se optimoidaan mallintamalla vahvistussäröä normaalijakaumalla. Lisäksi esitellään ja testataan yhteistoiminnallisille järjestelmille sopiva tehovahvistimen epälineaarisuuden poistotekniikan muunnelma, jolla saadaan lähellä lineaarista tapausta olevia tuloksia.

Avainsanat: Tehovahvistin, epälineaarisuus, OFDM, yhteistoiminnallinen viestintä, välitin, vahvistus ja edelleenlähetys.

HELSINKI UNIVERSITY OF TECHNOLOGY Abstract of the Master's Thesis

Author:	Jose Victor del Razo Sarmina
Name of the Thesis:	Nonlinear amplifier distortion in cooperative OFDM systems.
Date:	February 22, 2008 **Pages:** 115
Faculty:	Faculty of Electrical and Communications Engineering
Professorship:	Signal Processing
Supervisor:	Professor Risto Wichman
Instructor:	Dr. Fernando Gregorio

Orthogonal frequency division multiplexing (OFDM) is a promising technique for wireless communications because of its good performance under multipath environments. The concept of cooperative communications is currently under constant research. It uses antennas of other terminals to create virtual multiple input multiple output (MIMO) systems, providing capacity gains similar to those of MIMO systems.

This thesis studies the issue of nonlinear amplifier distortion when these two techniques are used together. The first chapters give a background on OFDM systems, nonlinear distortion in OFDM systems, and Cooperative Communications. The performance of OFDM cooperative systems under nonlinear distortion are measured by simulations. The performance is measured in terms of BER using a non-cooperative system and a linear cooperative system as references. In addition, the distortion term is also analysed. The system model includes a nonlinear amplifier at the relay, modelled as a solid state power amplifier (SSPA).

A technique for improving the performance of the system, by optimising the maximum ratio combiner (MRC), is introduced and tested. The MRC is optimised by modelling the distortion noise as Gaussian. Also, a modification to the power amplifier nonlinearity cancellation (PANC) technique, suitable to cooperative systems, is introduced and tested, showing results close to the linear case.

Keywords:	Power amplifier, nonlinearity, OFDM, cooperative communications, relay, amplify and forward.

Preface

The research work for this thesis was carried out in the Signal Processing Laboratory of Helsinki University of Technology in 2007.

I wish to express my gratitude to my supervisor, Professor Risto Wichman, for his continuous support and guidance and to my instructor, Dr. Fernando Gregorio, for his time, patience and advice. Special thanks to Dr. Stefan Werner for his constant interest and valuable help.

I would like to thank the people in the laboratory, especially to Mobien who always had time to answer to my questions and to Taneli for his interest in the topic and his advice. Thanks to the people at Bitville, particularly Antti Keurulainen for his support and flexibility when needed. Special thanks to Geoff White for the language comments on the text.

I want to thank my family, especially my parents and brothers for all their encouragement towards my decision of coming to Finland. Thanks to my mother for all the phone calls, the moral and sometimes economical support, and for her constant care. Thanks to Carlos and Mauricio for their continuous motivation and their visits on key moments. Thanks to my father for his advice and example. Special thanks to Magrande, Ricardo and Zahia for their everlasting interest in my well-being and their expressions of love. Thanks also to Doc for his constant support and his care towards us.

I want to thank my friends that made these years in Finland unforgettable. Andrea, who was a great study partner and even a better friend, deserves a special mention. My deepest gratitude to: Anna for her care and the trips and original activities; to Anel and Arto for the relaxed dinners and emergency housing; to Friederike for the motivation to keep going, her sweetness and the good food (particularly the desserts); to Laia for the nice company and a new concept of order; to Micke for redefining fun and friendship; to Niina and Poncho for their (atomic) energy and kindness; and to Virpi for the good time, long talks and lessons on finance. They are my Finnish family who helped winters to be warmer and summers to be cooler.

I also want to thank Evi Wager for her advice and support during this time. Special thanks to Carlos Martinez who, although far, was always there with wise words.

Helsinki, MMM dd$^{\text{th}}$, 2007

Victor del Razo

Contents

List of Acronyms

ADSL	Asymmetric Digital Subscriber Line
AF	Amplify and Forward
AM	Amplitude Modulation
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
CAR	Clipping Amplitude Recovery
CC	Coded Cooperation
CF	Compress and Forward
CP	Cyclic Prefix
CSI	Channel State Information
DAB	Digital Audio Broadcasting
DFT	Discrete Fourier Transform
DVB	Digital Video Broadcasting
FFT	Fast Fourier Transform
HPA	High Power Amplifier
IBI	Inter Bin Interference
ICI	Inter Channel Interference
IDFT	Inverse Discrete Fourier Transform
IFFT	Inverse Fast Fourier Transform
ISDB	Integrated Services Digital Broadcasting
ISI	Inter Symbol Interference
LO	Local Oscillator

LOS	Line Of Sight
MAC	Medium Access Control
MCM	Multi Carrier Modulation
MIMO	Multiple Input Multiple Output
MRC	Maximum Ratio Combiner
NLD	Non Linear Distortion
OBO	Output Back Off
OFDM	Orthogonal Frequency Division Multiplexing
PA	Power Amplifier
PANC	Power Amplifier Noise Cancellation
PAPR	Peak to Average Power Ratio
PM	Phase Modulation
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
SEL	Soft Envelope Limiter
SLM	Selective Mapping
SNR	Signal to Noise Ratio
SSPA	Solid State Power Amplifier
TWTA	Travelling Wave Tube Amplifier
WDM	Wave Division Multiplexing
WLAN	Wireless Local Area Network
WMAN	Wireless Metropolitan Area Network

List of Figures

List of Tables

Chapter 1

Introduction

Wireless communications are experiencing a fast evolution towards higher data rates and higher versatility in the services. Current offer of wireless services are already close to those available for fixed-line communications. Equipment manufacturers and network operators are constantly facing the challenge of increasing the available services while minimising costs to remain competitive. For doing so, an efficient use of the available resources and the reduction in the cost of the equipments have become essential issues.

1.1 Motivation

Orthogonal Frequency Division Multiplexing (OFDM) is one of the modulation techniques that have been used to achieve good performance at lower costs. This, because OFDM modulation has proved to very efficient for handling multipath environments at a low cost due to its spectral efficiency and simplicity of implementation and equalisation. For these reasons, OFDM is a big candidate for a large number of future technologies.

On the other hand, Cooperative Communications systems have recently been under investigation. In Cooperative Communications, neighbour devices are used as relays that cooperate with the transmitter and the receiver to provide diversity. These systems are expected to offer capacity gains similar to those of Multiple Input Multiple Output (MIMO) systems without the cost and complexity of increasing the number of antennas in mobile devices. This is achieved by using neighbour devices as virtual antennas.

A major disadvantage of OFDM systems is its sensitivity to amplifier nonlinearities. This problem has been studied and some solutions are already available. However, most of these solutions are complex and expensive to implement in mobile devices, where resources are more limited than in equipments connected to the network backbone. This means that the nonlinearity problem in Cooperative Communication systems has to be approached in a different way. So far, information available on this issue is limited.

1.2 Objective and Scope

This thesis has as a goal to analyse the effects of amplifier nonlinear distortion (NLD) in OFDM Cooperative Systems. By determining how the NLD affects this kind of systems, it is possible to evaluate if OFDM systems are good candidates for cooperation techniques.

The objective is, once the effects of NLD are known, to evaluate if it is possible to compensate the effects using this information. Knowing whether the performance of OFDM Cooperative Systems under NLD is close to the linear case (or at least significantly better than the non-cooperative case) brings these systems closer to their actual implementation.

The thesis is focused on a Cooperative System with a single Amplify-and-Forward relay. The modulation used is OFDM and no channel-coding or interleaving is applied. All channels involved in the system are considered to be independent and

Rayleigh fading.

1.3 Organisation of the Thesis

The thesis is organised in two main parts. The first part, that includes chapters 2-4, contains the relevant theoretical background available in the literature. The second part, chapters 5-8, consists of the contribution of this thesis for the analysis and mitigation of NLD in OFDM Cooperative Communications.

Chapter 2 presents the basic theory of OFDM. It also gives a good summary of the potential of this technique and the challenges of its implementation. In chapter 3 the problem of NLD in OFDM systems is explored in more detail. The chapter explains why OFDM systems are so sensitive to NLD and presents the existing techniques to identify it and mitigate it. Chapter 4 presents the theory behind Cooperative Communications. It reviews the methods and components of these systems and describes the details of the cooperative system to be used for the rest of the document.

In chapter 5, the system model is presented together with a brief description about how the simulations are performed. In this chapter the negative effects of NLD in OFDM Cooperative Systems are demonstrated and analysed. The causes of these negative effects are detected.

Chapter 6 presents an analysis of the relay. This includes how the signal arrives to the relay, how it is amplified and then forwarded to the destination. The effects of the position of the relay are also briefly discussed. In this chapter, the NLD is modelled and this model is used to optimise the maximum ratio combiner (MRC) at the destination; improving the results significantly.

Chapter 7 introduces a modification to the existing Power Amplifier Nonlinearity Cancellation (PANC) technique to be applied to cooperative systems. The results of applying this technique are also analysed in this chapter.

In chapter 8, the proposed solutions for NLD mitigation, as well as the system model used, are analysed. In this chapter the practicalities and challenges involving these solutions are described.

Finally, chapter 9 presents the conclusions of this thesis and some ideas for future work.

Chapter 2

OFDM Systems

Orthogonal Frequency Division Multiplexing (OFDM) is a multi-carrier transmission technique that has been recently used in different communication applications; for example, digital radio and television, wireless networking systems, and high data-rates over the fixed phone line. The principles behind OFDM modulation have been existing since several decades. However, in recent years OFDM has been implemented in modern communication systems and has proved to be a strong candidate for future developments. This chapter presents the basic concepts of OFDM and important practicalities about its implementation. The chapter explains also how the OFDM systems are influenced by the environment and presents the tools for optimising these systems. In addition, practical applications together with the most relevant advantages and disadvantages of OFDM systems are presented.

2.1 Basics of OFDM

OFDM can be seen as a hybrid of multi-carrier modulation (MCM) and frequency shift keying (FSK) modulation [1]. In MCM the data stream is divided into several parallel bit streams. In other words, a serial high-rate data stream is converted into multiple parallel low-rate sub-streams. Each sub-stream is then modulated using individual carriers or sub-carriers. Figure 2.1 shows an example of an MCM system with four sub-carriers.

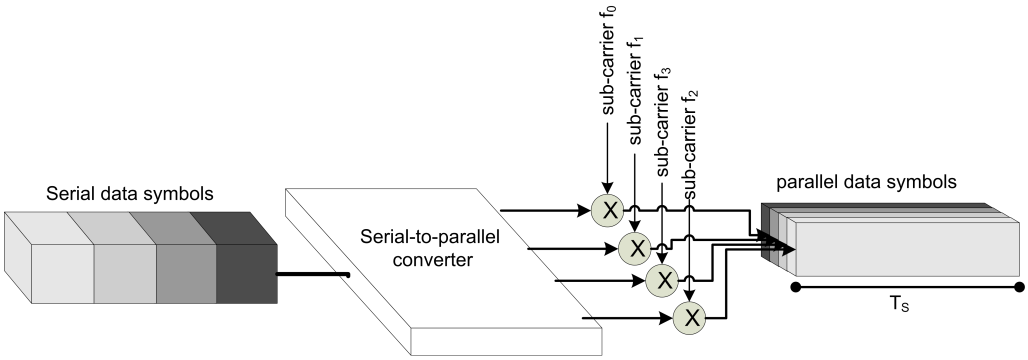

Figure 2.1: Multi-carrier modulation with $N_c = 4$ sub-channels.

When using parallel carriers, the symbol rate on each sub-carrier is significantly lower than the initial symbol rate, therefore the effects of delay spread, e.g. Inter-Symbol Interference (ISI), decrease. Among other advantages, the complexity of the equalizer is reduced [2].

FSK modulation is a modulation technique in which data is transmitted on one carrier from a set of orthogonal carriers in each symbol duration. This orthogonality is achieved by separating the carriers by an integer multiple of $1/T_s$, where T_s is the symbol duration. With OFDM, all the orthogonal carriers are transmitted simultaneously, thus, the entire allocated channel is occupied by the sum of the narrow orthogonal sub-bands.

The use of orthogonal sub-carriers is one of the key factors in OFDM systems. When the sub-carriers are orthogonal, it is possible to recover the individual sub-carriers' signal even if their spectra overlap. This permits a more efficient use of the available bandwidth.

A communication system using multi-carrier modulation with N_c parallel sub-carriers transmits the complex symbols S_n, $n = 0, ..., N_c - 1$, at a given time interval. Denoting the source symbol duration of the serial data as T_d, the OFDM symbol duration T_s resulting from the serial to parallel conversion is:

$$T_s = N_c T_d \tag{2.1}$$

In OFDM the objective is to modulate the N_c sub-streams on sub-carriers with a frequency spacing of:

$$F_s = \frac{1}{T_d} \tag{2.2}$$

that as explained earlier is the required spacing to achieve orthogonality between the signals on each sub-carrier; on the assumption that rectangular pulse shaping is used [2].

One OFDM symbol consists of the N_c parallel modulated source symbols S_n. The complex envelope of an OFDM symbol can be expressed as:

$$x(t) = \frac{1}{N_c} \sum_{n=0}^{N_c-1} S_n e^{j2\pi f_n t},\ for\ 0 \leq t \leq T_s. \tag{2.3}$$

where the N_c sub-carrier frequencies are defined as:

$$f_n = \frac{n}{T_s},\ n = 0, \ldots, N_c - 1. \tag{2.4}$$

From (2.1), it is possible to rewrite f_n as:

$$f_n = \frac{n}{N_c T_d},\ n = 0, \ldots, N_c - 1.$$

Sampling the complex envelope $x(t)$ in 2.3 at a rate $1/T_d$ results in:

$$x_v = \frac{1}{N_c} \sum_{n=0}^{N_c-1} S_n e^{j2\pi \frac{nv}{N_c}},\ v = 0, \ldots, N_c - 1. \tag{2.5}$$

Equation (2.5) can be compared to the definition of the Inverse Discrete Fourier Transform (IDFT). This is a very important advantage of OFDM from the implementation point of view. The sampled sequence x_v is the IDFT of the source symbol sequence S_n. Therefore, an OFDM transmitter can be implemented using an IDFT or an IFFT (Inverse Fast Fourier Transform) block, thus simplifying the implementation. These transformations can be viewed as mapping data onto orthogonal sub-carriers. Figure 2.2 shows a block diagram of a simple OFDM system using IFFT and FFT. In practice, OFDM systems are implemented using a combination of FFT and IFFT blocks. These are mathematically equivalent to the DFT and IFFT respectively but computationally more efficient.

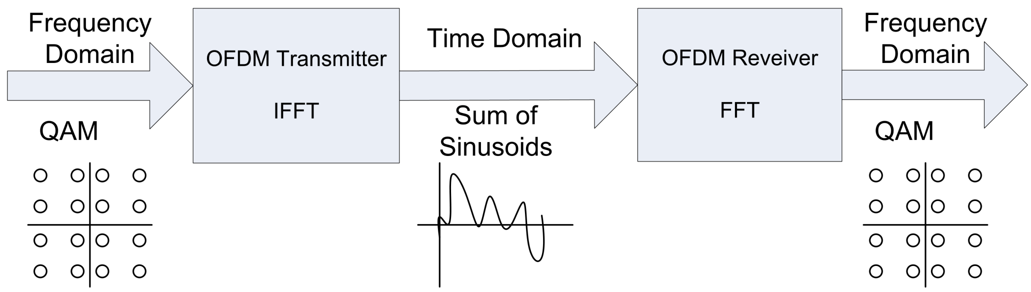

Figure 2.2: Block diagram of a simple OFDM system.

An OFDM system treats the source symbols at the transmitter as though they are in the frequency domain. These symbols are modulated in a complex envelope (e.g. M-QAM or M-QPSK) and are used as the input for the IFFT. The IFFT block transforms the signal into the time domain, N_c symbols at a time. Each of these inputs has a period of T_d seconds as mentioned before. As a result of using the IFFT, the time domain signal will be the sum of N_c sinusoidal functions with different but equally spaced frequencies (2.4). Taking this into consideration, each input symbol can be viewed as a complex weight for the corresponding sinusoidal function [3], determining both the amplitude and phase of the sinusoidal for each sub-carrier. Figure 2.3 shows the spectrum of an OFDM symbol together with the spectrum of the first sub-carrier. Ideally, the FFT block output will be the same original symbols that, when plotted in the complex plane, will form a constellation. This means that the detection of the symbol e.g. symbol slicing must be done in the frequency domain.

As the number of carriers N_c increases, the OFDM symbol duration T_s increases. When T_s becomes large compared to the duration of the channel impulse response

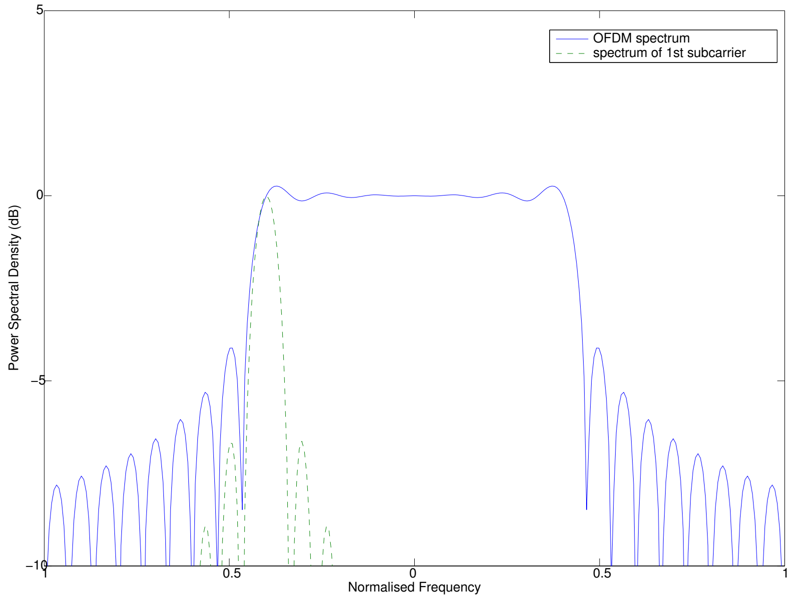

Figure 2.3: Spectrum of an OFDM symbol with 16 sub-carriers

τ_{MAX}, the amount of Inter-symbol Interference (ISI) is reduced significantly. However, given the importance of orthogonality between the signals on the sub-carriers, ISI must be completely avoided. The following section describes the importance of avoiding ISI and ICI (Inter-carrier Interference), as well as the techniques used to achieve this in OFDM systems.

2.2 Multipath Channels and the Cyclic Prefix

A major problem in most wireless systems is the presence of multipath propagation. In a multipath environment, the transmitted signal is reflected and diffracted by several factors, which include objects (obstacles) and terrain conditions, resulting in multiple delayed versions of the transmitter signal at the receiver. This situation is described in figure 2.4. The distortions caused by the wireless medium to the signal include delay spread, attenuation of the signal power, and frequency broadening [1].

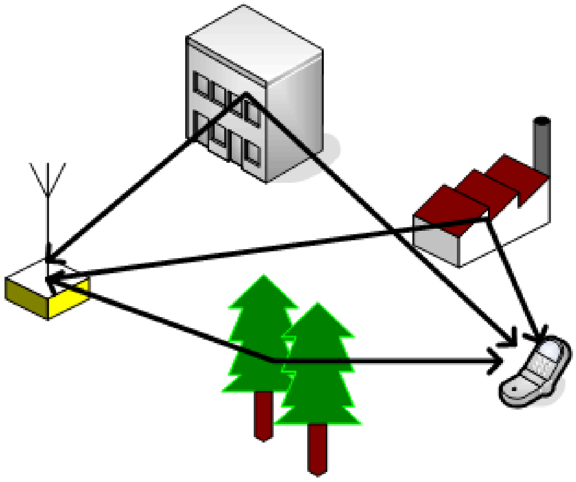

Figure 2.4: Multipath propagation

The channel characteristics vary in time in an unpredictable way. These time variations can be described by narrow-band random processes [4].When the number of signal reflections is large, the distortions can be modelled as complex Gaussian random processes, according to the central limit theorem. The envelope of the received signals is then a combination of two components: fast-varying fluctuations superimposed onto slow-varying fluctuations. When the mean of these envelopes suffers a significant strength reduction due to a destructive combination of the phase terms from each path, it is said that the signal is being faded.

In a fading channel, the relationship between the maximum excess delay time, τ_{MAX}, and the symbol time, T_s, defines two different degradation categories, frequency-selective fading and frequency non-selective or flat-fading [5]. A fading channel is frequency-selective when $\tau_{MAX} > T_s$, which means that the received multipath components of a symbol extend beyond the symbol duration. This multipath dispersion of the signal results in Inter-symbol Interference (ISI) which is referred as channel-induced ISI. A fading channel is frequency non-selective or flat-fading when $\tau_{MAX} < T_s$ which means that all the received multipath components of a symbol arrive within the symbol time duration. In this case, channel-induced ISI is not present but still there is a performance degradation. The multipath components can add up destructively yielding to a significant reduction in SNR. In OFDM systems, given the enlargement of the symbol duration, it is usual that $\tau_{MAX} < T_s$, therefore the wireless channel is in general considered as a flat-fading channel.

Another important factor to consider is how fast the statistics of the channel are changing. This is related to the channel coherence time. When the statistics of a channel remain constant during one or more symbol intervals T_s, the channel is considered to be a slow-fading channel. On the other hand, when the statistics of a channel vary within the duration of the symbol T_s, the channel is referred as fast-fading channel. In OFDM systems, T_s is considerably larger compared with single-carrier systems. Therefore it is possible that channels that are slow-fading for single-carrier systems can behave as fast-fading channels for OFDM systems. Whether modelling a channel as fast-fading or slow-fading depends on each situation.

2.2.1 The Channel Model

As mentioned above, it is possible to accurately describe a multipath channel as a random process, therefore the state of the channel is characterised by its correlation function. For multiple propagation paths to the receiver, a channel is characterised by time-varying propagation delays, attenuation factors, and Doppler shifts. The time-variant input response can be represented as [6]

$$c(\tau_n, t) = \sum_n \alpha_n(\tau_n(t)) e^{-j2\pi f_{D_n} \tau_n(t)} \delta[t - \tau_n(t)] \tag{2.6}$$

where $c(\tau_n, t)$ is the response of the channel at time t for an impulse applied at time $t - \tau_n(t)$; $\alpha_n(t)$ is the attenuation factor for nth path; $\tau_n(t)$ is the propagation delay for the nth path; and f_{D_n} is the Doppler shift for the signal received on the nth path. The Doppler shift depends on the relative motion between the transmitter and the receiver, and can be expressed as:

$$f_{D_n} = \frac{v \cos(\phi_n)}{\lambda} \tag{2.7}$$

where v is the relative velocity between the transmitter and the receiver, λ is the wavelength of the carrier, and ϕ_n is a random phase angle, uniformly distributed between 0 and 2π.

Equation (2.6) shows that $c(\tau_n, t)$ is the sum of Gaussian random processes. Therefore, the envelope of $c(\tau_n, t)$ has a Rayleigh distribution. The probability density function (pdf) of a Rayleigh fading channel is given by

$$f_Z(Z) = \frac{Z}{\sigma^2} e^{-\frac{z^2}{2\sigma^2}} \tag{2.8}$$

When there is a line of sight (LOS) between the receiver and the transmitter, then the envelope of $c(\tau_n, t)$ has a Rice distribution. A rician channel has a pdf

$$f_Z(Z) = \frac{Z}{\sigma^2} I_0 \left(\frac{Z\eta}{\sigma^2} \right) e^{-\frac{z^2+\eta^2}{2\sigma^2}} \tag{2.9}$$

I_0 is the modified Bessel function of order zero. The mean η is caused by the LOS or by the fixed scatters.

From the autocorrelation function for $c(\tau_n, t)$, $\Delta_c(\tau, \Delta t) = E\left[c(\tau, t)c^*(\tau, t + \Delta t)\right]$, the average output power of the channel can be obtained by setting $\Delta t = 0$ yielding to $\Delta_c(\tau)$. This is known as the delay power spectrum of the channel. In addition, the range of values of τ for which $\Delta_c(\tau) \neq 0$ is the multipath spread τ_{MAX}.

2.2.2 Effects of the Channel in OFDM Systems

The effects of a multipath channel causes two important problems for OFDM systems. The first problem is ISI, that occurs when the received OFDM symbol is distorted by the previously transmitted OFDM symbol. Unlike single-carrier systems, since the symbol period is longer, ISI in OFDM systems is caused only by one previous symbol. This happens because the OFDM symbol period T_s is usually longer than the time span of the channel τ_{MAX}.

As explained before, as a result of MCM, the OFDM symbol period is increased by a factor of N_c. The time span of the channel τ_{MAX} can be viewed in its discrete form as L_{ch} samples long. If $N_c > L_{ch}$ the effects of ISI will translate in the distortion of the first L_{ch} samples of the received OFDM symbol. Therefore, the use of a guard interval $T_g \geq \tau_{MAX}$, or equivalently $L_g \geq L_{ch}$, is enough to remove the effects of ISI in the OFDM symbol.

The second problem is only present in multi-carrier systems. Intra-symbol Interference or Inter-carrier Interference (ICI) is the result of the interference amongst an OFDM symbol's own sub-carriers. The use of the guard interval removes effects of ISI but does not remove the effects of ICI. It is possible, however, to solve the problem of ICI in a relatively simple way.

In continuous-time, a convolution in time is equivalent to a multiplication in the frequency-domain. This applies to discrete-time signals only if they are of infinite length or if at least one signal is periodic over the range of the convolution. It is obvious that an infinite-length OFDM symbol is not practical. However, it is possible to make the OFDM symbol appear somehow periodic over the range of the convolution. To achieve this, a cyclic prefix (CP) is used to fill in the guard interval. A CP is a copy of the last part of the OFDM symbol. The CP is discarded at the receiver, thus eliminating the effects of ISI. In addition it makes the effect of the channel to become multiplicative.

The length of the guard interval can be expressed as:

$$L_g \geq \left\lceil \frac{\tau_{max} N}{T_s} \right\rceil \tag{2.10}$$

The addition of the CP and the guard interval can be expressed as:

$$S_{n-m} = S_{N_c+m-n}, \quad for\ n-m \leq L_g$$

As a result, the OFDM symbol period is now extended to $T_s' = T_g + T_s$ and the sampled sequence in (2.5) with the cyclic extended guard interval is:

$$x_v = \frac{1}{N_c} \sum_{i=0}^{N_c-1} S_n e^{j2\pi \frac{nv}{N_c}},\ v = -L_g, \ldots, N_c - 1. \tag{2.11}$$

This sequence is converted to continuous-time and transmitted through the channel. The transmitted signal waveform $x(t)$, before RF conversion, will be the same as in (2.3) with increased duration T_s'. At the receiver, after RF conversion, the output of the channel can be expressed as:

$$y(t) = \int_{-\inf}^{\inf} x(t-\tau)h(\tau, t)d\tau + n(t) \tag{2.12}$$

that is the convolution of $x(t)$ with the channel impulse response $h(\tau, t)$ plus the additive noise $n(t)$.

After analog-to-digital conversion, at a sampling rate $1/T_d$, a sequence y_v', $v = -L_g, \ldots, N_c - 1$ is obtained. The first L_g samples are redundant information containing the CP and they are affected by ISI. Therefore, they can be removed previous to multi-carrier demodulation. The demodulation of the ISI-free signal y_v, $v = 0, \ldots, N_c - 1$ is performed using a DFT (or an FFT) resulting in the output sequence:

$$R_n = \sum_{v=0}^{N_c-1} y_v e^{-j2\pi \frac{nv}{N_c}},\ n = 0, \ldots, N_c - 1. \tag{2.13}$$

Given the multiplicative effect of the channel due to the addition of the CP, each sub-channel can be considered separately. Furthermore, assuming the fading on each sub-channel is flat [2], the received symbols can be expressed as:

$$R_n = H_{n,i} S_n + N_n,\ n = 0, \ldots, N_c - 1. \tag{2.14}$$

where $H_{n,i}$ is the flat fading factor and N_n is the noise of the nth sub-channel. For simplicity, $H_{n,i}$, which is the nth coefficient of the channel transfer function at time i, will be expressed as H_n.

Equation (2.14) shows that each sub-carrier of the OFDM symbol is multiplied by a complex number that is the frequency response of the channel at the sub-carrier's frequency. As a result, a frequency-domain equaliser, which is easier to implement than time-domain equalisers, can be employed, thus simplifying the equalisation process significantly. A frequency-domain equaliser consists of a single complex multiplication for each sub-carrier.

Using matrix-vector notation, we can express the received symbol sequence as:

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{2.15}$$

where $\mathbf{s}$ is the vector representing the N_c transmitted symbols,

$$\mathbf{s} = (S_0, S_1, \ldots, S_{N_C-1})^T \tag{2.16}$$

$\mathbf{H}$ is the $N_c \times N_c$ channel matrix, that in the absence of ISI and ICI is a diagonal matrix containing the flat fading coefficients assigned to the sub-channels. This is, the first N_c coefficients of the Fourier Transform of the impulse response of the channel.

$$\mathbf{H} = \begin{pmatrix} H_0 & 0 & \ldots & 0 \\ 0 & H_1 & & 0 \\ \vdots & & \ddots & \vdots \\ 0 & 0 & \ldots & H_{N_c-1} \end{pmatrix} \tag{2.17}$$

The vector $\mathbf{n}$ represents additive noise

$$\mathbf{n} = (N_0, N_1, \ldots, N_{N_c-1})^T \tag{2.18}$$

and $\mathbf{r}$ is a vector containing the received symbols after demodulation

$$\mathbf{r} = (R_0, R_1, \ldots, R_{N_c-1})^T \tag{2.19}$$

The matrix $\mathbf{H}$ is a diagonal matrix. Therefore the estimation of the channel and its equalisation is simpler compared with single-carrier systems.

These characteristics make OFDM in theory a simple yet powerful technique for communication systems. However, in practice there are other factors that have to be considered and that increase the complexity of OFDM systems. The following section gives an overview of the non-ideal effects in OFDM systems.

2.3 Non-ideal Effects in OFDM Systems

Non-ideal factors, e.g. impairments and receiver offsets, affect the performance of OFDM systems significantly. This section briefly describes some of these factors. As each of these factors is by itself a topic for research, it is not possible to go into the details. Nevertheless, understanding these complications gives a better understanding of the challenges to be faced when using OFDM systems.

One factor affecting OFDM systems is the local oscillator (LO) frequency offset. At start-up, the LO frequency at the receiver is usually different from its counterpart at the transmitter. In general, a carrier tracking loop is used to match these frequencies. However, OFDM systems are especially sensitive to frequency offsets. The LO frequency offset results in a frequency shift of the received signal spectrum which causes a loss of orthogonality. This happens because the bins of the FFT do not line up with the peaks of the received signal. As a result, a distortion called inter-bin interference (IBI) occurs. IBI can be defined as the energy spilling of one bin over the adjacent bins. This IBI will manifest as additive Gaussian noise, thus lowering the effective SNR of the system. To solve this problem, the signal can be multiplied by a correction factor which is a sinusoid of frequency equal to the LO frequency offset.

Another factor related to the LO is the LO phase offset, which also affects OFDM systems. LO phase offset causes a constant phase rotation for all of the sub-carriers in the frequency domain. As a result, the constellation points in each sub-carrier experience the same effect. If the LO phase offset is small, i.e. phase rotation does not go beyond the decision regions, the equaliser can correct this offset. If the LO phase offset is larger, a carrier tracking loop is needed.

FFT window location offset is a factor inherent to OFDM systems. It happens when the FFT is performed over samples that are not belonging to the same OFDM symbol. The offset can be reduced by using robust synchronisation algorithms. In addition, the presence of the CP gives enough headroom so that a small offset does not mean taking samples from an adjacent OFDM symbol. The offset is in practice a shift in time. When this shift is not large enough to go beyond the OFDM symbol boundary, it is equivalent to a linearly-increasing phase rotation in the frequency-domain constellations. This means that the lower frequency sub-carriers experience a smaller phase rotation while the higher frequency sub-carriers experience a larger phase rotation; these can be handled by the equaliser. Otherwise, FFT window location offsets are often corrected by performing a time-domain correlation with a known training sequence included in the transmitted signal.

Another harmful situation is the presence of sampling frequency offset. This occurs when the analog-to-digital converter output is sampled either too fast or too slow. Sampling too fast results in a contracted spectrum and sampling too low results in an expanded spectrum. Either type of sampling frequency offset results in IBI. This happens because in both cases the received sub-carriers are not able to line-up with the FFT bin locations. The sampling frequency offset can be corrected by generating an error term that is used to drive a sampling rate converter.

Phase noise is another factor that should be carefully considered in OFDM systems. Phase noise is added to the signal in the frequency-conversion stage. This occurs because the oscillator in the converter has inherently some phase noise, that can be viewed as the uncertainty of the actual frequency or phase of the signal. The phase noise in the oscillator is transferred to the signal. It is primary concentrated near the signal's central frequency. Since an OFDM signal contains multiple sub-carriers and each sub-carrier is affected by a phase noise, the performance degradation of an OFDM system is significantly higher than in the case of single carrier systems. The phase noise results in phase uncertainty in the constellation points in each sub-carrier. To handle this effect, pilot sub-carriers are generated by the IFFT and can be used to provide a stable phase reference to the receiver.

In addition to these factors, there is an additional cause of distortion which is the main topic for this work. Nonlinearities in the power amplifiers, used in the RF sub-system, can cause severe distortion of OFDM signals. This topic is discussed in detail in Chapter 3.

So far, the most relevant concepts and elements of OFDM, concerning this work, have been presented. However it is good to have a general idea of all the elements involved in an OFDM system. The following section gives an overview of a complete ODFM system.

2.4 The OFDM Transceiver Structure

In addition to the elements explained so far, the OFDM transceiver structure includes several blocks that play an important role in the performance of the system. This section presents these blocks and briefly goes through the functions and importance of some of them. Figure 2.5 presents a general block diagram of an OFDM transceiver.

The input data of the transmitter is the information to be sent and usually comes from a higher protocol layer. This data is first coded by the channel encoder for error detection and error correction purposes. Channel codes are generally defined by their coding gain and code rate. The coding gain is the reduction of the required E_b/N_o to achieve the same BER compared to the case when channel coding is not used. The code rate is the ratio of the input bits to the output bits of the encoder.

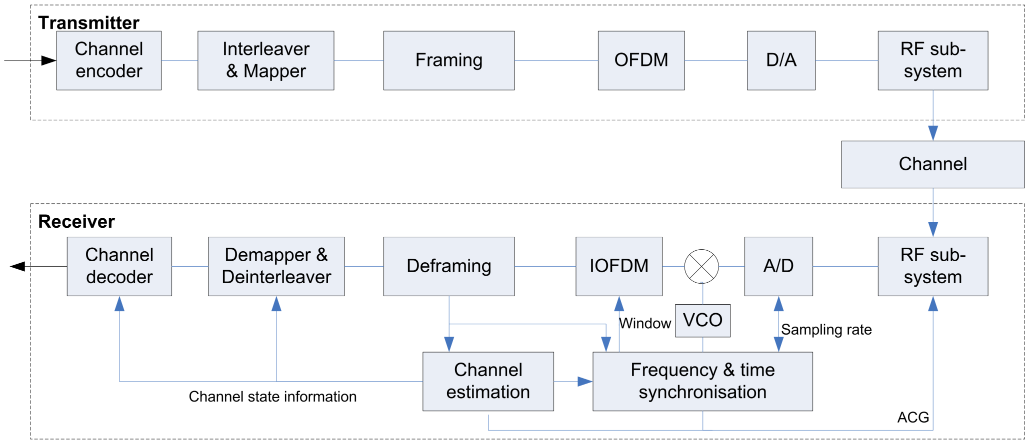

Figure 2.5: General block diagram of an OFDM transceiver

The coding gain denotes the reliability of the code while the code rate denotes the cost, in terms of bandwidth, of using the code. A description of the codes that are commonly used in ODFM systems can be found in [1].

In OFDM, as in other communication systems, channel coding is usually combined with interleaving. In wireless communications, errors due to the channel characteristics usually occur in several consecutive bits. The interleaver distributes the input bits either in time or frequency. By doing so, the error bits are also distributed making it easier for the channel decoder to correct these errors, improving the overall system performance. The use of an interleaver introduces certain delay that has to be considered. More detailed information about interleaving in OFDM systems can be found in [1].

The mapper or modulator maps the input bits to a symbol constellation, depending on the modulation scheme and size of the constellation. By far, QPSK and M-QAM are the most widely used modulation schemes in OFDM. The choice of the modulation scheme affects the BER, the Peak-to-Average Power Ratio (PAPR), and the RF spectrum shape. The larger the constellation is, the shorter the distance between the points of the constellation. The minimum distance d_{min} determines the amount of noise needed to generate a decision error. Also, depending on the phase-lock characteristics of the system, a modulation can be either coherent or non-coherent. More information about modulation schemes used in OFDM is available in [2] and [1].

After mapping the bit stream into symbols, these last are then framed or grouped. The size of the frame depends on the number of sub-carriers N_c of the OFDM system. The frame is then sent to the OFDM block which makes the serial to parallel conversion, performs the IFFT as explained in section 2.1 and makes the parallel to serial conversion. Later, an digital-to-analog conversion is performed and the output is sent to the RF sub-system. The RF sub-system processes the signal for transmission and includes pass-band modulation and amplification.

At the receiver side, the opposite operations are performed. In addition, blocks for the channel estimation as well as for the synchronisation are included. The importance of the synchronisation blocks and its accuracy for OFDM systems is explained in section 2.3. The channel estimation normally uses the principles explained in section 2.2 and it is proved to be simpler than in other communication systems.

2.5 OFDM Applications

So far, the concepts and practicalities of OFDM systems have been presented. It is clear that OFDM systems provide great advantages compared to single-carrier systems. However, it is also true that its implementation requires tight specifications for the elements to be used. Clearly, OFDM is not a solution for every communication system but in applications where the cost of a tighter implementation is worth the increase in performance, OFDM is with no doubt a clear option.

OFDM applications include a broad set of both wireless and wire-line applications. Maybe the most important wire-line technology in which OFDM is used is Asynchronous Digital Subscriber Line (ADSL), where it is better known as Discrete Multitone Modulation (DMT). Other applications in which OFDM is considered are in Wave Division Multiplexing (WDM) fibre optics and cable modems.

In the wireless scenario, OFDM is used in Digital Audio Broadcasting (DAB), Digital Video Broadcasting (DVB), Integrated Services Digital Broadcasting (ISDB), some Wireless LANs (WLAN), and Wireless MANs (WMAN) among others. In these techniques OFDM has resulted useful due to its performance in multipath environments and bandwidth efficiency.

DAB, for example, uses Coded OFDM (COFDM) with 1,536 carriers within the 1.5 MHz band. COFDM is an implementation of OFDM that includes coding for error detection and correction. DAB uses D-QPSK modulation.

In DVB, OFDM is considered in DVB-T (terrestial) and DVB-H (handheld). In DVB-T, OFDM is used for the transmission within the existing UHF and VHF frequency bands. The constellations used are BPSK, QPSK, 16-QAM and 64-QAM. Outer and inner coding and interleaving is used for the OFDM frame containing 68 OFDM symbols. Each OFDM symbol has either 1705 (1512 active) or 6817 (6048 active) sub-carriers, depending on the operation mode (2k and 8k respectively).

Regarding WLANs, OFDM is considered for IEEE 802.16 a and g systems, HyperLAN/2 and MMAC. In the IEEE 802.16 for example, 52 carriers with 4 pilot tones are used within the 5GHz band. This technology delivers a bitrate up to 54 Mbps. The constellations used are BPSK, QPSK, 16-QAM and 64-QAM.

Another significant example is the Broadband Wireless Access Systems (BWAS), based on the IEEE 802.16 WMAN. BWAS deliver voice, data, internet and video services within the 25GHz band. Modulation is performed using QPSK, 16-QAM or 64-QAM. In this case, OFDM allows the allocation in multiple frequencies in bands between 2 and 66 GHz. It uses 2048 carriers of which 1536 contain effective data.

ODFM is also considered in 3G Long Term Evolution (LTE) mobile systems mainly for the downlink. In addition, Fast Low-latency Access with Seamless Hand-off OFDM (Flash-OFDM) is being considered as a strong candidate due to its cross layer optimisation.

After reviewing the basic idea behind OFDM and its implementation, together with the applications in which OFDM is being considered, it is possible to define some advantages and disadvantages of OFDM systems [1]:

Advantages:

- High spectral efficiency due to nearly rectangular frequency spectrum for high numbers of sub-carriers.
- Simple digital realisation by using FFT operation.
- Low complex receivers due to the avoidance of ISI and ICI.
- Flexible spectrum adaptation can be realised.
- Different modulation schemes can be used on individual sub-carriers according to the transmission conditions of each of them.

Disadvantages:

- Multi-carrier signals with high peak-to-average power ratio (PAPR) require highly linear amplifiers.
- Loss of spectral efficiency resulting from the use of the guard interval.
- More sensitivity to Doppler spreads than single-carrier modulated systems.
- Phase noise caused by the imperfections of the transmitter and receiver oscillators influence the system performance.
- Accurate frequency and time synchronisation is required.

2.6 Summary of the Chapter

In this chapter the concept of OFDM and the basic ideas behind it were presented. The use of several sub-carriers result in longer symbol periods which is useful to avoid ISI.

The implementation of OFDM systems results very efficient due to the use of IFFT and FFT algorithms. By exploiting the properties of the Fourier transform and the addition of the cyclic prefix, it is possible to implement very reliable systems with high immunity to ICI and ISI.

However there are some issues to be considered which were explained in section 2.3 that include the need of very accurate frequency and time synchronisation. One of the main complication of OFDM is its high PAPR which makes it very sensitive to nonlinearities in the amplifier. This topic is explained in detail in next chapter.

In addition, an overview of a complete OFDM transceiver was presented in section 2.4 together with a brief explanation of the importance of its blocks. In section 2.5 some current and future applications were described and the main advantages and disadvantages of these systems were addressed.

Chapter 3

Nonlinear Distortion in OFDM Systems

3.1 Overview of Nonlinear Distortion in OFDM Systems

In the previous chapter, OFDM was shown to be a very effective technique for high speed digital communications in environments affected by high interference and multipath propagation. However, the principle of OFDM that brings this advantages (the use of several sub-carriers resulting in a proportionally longer symbol period in each sub-carrier) brings also one of the major drawbacks of this technique.

As the number of sub-carriers N_c increase, the OFDM signal dynamic becomes larger. This happens because for each sub-carrier there is a signal with random phase and amplitude. In other words, OFDM signals have a large peak-to-average power ratio (PAPR) [7]. This characteristic makes the OFDM system very sensitive to non-linear distortions. The PAPR is defined as:

$$PAPR_{OFDM} = \frac{max|x_i|^2}{\frac{1}{N_c}\sum_{i=0}^{N_c-1}|x_i|^2},\ i = 0, \ldots, N_c - 1 \tag{3.1}$$

where x_i are the transmitted time samples of an OFDM symbol.

Nonlinear distortions are mainly caused by the power amplifiers (PA). The nonlinear distortion at the transmitter causes interference both inside and outside the signal bandwidth. The in-band component affects the system BER [8] while the out-of-band component affects adjacent frequency bands.

The high PAPR in OFDM systems brings an additional condition in order to maintain their performance in an acceptable range. The nonlinear distortions have to be kept as low as possible. A linear behaviour is achieved when operating the amplifier sufficiently below its saturation point. As a result, a high output backoff (OBO) is required in the power amplifiers. The OBO is the ratio between the saturation power and the actual output power, that is the power at which the amplifier is operating. This can be expressed as

$$OBO = \frac{P_{sat}}{P_{out}} = \frac{V_{sat}^2}{E\{|y(t)|^2\}} \tag{3.2}$$

where V_{sat} is the saturation voltage and $y(t)$ is the output of the amplifier.

Generally high power amplifiers (HPA) show better power efficiency when driven close to their saturation point. Therefore, a higher OBO results in a lower power efficiency. Power efficiency is especially valuable in the cases, for example, of portable devices (mobile phones) and satellite systems, where power is in deed a very limited and expensive resource. Therefore a preliminary study of the system is required for finding a good trade-off between transmitted power and degradation [9].

3.1.1 The OFDM RF sub-system

The power amplifier (PA) is part of the RF sub-system of OFDM systems. The RF sub-system consists of the in-phase and quadrature (IQ) modulator, a baseband converter, a spectral filter and the power amplifier. Figure 3.1 shows a block diagram of an RF sub-system.

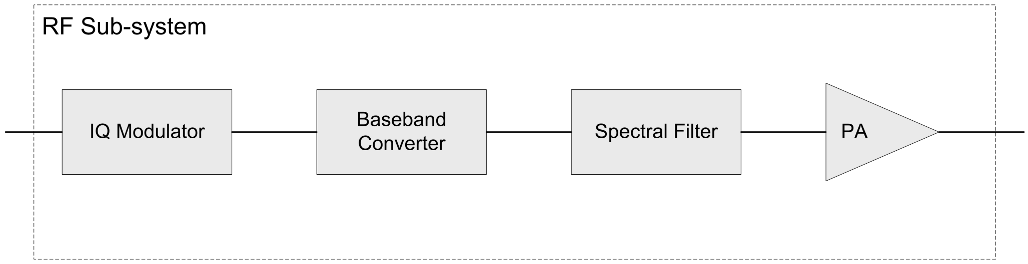

Figure 3.1: Block diagram of the RF sub-system

The IQ modulator splits the input data stream into two separate streams x_I and x_Q. Then it multiplies these streams by a sine and a cosine waveforms respectively. The resulting signals are later summed together to form a modulated intermediate frequency (IF) carrier. The envelope of the IF carrier A is given by:

$$A = \sqrt{x_I^2 + x_Q^2} \tag{3.3}$$

The modulated IF carrier is then converted to the RF carrier frequency via the baseband converter. The baseband converter is built up of a multiplier and a local oscillator of frequency f_c which is the carrier frequency or central frequency. The output of the baseband converter is then filtered by the spectral filter to limit the bandwidth of the RF carrier. This filtered signal is then amplified by the PA to an adequate power level for the RF transmission.

The amplifier introduces most of the nonlinear distortion in the system and therefore it is the main element to be analysed in this chapter. The following section describes the characteristics and models of power amplifiers used in OFDM systems.

3.2 Power Amplifiers in OFDM applications

Ideally, a power amplifier produces a scaled version of the input signal. Therefore, the output an ideal amplifier can be described as

$$V_{out} = GV_{in} \tag{3.4}$$

where G is the voltage gain of the amplifier. The output signal has the same waveform as the input signal and no new frequencies are introduced.

However, real amplifiers do not behave in this way. They act on the signal in an uneven way depending on the magnitude and frequency of the signal. This results in a distortion of the waveform that does not follow the superposition principle

$$\begin{aligned} f(Cx_1) &= C(f(x_1)) \\ f(x_1 + x_2) &= f(x_1) + f(x_2) \end{aligned} \tag{3.5}$$

and therefore is not linear.

Nonlinearities in amplifiers can be described by their amplitude transfer characteristics - also referred as Amplitude Modulation/Amplitude Modulation (AM/AM) conversion - and phase transfer characteristics or Amplitude Modulation/Phase Modulation (AM/PM) conversion. The AM/AM conversion ($F_A[\rho]$) describes the amplitude of the output signal for a given input amplitude (ρ) and the AM/PM conversion ($F_A[\rho]$) describes the phase of the output for a given input amplitude (ρ).

Both AM/AM conversion and AM/PM conversion can either depend on the frequency at which the amplifier is operating or be the same for all frequencies. Therefore, amplifiers can be classified, in terms of nonlinearity, into frequency-dependent and frequency-independent [10]. Frequency independent amplifiers can either be memoryless system or systems with memory. Frequency-dependent amplifiers are systems with memory.

In table 3.1 a summary of the classification and different distortion effects of nonlinear amplifiers is presented.

Frequency-Independent Systems	**Frequency-Dependent Systems**
Memoryless systems Gain distortion	Systems with memory Frequency-dependent gain distortion Frequency-dependent phase distortion
Systems with memory Gain distortion Phase distortion	

Table 3.1: Classification of distortion effects in nonlinear amplifiers

In a memoryless system there are no energy storing components, and any change in the input occurs instantaneously at the output. In other words, the output and the input are in-phase and there is only AM/AM conversion. In a system with memory there are energy storing devices, therefore the output depends on the previous inputs values. This results in phase distortion in addition to the amplitude distortion. Detailed information about nonlinearities with and without memory are presented in [11] and [12].

For the analysis of nonlinear distortion in OFDM systems, amplifiers are often considered frequency-independent. In addition, even considering that in general the overall system is considered to have memory[1], there is no ISI between OFDM symbols as explained in Chapter 2. Furthermore, in [9, 13, 14] the analysis is performed for traveling-wave tube amplifiers (TWTA) and solid-state power amplifiers (SSPA). In addition, [15] uses the soft envelope limiter (SEL). An overview of the models for these amplifiers is presented below. Models for other amplifiers as well as a deeper discussion about amplifier nonlinearities is available in [10].

[1]This, because all practical amplifiers contain elements with a certain level of energy storage, but also because of the filtering processes performed before the amplification.

3.2.1 High Power Amplifiers

Today, the most commonly used devices are the ones based on electron beam tubes, usually Travelling Wave Tube Amplifiers (TWTA), and the Solid State Power Amplifiers (SSPA) as shown in table 3.2, for different microwave bandwidths.

TWTA operation is based on the same principle of the incandescent electric light bulb: making an electric current flow through a glass tube surrounding a vacuum.

An electron emitter, which mainly consists of a heater, a cathode and an anode, starts the stream of electrons by heating the cathode. The stream passes through the anode and travels through the helix. At this point the energy is transferred to the signal travelling around the helix thus amplifying it. Because of their construction, TWTAs offer large currents and therefore, high output powers can be obtained.

TWTAs can be modelled, for the non-frequency selective case, using the model proposed by Saleh in [16] as:

$$F_A[\rho] = \frac{v\rho}{1+\beta_a\rho^2} \tag{3.6}$$

$$F_P[\rho] = \frac{\alpha_p\rho^2}{1+\beta_p\rho^2} \tag{3.7}$$

$F_A[\cdot]$ and $F_P[\cdot]$ are the AM/AM and AM/PM conversion functions respectively, v is the small-signal gain of the amplifier, β_a depends on the input saturation voltage and α_p and β_p depend on the maximum phase displacement that can be introduced by the amplifier. These values are usually chosen to be [17]:

$$v = 1,\ \beta_a = 0.25,\ \alpha_p = \pi/12,\ \beta_p = 0.25 \tag{3.8}$$

The term "solid state" refers to devices whose operation does not depend on wave tubes. SSPAs are made of several Gallium Arsenide (GaAs) metallic semiconductor Field Effect Transistors (FETs). The different output power levels can be achieved by the combination of serial and parallel FET arrays.

A model for the SSPA, presented in [18] describes the SSPAs in terms of their AM/AM and AM/PM conversion.

$$F_A[\rho] = \frac{v\rho}{[1 + (v\rho/V_{sat})^{2r}]^{1/(2r)}} \tag{3.9}$$

$$F_P[\rho] = 0 \tag{3.10}$$

$F_A[\cdot]$ and $F_P[\cdot]$ are the AM/AM and AM/PM conversion functions respectively, v is the small signal gain of the amplifier and r is the smooth factor of the transition between the linear operation and the saturation operation. Note that the AM/PM is zero [2], therefore it can be said that SSPAs only introduce AM/AM distortion.

TWTAs and SSPAs can be used in different frequency bands as presented in table 3.2.

Band name	**Frequency-band GHz**	**Solid-state type**	**Tube type**
X-band	10.6 – 10.7	20 W (GaN device)	3000 W (TWT)
Ka-band	22 – 36.5	6 W (0.15 μm PHEMT devices)	1000 W (Klystron)
Q-band	42.4 – 49.04	4 W	-
W-band	86 – 92	0.5 (TRW)	1000 W (EIKA)

Table 3.2: SSP and TWT amplifier types by frequency band.

There are several differences between these types of amplifiers, and the amplifier to be used depends on the application. Table 3.3 presents a summary of the advantages, disadvantages and applications of TWT and SSP amplifiers.

Other model that can be considered is the Soft Envelope Limiter (SEL) that although it does not exactly model a real amplifier it can be used as a model for pre-distorted

[2]The effect of the AM/PM conversion is not exactly zero, but it is very small and thus it is not considered in the model.

	TWTA	**SSPA**
Advantages	-Wide range of frequencies at the same time. -Excellent performance in audio and satellite devices -Efficient and less expensive for high power outputs (>10kW) and frequencies (>50Mhz)	-Small size -Amplification in stages -Continues to operate after partial failures
Disadvantages	-Difficult to repair -Expensive and complex power supply -Short active life (4 – 6 years)	-High power consumption -Non-stable behaviour due to failures -Not recommended for low frequencies.
Applications	-High power applications (e.g. remote sensing) -Earth stations and communication satellites -Professional audio -High-power UHF TV stations -FM broadcast stations	-Low and medium power applications -AM and FM broadcast transmitters -TC, HF/VHF, lower power UHF, OFDM and HDTV broadcast -Broadband and wireless RF -Mobile phone handsets and base station transmitters.

Table 3.3: Advantages, disadvantages, and applications of SSPAs and TWTAs

signals[3]. The SEL can be modelled as:

$$F_A[\rho] = \begin{cases} \rho & \rho \leq V_{sat} \\ V_{sat} & \rho > V_{sat} \end{cases} \tag{3.11}$$

Now that the models for different amplifiers have been presented, the next step is to characterise the nonlinearities caused by these amplifiers in OFDM systems. The following section provides a description of the main nonlinear effects in OFDM systems and explains how these nonlinearities are characterised and measured.

3.3 Characterisation of Nonlinear Effects in OFDM Systems

In section 3.1, it was shortly explained that the distortions caused by nonlinearities can be classified into two components: the in-band component and the out-of-band component. The last causes mainly adjacent-channel interference (ACI) [9] affecting other systems in adjacent bands but has no effects on the performance of the actual system.

The in-band component, on the other hand, affects the performance of the system. The effects of these distortions are:

- Interference between the in-phase and quadrature (I/Q) components due to AM/PM conversion (if present).
- Intermodulation effects on the sub-carriers.
- Wrapping of the signal constellation in each sub-channel.

Degradation in the performance of a communication system generally results in

[3] Pre-distortion is a technique for mitigating nonlinear distortion described in section 3.5.

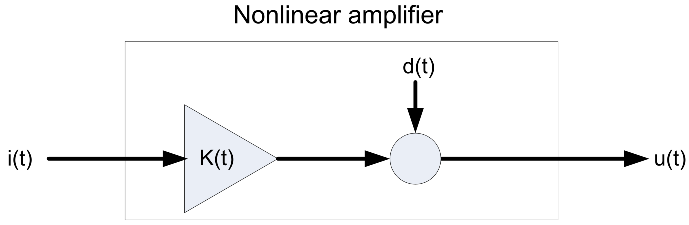

Figure 3.2: Alternative model of a nonlinear amplifier.

an increase in the BER. The following analysis (proposed by Dardari in [14, 15]) explains, quantitatively, the impact of the nonlinearities in the BER.

At the input of the amplifier, the complex envelope of the OFDM signal is defined as:

$$i(t) = \rho(t)e^{j\phi(t)} = \sum_{m=-\infty}^{\infty} \sum_{n=0}^{N_c-1} i_n^m g(t - nT_s - mN_cT_s) \tag{3.12}$$

Where N_c is the number of sub-carriers, T_s is the channel symbol time, $g(t)$ is the pulse shaping used in the digital-to-analog converter and i_n^m represents the channel symbols at the output of the IFFT in the m-th OFDM frame. It is defined as:

$$i_n^m = \frac{1}{N} \sum_{k=0}^{N_c-1} S_k^m e^{j2\pi \frac{nk}{N_c}}, n = 0, \dots, N_c - 1 \tag{3.13}$$

S_n^m are modulation symbols from an M-ary constellation. The analysis presented here assumes M-QAM modulation and rectangular pulse shaping.

The output of the HPA can be obtained by expressing the output in terms of the input signal and the AM/AM and AM/PM conversion functions. For the general case, the output of the HPA is expressed as:

$$u(t) = F_A[\rho(t)]e^{F_P[\rho(t)]}e^{j\phi(t)} = F\big(\rho(t)\big)e^{j\phi(t)} \tag{3.14}$$

The input signal $i(t)$ is built by the sum of N_c independent contributions with the same statistics that are the transmitted symbol in each sub-channel. As a consequence of the central limit theorem, it is possible to state that for a sufficiently large number of sub-carriers N_c, the input signal $i(t)$ can be handled as a complex Gaussian non-stationary process. For this case, as presented in figure 3.2, the output of the HPA can be expressed in terms of a deterministic complex function $K(t)$ and an additive noise $d(t)$:

$$u(t) = K(t)i(t) + d(t) \tag{3.15}$$

$d(t)$ is a zero-mean process not correlated to $i(t)$ and $K(t)$ is defined by the expectation of the output signal. Furthermore, when the waveform is rectangular, $K(t)$ does not longer depend on t and, for a sufficient number of sub-carriers N_c, can be defined as a constant given by [14]:

$$\begin{aligned} K_o &= \frac{1}{2}E\left[F'(\rho) + \frac{F(\rho)}{\rho}\right] \\ &= \frac{1}{2}\int_0^{\infty} \frac{\rho}{P_i} e^{-\rho^2/2Pi}\left[F'(\rho) + \frac{F(\rho)}{\rho}\right] d\rho \end{aligned} \tag{3.16}$$

where the function $\rho/P_i e^{\rho^2/2P_i}$ is the pdf of the random variable ρ, when the power of the OFDM signal at the input of the HPA is P_i.

The output $u(t)$ is sent through the channel, passed through a matched filter and sampled. The phase rotation caused by K_o is compensated at the receiver. After performing the FFT, the received symbol can be expressed as:

$$V_k^m = V_o|K_o|S_k^m + D_k^m + W_k^m \tag{3.17}$$

where W_k^m is the Gaussian noise component (or thermal noise) with variance σ_W^2

and D_k^m is an additive zero-mean component with variance σ_D^2 and does not depend on the useful component of V_k^m. D_k^m is also known as the nonlinear distortion noise (NLD).

The NLD D_k^m is the result of the FFT of the filtered distortion component $d_r(t)$ and can be expressed as:

$$D_k^m = \sum_{n=0}^{N_c-1} d_{rn}^m e^{-j2\pi nk/N} \tag{3.18}$$

Therefore, it can also be assumed that since all $d_i(t)$ can be considered independent and uncorrelated to $i(t)$, the NLD D_k^m is a complex Gaussian random variable with variance σ_D^2. After some manipulation, it is possible to express σ_d^2 as:

$$\begin{aligned} \sigma_d^2 &= E\Big[|d_n^m|^2\Big] \\ &= E\Big[|u_n^m|^2\Big] - |K_o|^2 E\Big[|i_n^m|^2\Big] \\ &= E\Big[|F(\rho)|^2\Big] - |K_o|^2 E\Big[\rho^2\Big] \end{aligned} \tag{3.19}$$

and the variance of the NLD is expressed as:

$$\sigma_D^2 = N_c \sigma_d^2 \tag{3.20}$$

Now we have a signal that is distorted by two independent additive Gaussian noise components. The signal-to-noise ratio can be expressed in terms of these two noise components as:

$$\gamma = \frac{|K_o|^2}{\sigma_W^2 + \sigma_D^2} \tag{3.21}$$

The BER can then be approximated in a conventional way since it is a function of the modulation (with a constellation of size M) and the signal-to-noise ratio.

$$P_b = \frac{2}{\log_2 M} \frac{\sqrt{M}-1}{\sqrt{M}} \operatorname{erfc}\sqrt{\frac{|K_o|^2}{\sigma_W^2 + \sigma_D^2}} \tag{3.22}$$

The process presented above, proposes an analytical method for evaluating the signal-to-noise ratio and BER in case of nonlinear distortion. Based on this process, it is possible to establish a method for evaluating the degradation in performance when nonlinearities are present, compared with the case where there is no nonlinear distortion. This method is described in the following section.

3.4 Performance Evaluation

It is important to have a measure that can quantitatively describe the impact of the nonlinear distortion compared with the case where there are no nonlinearities in the system. A useful performance measure (used in [9, 14]) is the total degradation (TD) as a function of the HPA output backoff:

$$\mathrm{TD}_{\mathrm{dB}} = \mathrm{SNR}_{\mathrm{dB}} - \mathrm{SNR'}_{\mathrm{dB}} + \mathrm{OBO}_{\mathrm{dB}} \tag{3.23}$$

$\mathrm{SNR}_{\mathrm{dB}}$ is the required SNR in decibels at the input of the detector to obtain a fixed BER (e.g. 10^{-4}) for a given value of OBO. $\mathrm{SNR'}_{\mathrm{dB}}$ is the required SNR to obtain the same BER in the absence of nonlinearity.

From equation (3.2), an alternative expression for the OBO in dB can be:

$$\mathrm{OBO}_{\mathrm{dB}} = 10 \log \left(\frac{A_o}{2P_o} \right) \tag{3.24}$$

A_o is the output saturation voltage of the HPA, and P_o is the power of the signal at the output of the HPA. Based on equation (3.15), P_o can be expressed as:

$$P_o = |K_o|^2 P_i + P_d \tag{3.25}$$

After some manipulations, it is possible to express the signal-to-noise ration as:

$$\gamma = \left[\frac{N_o}{E_b} \frac{p_i}{\log_2 M} (1 + \eta) + p_i \theta \right]^{-1} \tag{3.26}$$

where $p_i = 2N_C P_i = 2(M-1)/3$ is the normalised input power and

$$\begin{aligned} \eta &= \frac{P_d}{|K_o|^2 P_i} \\ \theta &= \frac{\sigma_{dr}^2}{2|K_o|^2 P_i} \end{aligned} \tag{3.27}$$

are the nonlinear distortion noise powers before and after the receiver filter respectively. In this sense, η represents the total nonlinear distortion while θ represents the filtered part of it, in other words, the in-band component. If the waveform is rectangular, it can be assumed that all the noise is in-band [14], therefore $\eta = \theta$.

Both K_o and P_o/P_i depend only on the factor $\beta = A_o^2/2P_i$. In addition, the OBO can be rewritten as $OBO = \beta P_i/P_o$. The actual value for K_o and P_o/P_i in terms of β, depends on the model of HPA used. Analytical expressions for SSPA amplifiers and rectangular pulse shape can be found in [14]. Alternatively, it is possible to find the value of θ as a function of β by simulation.

Considering the signal-to-noise ratio required to achieve a predefined BER, for a given constellation size M, as γ_M from (3.22), it is possible to rewrite equation (3.23) by using the following substitutions from (3.26):

$$\text{SNR} = \frac{p_i}{\log_2 M}\gamma_M \frac{1+\eta}{1-p_i\gamma_M\theta} \tag{3.28}$$

$$\text{SNR'} = \frac{p_i}{\log_2 M}\gamma_M \tag{3.29}$$

$$\text{OBO} = \beta\frac{P_i}{P_o} \tag{3.30}$$

resulting in an expression depending only on β

$$\text{TD}_{\text{dB}} = 10\log_{10}\left(\frac{1+\eta(\beta)}{1-p_i\gamma_M\theta(\beta)}\right) + 10\log_{10}\beta\frac{P_i}{P_o} \tag{3.31}$$

From equation (3.31), the minimum TD can be found numerically or by inspection of the curve. It is also possible to obtain an expression analytically. In this case, the minimum asymptotic OBO_{dB} that results in the wanted BER can be obtained from the value of β that satisfies the equation

$$\theta(\beta) = \frac{1}{p_i\gamma_M} \tag{3.32}$$

Note that there is a dependency on the size of the alphabet M since p_i and γ_M depend on M as well.

With this performance measure it is possible to evaluate the effects of nonlinear distortion in the OFDM system. This is very useful since it can help to find a good trade-off between power efficiency and system performance. Yet, this performance can be improved using different techniques. Some of these techniques are presented in the following section.

3.5 Mitigation of Nonlinear Effects in OFDM Systems

There are several techniques to overcome the effects of nonlinear distortion in OFDM systems. This section provides a brief description of some of these techniques.

One approach for reducing the effects on nonlinearities is the use of pre-distortion techniques. These techniques aim to compensate the distortions caused by the amplifier before feeding the signal into it. This is, to modify the signal before the amplification process takes place so that the output of the amplifier is closer to the original signal. These modifications can be performed either in a non-adaptive way or in an adaptive way.

The most common non-adaptive pre-distortion technique is amplitude clipping. Figure 3.3 presents a block diagram for this technique.

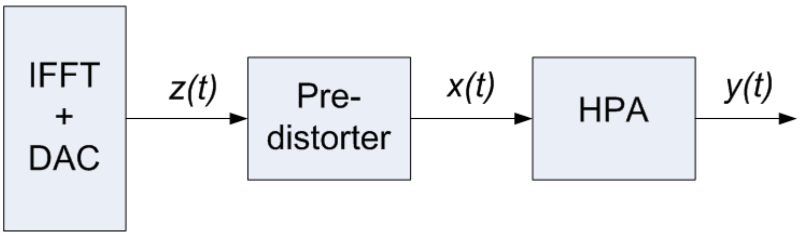

Figure 3.3: Block diagram of non-adaptive pre-distortion.

The input of the pre-distorter $z(t)$ is the signal we want to transmit, in other words, the output of the IFFT module. The pre-distorter modifies the signal and outputs $x(t)$ which is later fed to the HPA resulting in an output signal $y(t)$. The objective of this technique is to have the following relation between the output of the IFFT module $z(t)$ and the output of the amplifier $y(t)$

$$y(t) = \begin{cases} z(t) & \text{if } |z(t)| \leq V_{sat} \\ V_{sat}\frac{z(t)}{|z(t)|} & \text{if } |z(t) > V_{sat}| \end{cases} \tag{3.33}$$

The output of the pre-distorter can be expressed as:

$$x(t) = |x(t)|e^{j\varphi(t)} = r(t)e^{j\varphi(t)} \tag{3.34}$$

where r(t) depends on the HPA amplifier function.

It is possible to see from (3.33) that for all values below the saturation voltage, the pre-distorter should perform the inverse of the HPA operation; as for values exceeding the saturation voltage, it will limit the amplitude to the saturation voltage but keep the phase. Therefore, amplitude clipping adds another source of noise that needs to be considered carefully.

There are techniques that help mitigating the noise caused by amplitude clipping. These techniques are known as Clipping Amplitude Recovery (CAR). A decision-aided method and a Bayesian-inference method for CAR can be found in [2].

Adaptive pre-distortion techniques follow a similar principle but estimate the function of the HPA in an adaptive way. More information about these techniques can be found in [19–21].

Other approach to reduce the effects of nonlinearities in OFDM systems is the use of coding techniques. These techniques aim to reduce the PAPR of the OFDM signal. These methods usually require side information to be known both at the receiver and the transmitter.

One set of coding techniques within this group is the Partial Transmit Sequence techniques. This technique was first introduced by Muller and Hüber in [22]. There are several algorithms for implementing these techniques. However, the principle remains the same. The objective is to reduce the PAPR of the transmitted signal. Figure 3.4 shows the PTS functional block diagram.

Each data block is partitioned into M_s disjoint sets. The weighting factors b_m have to be optimised so that the combination of the clusters, and therefore the PAPR,

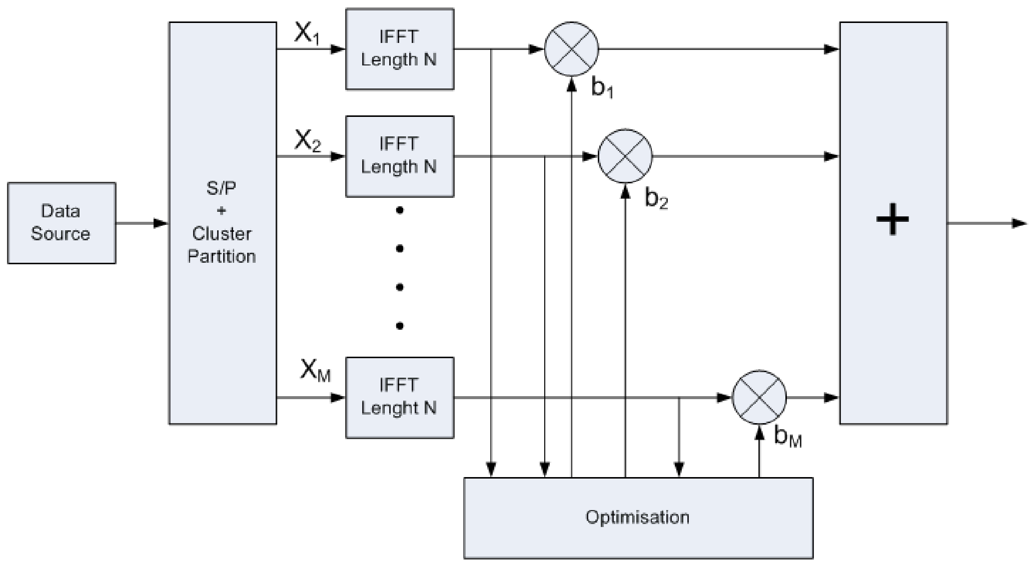

Figure 3.4: PTS functional block diagram.

are minimised:

$$\mathbf{X}' = \sum_{m=1}^{M_s} b_m \mathbf{X}_m \tag{3.35}$$

The weighting factors b_m are pure phase rotations. After the IFFT, the expression above can be written as:

$$\mathbf{x}' = \sum_{m=1}^{M_s} b_m \mathbf{x}_m \tag{3.36}$$

It is then possible to set one rotation factor b_m to one, and the remaining can be found with:

$$\hat{b}_m = \arg\min \left(\max_{b_m} \left\{ b_1 \mathbf{x}_1 + \sum_{m=2}^{M_s} b_m \mathbf{x}_m \right\} \right), \ m = 2, \ldots, M_s \tag{3.37}$$

This technique offers a good improvement in performance with a relatively low increase in complexity of the transmitter.

Another coding technique was introduced by Baümi in [23]. It is called Selective Mapping (SLM). The idea is, that given M_s statistically independent OFDM symbols having the same information, to select the symbol with the lowest PAPR for transmission. These M_s statistically independent OFDM symbols are generated by weighting the OFDM symbols with M_s random sequences of length N_c. This can be performed using Walsh sequences [23] or a random interleaver [24]. This technique, combined with block coding, can provide both error protection and PAPR reduction when only the code-words with small PAPR are selected.

3.5.1 Power Amplifier Nonlinearity Cancellation

Power Amplifier Nonlinearity Cancellation (PANC) is introduced in [25]. This technique attacks the nonlinearity problem from the receiver point of view. It requires the knowledge of the channel coefficients and the characteristics of the amplifier. The main idea of this technique is, based on the received symbols, to estimate the NLD noise term caused by the amplifier and subtract it from the received signal. By repeating this process for several times, a better estimation of the received signal is achieved.

Figure 3.5 presents a block diagram of the PANC. An IFFT is applied to the decoded symbols and then passed through the amplifier model. The input of the amplifier is then subtracted from the output to obtain an estimation of the NLD noise term. The result is then transformed to the frequency domain by the FFT block and subtracted from the received signal. Since the distortion term is subtracted in the frequency domain, this is equivalent to a symbol-by-symbol subtraction.

One important advantage of this technique is that it is applied at the receiver side. In the cases where applying the techniques presented above to a certain transmitter is not practical, the PANC can be applied at the receiver to mitigate the effects of the nonlinear amplifiers with very good results. If we consider a mobile system as an example, the transmitter-based techniques can be applied in the base station for the downlink, while PANC can be applied for the uplink. As a result, the mobile terminal, where resources are more limited, does not need to take any actions related to NLD mitigation.

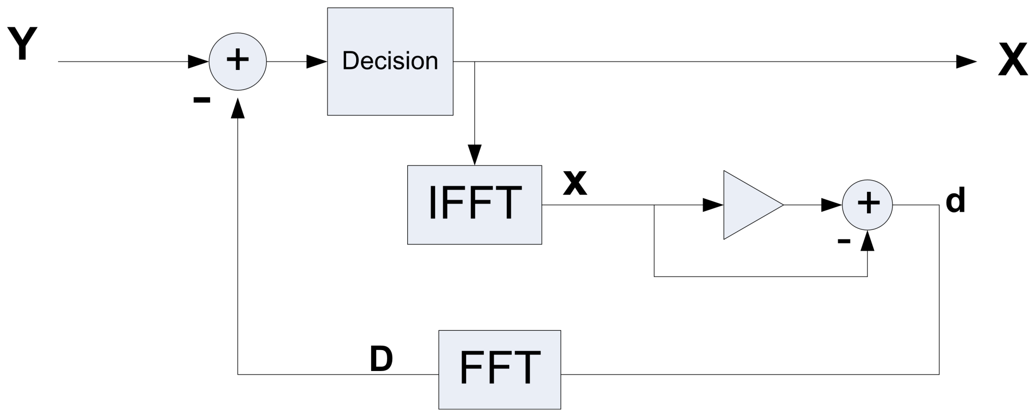

Figure 3.5: Power Amplifier Nonlinearity Cancellation (PANC).

3.6 Summary of the Chapter

In this chapter, nonlinear amplifier distortion in OFDM systems is described. The OFDM RF sub-system is described. The characteristics and models of the most common power amplifiers used in telecommunications are also briefly introduced. A method for characterisation of the NLD effects as well as a method for performance evaluation are explained. Finally, some techniques for mitigating the effects of NLD are briefly presented.

OFDM systems are highly sensitive to NLD because of the high PAPR inherent to this modulation technique. Unfortunately, HPAs are more efficient when operated close to their saturation point, where the nonlinearities increase. The impact on the performance has to be evaluated to find a good trade-off between efficiency and performance. The NLD term can be modelled as an additive Gaussian noise in the system. Several techniques exist for mitigating the effects of NLD noise, specifically PANC technique can be applied at the receiver in cases where other techniques are not available at the transmitter.

Chapter 4

Cooperative Communications

Cooperative communications origins date back to the relay channel model, containing a source, a destination and a relay. The goal of the relay is to facilitate information transfer from the source to the destination. The relay channel was introduced by Van der Meulen [26] and investigated extensively by Cover and El Gamal [27]. The last provided a number of relaying strategies, found achievable regions and provided upper bounds to the capacity of a general relay channel. They also provided an expression for the capacity of the degraded relay channel, in which the communication channel between the source and the relay is physically better than the source-destination link. These studies analysed the relay as a helper for the main channel and only from the capacity point of view. Cooperative communications evolved from this first approach to become a source of diversity. First proposed by Laneman in [28], cooperative communications are currently a topic under intensive research.

4.1 Overview

Diversity techniques have been widely accepted as effective ways to combat multipath fading in wireless communications. Transmitting independent copies of a signal generates diversity that can effectively counteract the effects of fading in the wireless environment. In particular, spatial diversity is generated by transmitting these copies from different locations, allowing independent faded versions of the signal at the receiver.

Multiple-input multiple-output (MIMO) systems are often used as a source of spatial diversity. However, antenna arrays may not be viable in certain cases, for example: handsets or sensor network nodes, where size and hardware limitations complicate the use of multiple antennas in a single device. Cooperative communications offer an alternative in these cases. The main idea is to share the antennas, belonging to different terminals, to form a virtual MIMO system. However, cooperative communication systems differ from MIMO systems mainly on the fact that the channel between the two antennas is a noisy and fading channel. Diversity achieved by the use of cooperative communication is often referred to as cooperative diversity. Figure 4.1 shows a model of a simple cooperative system.

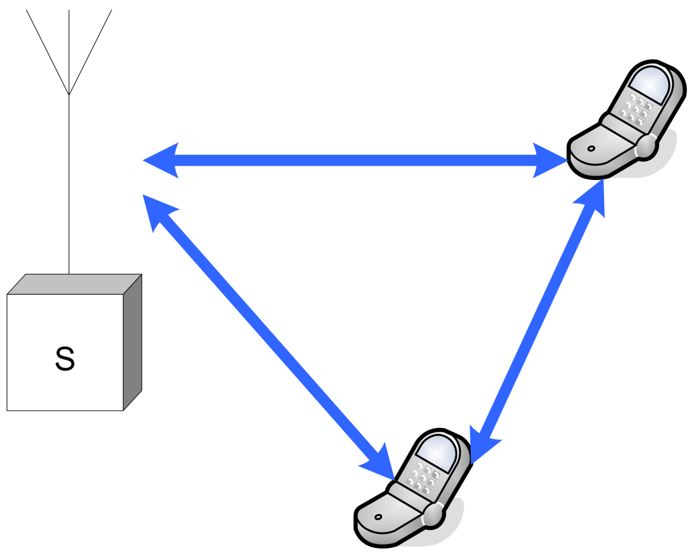

Figure 4.1: Simple cooperative communication system.

Two factors in today concept of cooperative communications, first presented in [29] and further investigated in [30], mark a difference with the idea presented by Van der Meulen. The first is that the destination receives the signal from both the source and the relay. The second is that in a fully cooperative scheme, each mobile device or node works both as a source/destination of its own information and as a relay for other device.

Cooperation can be used both for the uplink and for the downlink. There can be multiple relays, in parallel or in series. The way in which the relay retransmits the signal varies as well. However, a cooperative system can be generally defined as having:

- one source
- one destination
- one or more relays
- three or more independent fading channels

Figure 4.2 show a more general model of a cooperative system.

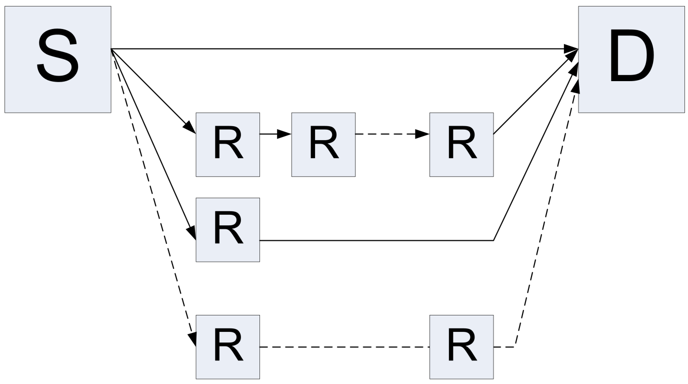

Figure 4.2: General cooperative communication system.

The source sends the information that is listened by the destination and the relay. The relay retransmits the information received by the source towards the destination. At the destination, both signals are combined and decoded.

4.2 Multiple Access

A relay can operate in the half-duplex mode or in the full-duplex mode. A relay operating in the full-duplex mode transmits and receives at the same time and at the same frequency. Moreover, it retransmits the same sample it is receiving. Full-duplex relays have a high complexity since they require separate electronic elements for transmission and reception and even separate antennas with a significant distance between them. In addition the delay between receiving and transmitting must be kept very low. This means, for example, in the case of OFDM that the overall delay (propagation delay, delay spread and relay delay) must be shorter than the length of the cyclic prefix. These relays can be used, for example, in digital television systems where relays can be fully equipped and space and power are not a limitation. Full-duplex relays, however, cannot be used for cooperation.

For the diversity through cooperation to be successful, the destination must be able to receive the original and relayed transmissions separately. This is achieved by sending the two parts orthogonally. This orthogonality can be achieved by separation in time or in frequency, or by using orthogonal spreading codes [31], [30]. In practice, this means that the relay operates in the half-duplex mode. For simplicity, the rest of the chapter focuses on time separation. However, this does not result in a loss in generality. In principle, this means that the cooperation is performed in two time frames:

	Time-frame 1	Time-frame 2
Source	Transmit	Standby
Relay	Receive	Transmit
Destination	Receive (S)	Receive (R)

Table 4.1: Multiple access in cooperative systems.

The following sections cover the main aspects of cooperative communications. Given the complexity of cooperative communications systems, it is important to define some terminology:

	S-R link	S-D link	R-D link
Transmitted signal	x_{SR}	x_{SD}	x_{RD}
Channel	h_{SR}	h_{SD}	h_{RD}
Received signal	y_{SR}	y_{SD}	y_{RD}

Table 4.2: Terminology in the cooperative system.

Figure 4.3 presents a simple system model of the cooperative system that is analysed in these sections.

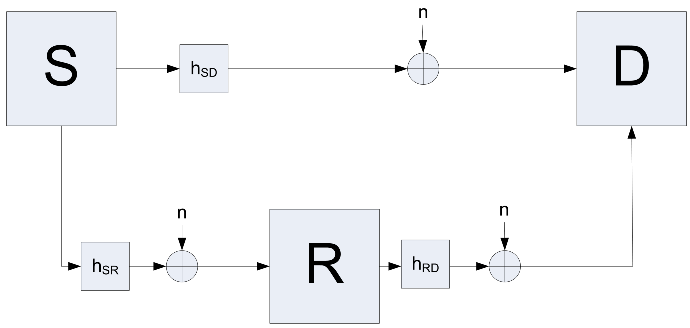

Figure 4.3: System model of a cooperative communication system.

4.3 Cooperation Methods

The way in which the relay processes and retransmits the signal, defines the cooperation method. These methods are [31, 32]:

- Amplify and Forward (AF)
- Decode and Forward (DF)
- Compress and Forward (CF)
- Coded Cooperation (CC)

These methods are described in the following sections.

4.3.1 Amplify and Forward Cooperation

This method was first introduced in [33]. AF relays receive the signal from the source, amplify it, and retransmit it to the destination. The signal received from the source is affected by the channel and the thermal noise. This means, in practice, that what is amplified is not the signal itself but a distorted version of it. Furthermore, the relay amplifies also the noise.

The signal received by the relay is:

$$y_{SR}(k) = h_{SR}(k) * x_{SR}(k) + n_{SR}(k) \tag{4.1}$$

After amplification, the relay transmits the signal:

$$x_{RD}(k) = F\Big(y_{SR}(k)\Big) = F\Big(h_{SR}(k) * x_{SR}(k) + n_{SR}(k)\Big) \tag{4.2}$$

Therefore, the signals received at the destination are:

$$y_{SD} = h_{SD}(k) * x_{SD}(k) + n_{SD}(k) \tag{4.3}$$

$$y_{RD} = h_{RD}(k) * x_{RD}(k) + n_{RD}(k) \tag{4.4}$$

$$= h_{RD}(k) * F\Big(h_{SR}(k) * x_{SR}(k) + n_{SR}(k)\Big) + n_{RD}(k) \tag{4.5}$$

Where y_{SD} and y_{RD} are the signals received from the source (in the time-frame 1) and the relay (in the time-frame 2), respectively.

These signals are then combined. For this method, the MRC is the preferred combiner since it yields to the ML combiner [34]. It is important to note that the noise in the two branches is not equal and therefore the MRC is not the same as in single-hop systems (e.g. MIMO systems). A detailed description on how this is implemented is presented in section 4.5.

Although the noise is amplified at the relay, the destination receives two independently faded versions of the signal and can make better decisions on the detection. In [33] it is shown that for the single relay case, this method achieves diversity of order two.

This cooperation method is rather simple, since there is relatively little processing in the relay node. There are, however, some things that need to be considered when the MRC is used. The interuser channel coefficients (h_{SR}) need to be known, estimated or solved[1]. This method is particularly sensitive to amplifier nonlinearities since they are also transferred to the destination. This issue may become more critical when more than one relay is used. Finally, sampling, amplifying and retransmitting analog values is also challenging [31].

4.3.2 Decode and Forward Cooperation

This method was first presented in [30]. A DF relay first reproduces a copy of the information symbol sent by the source and then retransmits it.

$$x_{RD} = F\left(\hat{x}_{SR}\right) = F\left(dec(y_{SR})\right) \tag{4.6}$$

Where $\hat{x}_{SR} = dec(y_{SR})$ is the decoded version of the received signal at the relay.

This means that the relay must equalise, demodulate, decode and detect the symbol before sending it to the destination. In this case, the retransmitted signal ideally contains the same information than the original signal[2]. In practice, this means that the noise is not amplified and that the effects of the channel are mitigated before sending the signal to the destination. However, if an error occurs in the detection at the relay, this error is forwarded to the destination.

The combination at the destination can be performed using ML combining. The

[1]Even when non-coherent demodulation techniques are used, the effects of h_{SR} are present in the received signal.

[2]Note that although the information bits are equal, the transmitted signal may be different due to e.g. different channel coding

main difference with single-hop systems is that the conditional PDF of y_{RD} is a Gaussian mixture consisting of two components: the correct and the incorrect decision made by the relay. This results in a nonlinear ML combiner [35]. This means that the destination needs to know the error characteristics of the interuser channel. In [33] a hybrid DF system is proposed in which cooperation is only performed if the interuser channel has a high SNR, switching to the non-cooperative scheme whenever this condition is not satisfied.

The main advantages of this method are that noise is not amplified and forwarded to the destination and that the destination does not need to have information about h_{SR}. However, the error characteristics of the channel must be known and the processing load at the relay is significantly higher than in the AF case.

4.3.3 Compress and Forward Cooperation

In Compress and Forward relays (CF), instead of decoding the source messages, the relay helps the destination by forwarding a compressed version of its received signal.

The transmission cycle is divided into two stages. During the first stage, both the relay and the destination listen to the source node. In the second stage, the relay quantises its observations and sends the quantised data to the destination. In general, the correlation between the relay and destination observations can be exploited using Wyner-Ziv coding to reduce the data rate at the relay [32]. During the second stage, new information is also sent by the source to further boost the total throughput.

A practical approach to the CF cooperative scheme is the Slepian-Wolf (SW) cooperation proposed in [36]. Slepian-Wolf cooperation exploits practical Slepian-Wolf codes in wireless user cooperation to help combat inter-user outage.

This type of relays can be useful in the cases where the source-relay link is very noisy since requiring the relay node to decode the message under these circumstances before starting to help the destination, may adversely affect the performance.

4.3.4 Coded Cooperation

Coded Cooperation integrates cooperation into channel coding. In this method, portions of the user's codeword are sent through independent fading paths. The idea behind this is that each user tries to retransmit incremental redundancy to its partner. In the cases where this is not possible, the system switches to a non-cooperative scheme. The big challenge in the CC method (for it to be efficient) is that it should be managed by the code design, to avoid the need of feedback between the users. A detailed description can be found in [37, 38].

The data transmission period for each user is divided into two time segments of N_1 and N_2 bit intervals, respectively. N_1 and N_2 correspond to the size of portions in which the codeword was divided. Each portion is a codeword itself although weaker than the original. For the first interval, each user transmits a codeword consisting of the N_1-bit code partition. Each user also tries to decode the transmission of its partner. If 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, which also contains N_2 bits. Thus, each user always transmits a total of $N = N_1 + N_2$ bits per source block over the two frames. The level of cooperation is defined as N_2/N that is the percentage of the transmitted bits corresponding to the partner.

This method has the advantage that the amount of information sent remains the same, compared with a non-cooperative scheme. Furthermore, by trying to decode the partner's codeword, it has also an embedded method for measuring the quality of the interuser channel to evaluate if it is worth cooperating. One disadvantage is that a complex protocol is needed to define whether there was cooperation or not. Other is that the processing load on the relay is higher compared with the AF case.

4.4 Modulation, Demodulation and Combining Methods

Modulation techniques in cooperative systems are by nature *distributed* since the signal is first modulated at the source and subsequently at the relay. Modulation and demodulation in cooperative systems can be coherent or non-coherent (differential). Coherent techiques require the knowledge of the channel coefficients (achieved by estimation or other means). Differential techniques do not require the knowledge of the channel. Which technique should be used depends greatly on the application. In slow-fading scenarios, coherent demodulation may be the choice, but in fast-fading environments, using coherent demodulation reduces the effective transmission rate since pilot sequences need to be inserted more often. Furthermore, in [39] non-coherent modulation and demodulation are considered to be robust methods to implement control signalling in wireless networks. In [35] distributed modulation techniques are described for AF and DF cooperative methods.

Together with the choice on which modulation and demodulation technique to use, it comes also the combining method to apply. Which combining method to use depends greatly on how much information on the channel is available (either by estimation or by sufficient statistics) and the combination method. In the literature is quite common, for example, the use of MRC for AF methods and ML combiner for DF methods. Specifically in [39] a piecewise-linear (PL) combiner is also proposed.

Combination in cooperative communications is a major challenge. If MRC is considered, the received signals from the different branches have different SNR. This is caused by the two sources of noise present at the source-relay-destination path compared to only one source at the source-destination path. If ML combiner is considered, the received signals from the different braches have a different conditional probability. This is because the processes at the relay are also source of errors that should be considered by the combiner. In both cases the end idea is the same: you cannot treat both branches in the same way because they were affected differently in the process. In [35] MRC and ML combiners are presented for AF and DF methods using both coherent and differential modulation.

4.5 Coherent Amplify and Forward Cooperation

Studying all the variations of cooperating methods, modulation and combination possibilities, is out of the scope of this work. As explained before, AF methods are more sensitive to amplifier nonlinearities which are the main interest of this thesis. In addition, to keep the focus on the amplifier nonlinearities, an approach using coherent demodulation (assuming the knowledge of the channel state) is used. In this section we analyse a cooperative system using coherent demodulation, AF method and MRC combiner. For clarity, the time k is removed from the equations below but these refer to instantaneous time-domain values.

The source transmits the information that is available to both the destination and the relay:

$$x = x_{SD} = x_{SR} \tag{4.7}$$

At the relay, the received signal is:

$$y_{SR} = h_{SR} * x_{SR} + n_{SR} \tag{4.8}$$

The signal is then amplified at the relay:

$$\begin{aligned} x_{RD} &= G y_{SR} && (4.9)\\ &= G\Big(h_{SR} * x_{SR} + n_{SR}\Big) && (4.10)\\ &= G\Big(h_{SR} * x_{SR}\Big) + G n_{SR} && (4.11) \end{aligned}$$

Where $G(\cdot)$ is the gain of the amplifier, which is considered ideal for this analysis.

That is:

$$F(\rho) = G\rho \tag{4.12}$$

In chapter 5, however, the simulations are performed using a model of an SSPA.

As shown in 4.9 the noise is also amplified. The original signal is affected by the channel before being amplified. At this point there is one constraint that has to be considered. The power at the relay is limited, and it is not possible to know if after the channel distortion and amplification, this power constraint is fulfilled.

The amplification process is limited by the available power at the relay. An AF relay can have fixed gain or variable gain. In a fixed gain relay, the amplification is constant but the transmitted power P_S from the source to the relay is controlled to meet the power constraint.

$$G = \frac{1}{\sqrt{P_S \sigma_{h_{SR}}^2 + \sigma_{SR}^2}} \tag{4.13}$$

In a variable gain relay, the gain of the amplifier varies depending on the channel and the noise power in order to keep the output power on the limit. In this case, the gain G varies so that the output of the amplifier has the same power than the original signal.

$$G = \frac{1}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} \tag{4.14}$$

Thus, the signal transmitted by the relay can be expressed as:

$$x_{RD} = \frac{h_{SR} * x_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} + \frac{n_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} \tag{4.15}$$

At the destination, there are two versions of the signal. On one branch the signal received from the source is:

$$y_{SD} = h_{SD} * x_{SD} + n_{SD} \tag{4.16}$$

On the other branch we have the signal received from the relay:

$$\begin{aligned} y_{RD} &= h_{RD} * x_{RD} + n_{RD} && (4.17)\\ &= h_{RD} * \left[\frac{h_{SR} * x_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} + \frac{n_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} \right] + n_{RD} && (4.18) \end{aligned}$$

It is to be noticed that y_{RD} has two noise components. One that corresponds to the noise from the source-relay path, which is first amplified and then distorted by h_{RD}. The other component is simply the noise from the relay-destination path. If we define:

$$\begin{aligned} n_{SRD} &= \frac{h_{RD} * n_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} + n_{RD} && (4.19)\\ \sigma_{SRD}^2 &= \frac{|h_{RD}|^2 \sigma_{SR}^2}{|h_{SR}|^2 + \sigma_{SR}^2} + \sigma_{RD}^2 && (4.20)\\ h_{SRD} &= \frac{h_{RD} h_{SR}}{\sqrt{|h_{SR}|^2 + \sigma_{SR}^2}} && (4.21) \end{aligned}$$

y_{RD} can be expressed as:

$$y_{RD} = h_{SRD} * x_{SR} + n_{SRD} \tag{4.22}$$

From (4.7), $x_{SR} = x_{SD} = x$, since the information sent by the source is the same for the two branches.

To optimally combine the signals, we can apply the Maximum Likelihood principle by maximising the probability that the combined signal is indeed the sent signal.

The received signals can be expressed as:

$$\mathbf{y} = \mathbf{h}x + \mathbf{n} \tag{4.23}$$

where:

$$\mathbf{y} = \begin{bmatrix} y_{SD} \; y_{RD} \end{bmatrix}^T \tag{4.24}$$

$$\mathbf{h} = \begin{bmatrix} h_{SD} \; h_{SRD} \end{bmatrix}^T \tag{4.25}$$

$$\mathbf{n} = \begin{bmatrix} n_{SD} \; n_{SRD} \end{bmatrix}^T \tag{4.26}$$

$$\tag{4.27}$$

The likelihood function can be expressed then as:

$$p(y;x) = \frac{1}{(2\pi \det \mathbf{C})^{1/2}} e^{-\frac{1}{2}(\mathbf{y}-\mathbf{h}x)^H \mathbf{C}^{-1}(\mathbf{y}-\mathbf{h}x)} \tag{4.28}$$

where

$$\mathbf{C} = \begin{bmatrix} \sigma_{SD}^2 & 0 \\ 0 & \sigma_{SRD}^2 \end{bmatrix} \tag{4.29}$$

The objective is to find an x such that:

$$x = \arg\left\{ \min_x \left[(\mathbf{y} - \mathbf{h}x)^H \mathbf{C}^{-1} (\mathbf{y} - \mathbf{h}x) \right] \right\} \tag{4.30}$$

After some operations, x yields:

$$x = (\mathbf{h}^H \mathbf{C}^{-1} \mathbf{h})^{-1} \mathbf{h}^H \mathbf{C}^{-1} \mathbf{y} \tag{4.31}$$

The term $(\mathbf{h}^H \mathbf{C}^{-1} \mathbf{h})^{-1}$ may be considered as a scaling term, resulting in:

$$x = \frac{h_{SD}^*}{\sigma_{SD}^2} y_{SD} + \frac{h_{SRD}^*}{\sigma_{SRD}^2} y_{RD} \tag{4.32}$$

This result is equivalent to the MRC for branches with different SNR. The output of the MRC is:

$$y = \frac{h_{SD}^*}{\sigma_{SD}^2} y_{SD} + \frac{h_{SRD}^*}{\sigma_{SRD}^2} y_{RD} \tag{4.33}$$

With SNR

$$\gamma = \gamma_{SD} + \gamma_{SRD} \tag{4.34}$$

Where

$$\gamma_{SRD} = \frac{\gamma_{SR}\gamma_{RD}}{\gamma_{SR} + \gamma_{RD} + 1} \tag{4.35}$$

In the next chapter, this cooperative system is combined with OFDM modulation and a nonlinear amplifier at the relay. However, before starting this analysis, it is important to consider what cooperation means from the practical point of view and why it is worth analysing it.

4.6 Advantages of Cooperative Communications

Cooperative communications are being studied as an alternative for generating virtual MIMO systems in cases where multiple antennas are not practical to implement in a single device, given their size or battery limitations. Providing the receiver of two uncorrelated faded versions of the signal increases the chances of the receiver to make a correct decision on the information received. As a result of spatial diversity, two important benefits can be identified by using cooperative communications:

- Power efficiency
- Spectral efficiency

In the case of power efficiency it may be argued that more power is needed because each user is transmitting for more than itself. However, the baseline transmit power from all users may be reduced because of the diversity i.e. lower SNR levels are required to achieve the same BER. Therefore it is expected that the net transmit power is reduced.

When it comes to the rate of the system, a counter argument can be that each user transmits its own bits plus the partner's bits, which would decrease the rate of the system. However, due to diversity, the channel code rates can be increased, increasing the spectral efficiency of each user. Even in the cases where the different paths are correlated, the combination of correlated links provide also extra coding gains due to repetition of information [39].

Cooperative communications also offer the possibility of using different network topologies, providing higher flexibility. In [40] some of these topologies, e.g. cooperative gateways and randomised cooperation, are presented. The potential and challenges from the network topology point of view are also described.

In addition to the benefits resulting from cooperative diversity, there are also benefits related to the use of the relay itself. Higher coverage areas can be achieved by the efficient use of terminals that work both as user and as relays. These can translate in solutions for serving hotspots or providing better coverage in corporate buildings for example.

At the network level, studies have shown that diversity provided by MIMO space-time codes can improve performance at the medium access control (MAC), network and transport layers. This improvement in performance is also valid for cooperative systems [31].

The question whether these benefits are worth the incurred cost has been positively answered by several studies and demonstrated in [31].

4.7 Main Challenges in Cooperative Communications

Although the use of cooperative communications as a source of diversity provides certain benefits, there are practical issues that must be considered.

It was previously explained here that there should be orthogonality between the two signals received by the destination. This is not an issue when the very simple case is analysed. However, considering that each relay is also a destination itself, meaning that it is also a user, this may result quite complex to implement. In cellular systems, for example, there is one frequency band for the uplink and one for the downlink. From the handset pioint of view, using cooperation means that it must also be able to transmit in the downlink band and receive in the uplink band. This adds additional hardware requirements to the mobile devices since both transmission and reception are now performed in both bands. This is not such an issue in ad hoc wireless networks or time-division-duplexing (TDD) systems, for example.

Another issue to be considered is transmission and reception requirements in the terminals. In CDMA, for example, mobiles may require to transmit and receive at the same time. In this cases, the transmit signal power may be more than 100 dB over the received signal power. This is beyond the isolation achieved by existing directional couplers. This issue requires additional network considerations such as frequency separation or even time separation.

It is also important to consider that the destination needs certain information about the interuser channel. How to make the destination aware of this information is also an issue. In DF methods, for example, the destination needs to know the error probability of the interuser channel in order to be able to perform an optimal detection. In AF methods, the destination must be able to estimate the interuser channel characteristics. In hybrid DF and CC, even when the information about the interuser channel is not needed, the destination still has to know wether the users cooperated or not, so it can know whose bits are being received during the second frame.

Another issue to be considered is the development of power control mechanisms for cooperative networks. Additional mechanisms may even improve the performance of cooperative systems, by assigning power based on the interuser channel conditions. Power allocation for a fixed-gain relay is analysed in [41].

Another big challenge is to control how users cooperate, meaning which users cooperate with each other. This problem has different approaches depending on the application. In applications where the control is centralised (e.g. cellular networks), techniques to assign partners based on the optimisation of a certain parameter (e.g. average block error rate) have to be developed. In applications with no centralised control (e.g. ad hoc networks) a distributive cooperation protocol has to be implemented, where each user can independently decide which other node to cooperate with. The possible solutions to this issue, must provide a fair treatment to all users without requiring significant additional system requirements. Furthermore, these solutions must be possible to implement together with the existing systems' multiple access protocols. Some initial work related to this issue is presented in [42].

4.8 Summary of the Chapter

This chapter describes the most important aspects of cooperative communications. Cooperative communications are being studied as a source of diversity in cases where having multiple antennas on a single device is not practical. The factors that define how a cooperative system works are the cooperation method, the modulation and demodulation method and the combining method. The AF and DF methods have been widely studied. Since the objective of the thesis is the analysis of nonlinear distortion, the AF method is chosen to be analysed in detail. A coherent AF cooperative system with a single relay and MRC is analysed in detail. Expressions for the received signal and the equivalent SNR are also presented. Finally, the main advantages, as well as the practical issues related to cooperative communications are introduced.

So far in this work, OFDM systems, amplifier distortion in OFDM systems and cooperative communications have been introduced. These three chapters are the theoretical backbone of the thesis. The following chapters cover the case when these three topics are combined. The idea: to implement a cooperative OFDM system, with an AF relay that contains a nonlinear amplifier and evaluate how the nonlinearities affect the performance of this system.

Chapter 5

Nonlinear distortion in OFDM cooperative systems

OFDM, as presented in chapter 2, is a very useful technique with numerous current and potential applications in communications networks. In a similar way, cooperative communications schemes, presented in chapter 4 offer an interesting alternative for using other mobile receivers as a source for diversity. Combining these two techniques would result in a powerful system with a reliable modulation against multipath environments and optimised resource coverage and resource usage. However, as presented in chapter 3, NLD in OFDM systems has to be carefully considered.

The techniques for mitigating amplifier distortion in traditional OFDM systems can be applied in cooperative schemes as well but just solve the problem partially. The mitigation of NLD becomes particularly challenging in the relay element of the cooperative system. If we consider that, in most of the cases, the relay element is a mobile device, the processing and power resources are considerably scarcer than in the source element or base station. As a consequence, the techniques presented in chapter 3 can be difficult to implement in relays, where processing and power resources are limited by the size and battery life of the device.

Therefore, it is important to evaluate how does the NLD affect the performance in cooperative communications when it is not compensated in the relay. In this chapter a method for evaluating this performance is presented. This method is based on simulations.

5.1 The System Model

The cooperative system to be considered consists of one source (S), one relay (R), and one destination (D); all of them with single antennas. Figure 5.1 shows the system model for the cooperative scheme.

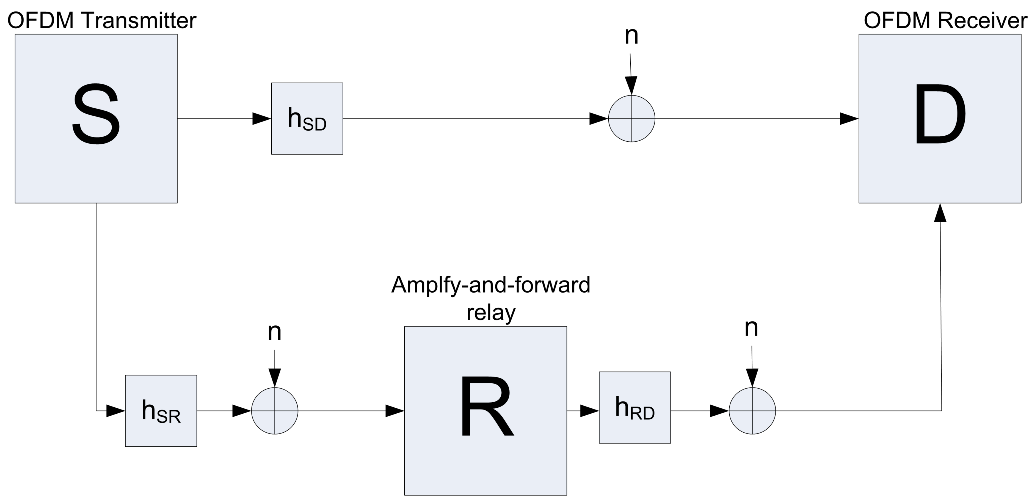

Figure 5.1: System model of the cooperative communication system.

The three channels involved in this systems (h_{SD}, h_{SR}, h_{RD}) are assumed to be Rayleigh fading channels. Each link is assumed to be distorted by AWGN.

S is the transmission point, or the base station (BS). It uses OFDM for transmission. For simplicity, no channel-coding, interleaving, or mapping is considered. In addition, the amplification is assumed to be linear and with unit gain. This assumption can be justified by considering that the BS has enough processing and power resources to apply some of the techniques considered in chapter 3. Figure 5.2 shows the block diagram of S.

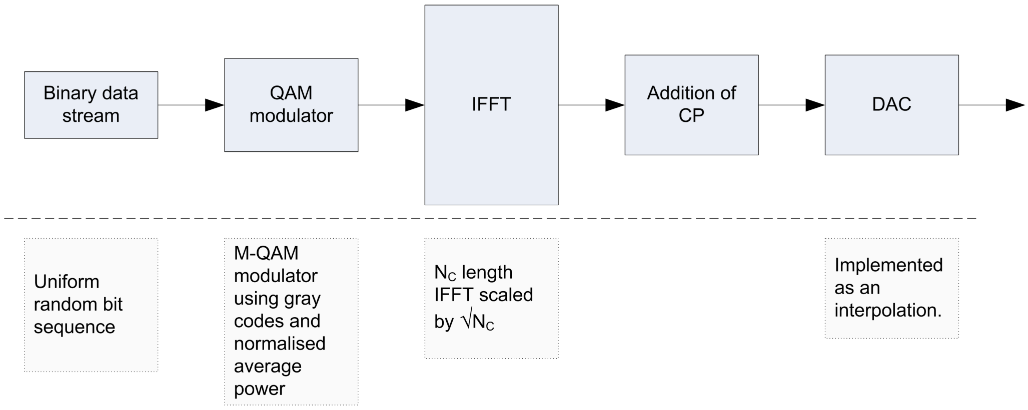

Figure 5.2: Block diagram of the source element.

The bit sequence to be transmitted is randomly generated with uniform distribution. The bit sequence is then modulated into a M-QAM constellation. After the symbols are generated, an IFFT is applied to generate the OFDM symbol to be transmitted.

R is the relay, or relay station (RS). It is assumed to work as an amplify-and-forward (AF) relay. Figure 5.3 shows the block diagram of R.

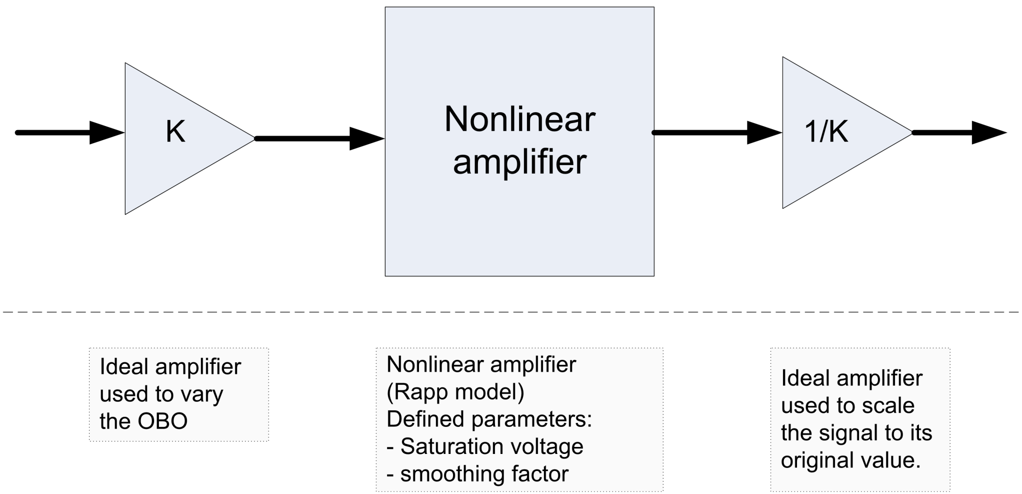

Figure 5.3: Block diagram of the source element.

R only consists of an amplifier which is considered to be an SSPA amplifier. From

(3.9) considering the small signal gain v equal to one[1] the amplifier can be modelled as:

$$F_A[\rho] = \frac{\rho}{[1 + (\rho/V_{sat})^{2r}]^{1/(2r)}} \tag{5.1}$$

$$F_P[\rho] = 0 \tag{5.2}$$

In order to simulate the variations in the output backoff (OBO), the input signal at the relay is multiplied by a factor K and then divided by the same factor at the output of the amplifier.

Note that in R, the power constraint principle presented in Chapter 4 is not considered in this case. This is because to apply this principle R would need to estimate the channel coefficients. This adds computational load that reduces the advantages of an AF cooperation mode. In addition, the output power of the amplifier is already limited by the saturation voltage in the model. Therefore a fixed-gain amplifier is used.

D is the destination or mobile station (MS). The signal is received, combined, and decoded. Figure 5.4 shows a block diagram of D.

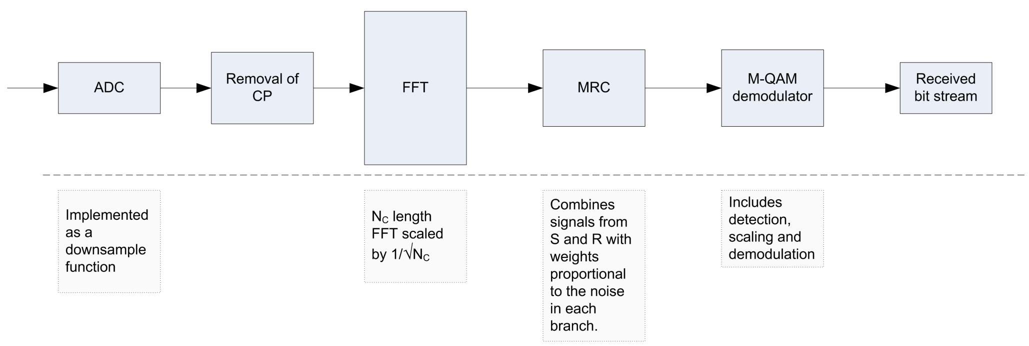

Figure 5.4: Block diagram of the destination element.

[1]This means in other words that the input saturation voltage and the output saturation voltage have the same value.

The FFT decodes the OFDM symbol. The signal received from S and from R are assumed to be orthogonal; for example, S and R transmit on different time instants.

The signals received from the two paths are:

$$Y_{SD} = H_{SR}X + N_{SD} \tag{5.3}$$

$$Y_{SRD} = H_{SRD}X + N_{SRD} \tag{5.4}$$

where N_{SD} is the at the source-destination link and H_{SD} is the Fourier transform of the channel impulse response at the source-destination link. To perform the combination, the amplifier is assumed to be linear with a fixed gain G. Therefore, $H_{SRD} = H_{RD}GH_{SR}$ is the equivalent channel frequency response for the source-relay-destination link and N_{SRD} is the noise sources on this link with variance:

$$\sigma_{SRD} = |H_{RD}|^2 G^2 \sigma_{SR}^2 + \sigma_{RD}^2 \tag{5.5}$$

Under these circumstances, an MRC is the best choice for exploiting diversity [35]. The MRC for equivalent power branches would be:

$$Y_D = H_{SD}^* Y_{SD} + H_{SRD}^* Y_{SRD} \tag{5.6}$$

It is possible to use this approach if the noise is normalised. By defining:

$$\tilde{Y}_{SD} = \frac{Y_{SD}}{\sigma_{SD}} = \frac{H_{SR}}{\sigma_{SD}}X + \frac{N_{SD}}{\sigma_{SD}} \tag{5.7}$$

$$\tilde{Y}_{SRD} = \frac{Y_{SRD}}{\sigma_{SRD}} = \frac{H_{SRD}}{\sigma_{SRD}}X + \frac{N_{SRD}}{\sigma_{SRD}} \tag{5.8}$$

both branches have a noise term with variance 1. Substituting these in (5.6), results in:

$$Y_D = \left(\frac{H_{SD}}{\sigma_{SD}}\right)^* \tilde{Y}_{SD} + \left(\frac{H_{SRD}}{\sigma_{SRD}}\right)^* \tilde{Y}_{SRD} \tag{5.9}$$

In terms of the received signal, the combined signal is:

$$Y_D = \frac{H_{SD}^*}{\sigma_{SD}^2} Y_{SD} + \frac{H_{SRD}^*}{\sigma_{SRD}^2} Y_{SRD} \tag{5.10}$$

which is consistent to the MRC presented in (4.33).

After the signal is combined, it is normalised by a factor:

$$\frac{|H_{SD}|^2}{\sigma_{SD}^2} + \frac{|H_{SRD}|^2}{\sigma_{SRD}^2} \tag{5.11}$$

to keep the amplitude levels used in the transmission. This scaling factor can be compared to the one found in (4.31).

The signal is then passed through the detector and the M-QAM demodulator.

It is important to notice that the system model assumes unit average power of the constellation and that the IFFT and FFT are scaled for giving unitary power in the transmission.

5.2 Structure of the Simulations

Based on the system model presented in the previous section, the objective is to establish the structure of the simulation in a modular way, where changing any parameter of the simulation can be performed in a simple way. For achieving this, the following modules are defined:

- M-QAM modulator
- OFDM modulator
- Rayleigh channel generator
- OFDM demodulator

- Maximum Ratio Combiner
- M-QAM demodulator

These modules are implemented as Matlab functions. The decision of implementing these functions is that they are the tasks that are used constantly in every simulation. On the other hand, having these task implemented as functions gives more flexibility when combining or improving the code.

The M-QAM modulator receives the input signal, the size of the constellation M and the number of carriers N_c used. It frames the data, modulates the signal with gray coding, and scales the constellation so the average power is equal to one.

The OFDM modulator receives the M-QAM symbols, the length of the cyclic prefix, the number of carriers N_c and the up-sampling rate to be used. It applies the IFFT and scales it by a factor $\sqrt{N_c}$ for keeping the unitary reference in the frequency domain. Then it inserts the cyclic prefix and interpolates the signal.

The Rayleigh channel generator, returns the impulse response of a SISO Rayleigh channel based on the following input parameters:

- delay profile
- power profile in dB
- channel impulse response length
- terminal velocity (km/h)
- carrier frequency (MHz)
- bandwidth in Hz
- sub-symbol period in μs

The OFDM demodulator receives the signal, down-samples it to the given rate, then removes the cyclic prefix and applies the FFT scaled by $1/\sqrt{N_c}$ for keeping the time domain signal around unitary average power.

The Maximum Ratio Combiner performs the operation described in (5.10)

The M-QAM demodulator, detects the input signal and returns the corresponding bit sequence.

Using these modules, the complete cooperative system can be implemented in the following way:

<table>
<tr><td colspan="2">- Generate the binary data (uniform random sequence)
- M-QAM Modulation
- OFDM Modulation
- Generate the 3 channels</td></tr>
<tr><td>- Filter the signal with channel h_{SR}
- Add AWGN
- Amplify using the SSPA model
- Filter the output with the channel h_{RD}
- Add AWGN
- OFDM demodulation</td><td>- Filter the signal with channel h_{SD}
- Add AWGN
- OFDM demodulation</td></tr>
<tr><td colspan="2">- Maximum ratio combining
- M-QAM demodulation</td></tr>
</table>

Table 5.1: The complete cooperative system algorithm

Based on this algorithm, the following simulations are implemented:

1. NLD effects on the OFDM symbol spectrum.

2. NLD effects on the BER of the cooperative system.

3. Cooperative system BER with and without NLD under the following conditions:

- Ideal channel with no noise
- Ideal channel with AWGN
- Rayleigh channel with no noise
- Rayleigh channel with AWGN

4. OBO effects on the NLD and the BER of the cooperative system.

In the following sections these simulations are described in detail and analysed.

5.3 NLD Effects on the OFDM Symbol Spectrum

The objective of this simulation is to analyse how the NLD affects the spectrum of the OFDM symbol. This is presented in figure 5.5.

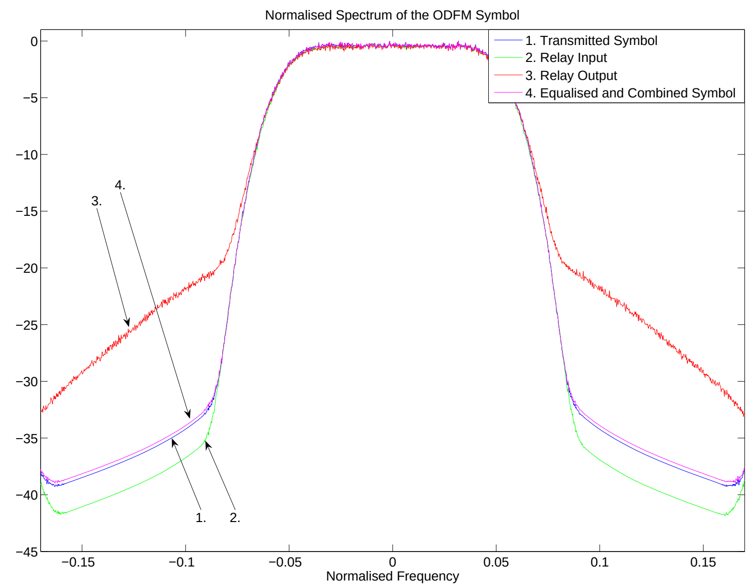

Figure 5.5: Different spectra of the OFDM symbol

It is possible to see from the figure that the OFDM symbol spectrum is affected by an out-of-band distortion that is reflected in the "shoulders" that appear in the figure. At the receiver, the signal is filtered and the out-of-band distortion is not visible any more.

This results show that the spectrum of the symbol is affected by the NLD. However, it is not possible to notice any impact in the performance from these results. In order to analyse the performance of the system, a performance measurement is needed. The BER is a chosen as the measure of performance.

5.4 NLD Effects on the BER of the Cooperative System

The objective of this simulation is to evaluate the performance of the cooperative system affected by the NLD using the same cooperative system without NLD and a non-cooperative (conventional one-to-one) system as references. This is done by evaluating the BER for several values of SNR at the transmitter and different cases of saturation voltage. The transmitted SNR is calculated as the SNR of the S-D link. The results are presented in figure 5.6.

The BER increases significantly even for large values of saturation voltage compared to the transmitted voltage. A large value of saturation voltage can be interpreted as a small NLD. Given that the transmission power is normalised, a $V_s = 1.4V$ should be sufficient for an acceptable performance; somehow, this is not the case. Therefore, there is a new factor in the cooperative systems causing big distortions even for large values of saturation voltage. On the other hand, the BER keeps increasing as the saturation voltage decreases. This is expected since the NLD increases causing additional errors in the reception.

The cause of such a significant increase in BER even for large values of saturation voltage cannot be explained with these results. Therefore, there is the need to use different scenarios to try to isolate the cause of such distortion.

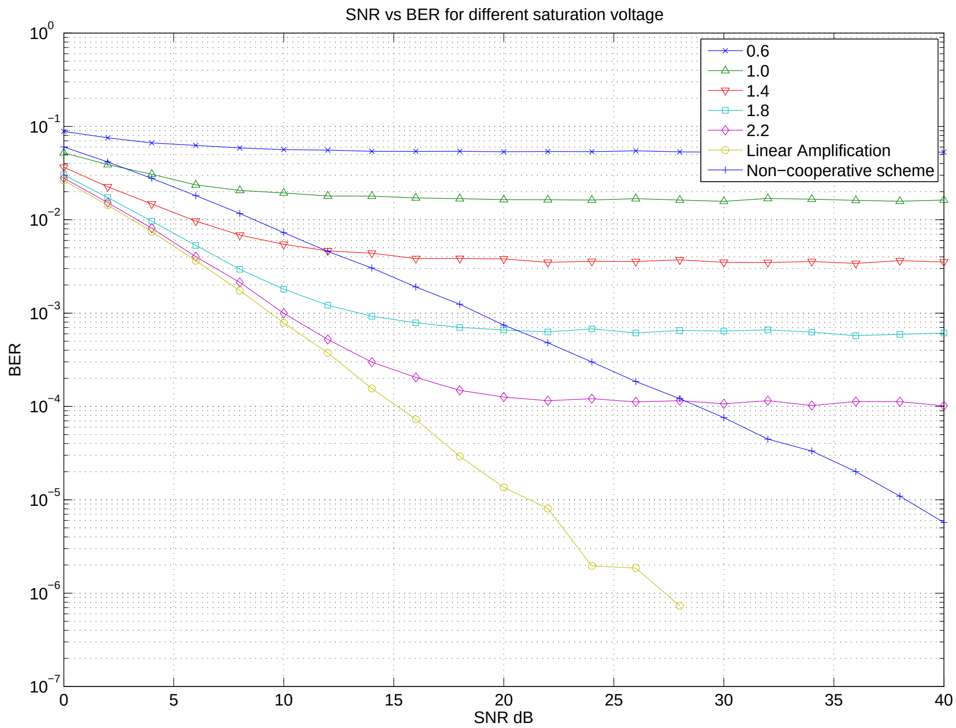

Figure 5.6: BER for different levels of saturation voltage

5.5 Cooperative System BER with and without NLD under Different Scenarios

In order to isolate the cause of the distortion, several simulations - equivalent to the one presented in the previous section - are performed with different scenarios but only for one high value of saturation voltage ($V_s = 1.4V$). The different scenarios only apply to the S-R branch of the system. The rest of the model is not affected. These modifications were:

- Ideal channel with no noise
- Ideal channel with AWGN

- Rayleigh channel with no noise
- Rayleigh channel with AWGN

The results of these simulations are summarised in figure 5.7.

As it can be seen from graphs 5.7.a and 5.7.b, in the case of the ideal channel the BER in both cases keeps close and almost parallel. This means that the insertion of AWGN before the amplifier is not the reason for the increase in the BER found in the previous section.

When analysing 5.7.c and 5.7.d, the inclusion of a Rayleigh channel causes a significant increase in the BER. Therefore it can be stated that the distortion introduced by the channel causes that the signal to be amplified falls more often into the saturation levels (i.e. the nonlinear operation point). The NLD increases as the signal voltage reaches the saturation voltage, where it is known that the amplifier's nonlinear behaviour is stronger. This would explain the increase in NLD when the channel is introduced. In chapter 6 a deeper analysis of the behaviour of the signal in the relay is presented.

5.6 OBO Effects on the NLD and the BER of the Cooperative System

In the previous section it was shown that the Rayleigh fading channel at the source-relay path causes higher nonlinear distortion compared with an ideal unitary channel. A possibility for reducing this distortion is to modify the OBO on the amplifier at the relay. Figure 5.8 shows how the BER changes when modifying the OBO for a fixed saturation voltage of 1.4 V.

It is clear that when the value of OBO increases, the BER decreases. This is expected. The OBO is the ratio between the saturation power and the output power. This reduces the NLD because the amplitude levels of the signal are further

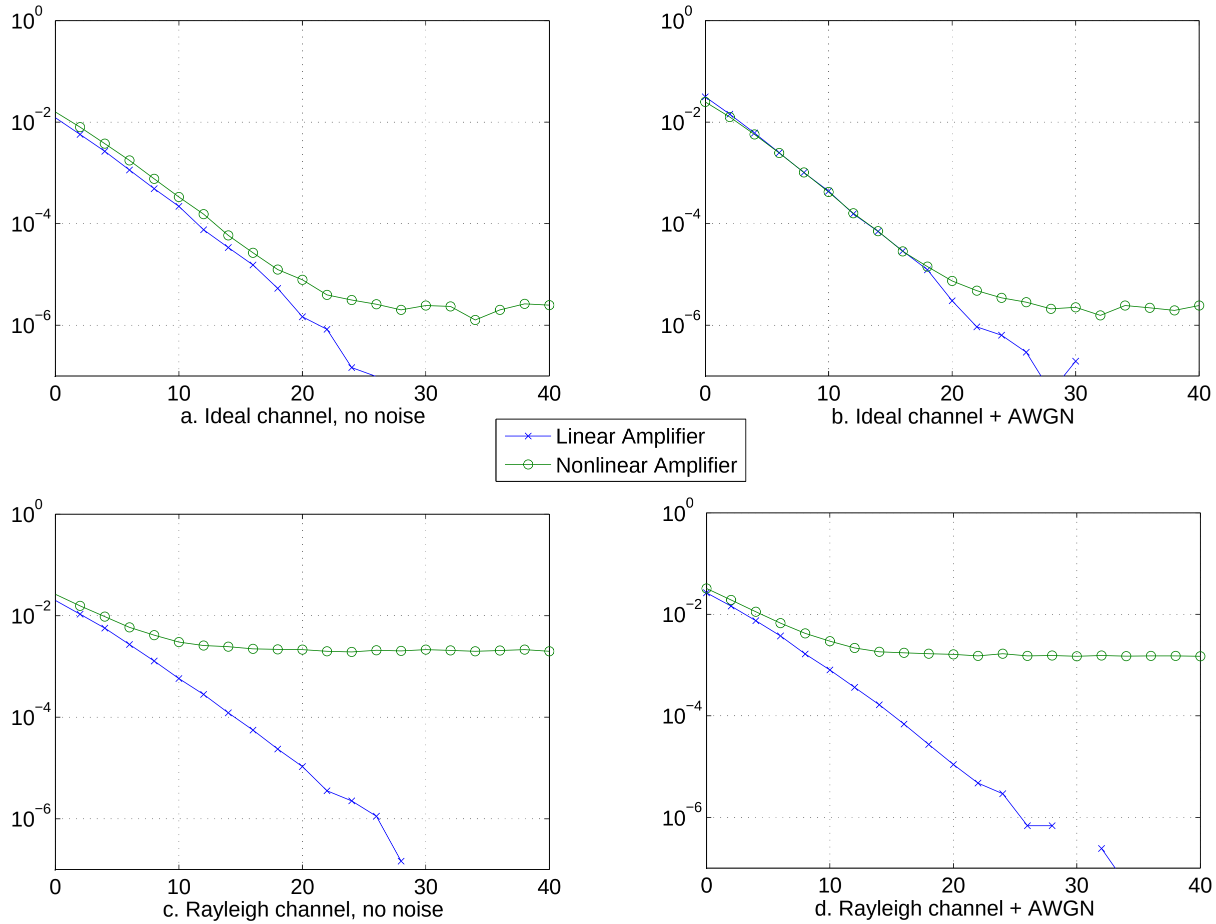

Figure 5.7: BER for different conditions of the S-R branch.

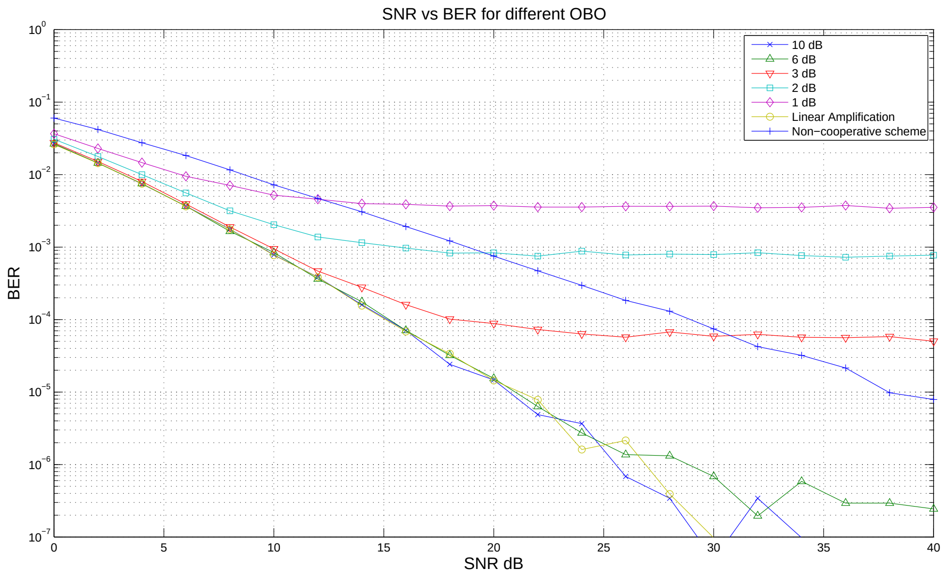

Figure 5.8: BER for different levels of OBO.

below the saturation voltage. However, increasing the OBO results in an inefficient use of resources because the HPAs' maximum efficiency is achieved when they are operating close to their saturation point.

5.7 Summary of the Chapter

When using OFDM in cooperative systems, the NLD becomes an issue that has to be considered carefully. This chapter presents a method for evaluating how the NLD affects the performance of the system, based on simulations. The system model and the basic structure of the simulations is also presented. To evaluate the performance, the BER is chosen as the performance measure.

The NLD in OFDM cooperative systems, as in other systems, produces two components. The in-band distortion and the out-of-band distortion. The first is of more interest since it is the one affecting the link. The out-of-band distortion affects other systems using adjacent bands.

The NLD distortion affects the performance of the system more than expected. It is shown that this is mainly because the signal is distorted by a Rayleigh fading channel before being amplified, causing higher levels of distortion compared with the case when an AWGN channel is used.

A possibility for reducing the NLD is to increase the OBO of the amplifier at the receiver. This improves the performance of the system but it makes it inefficient from the resources point of view. The HPAs are more power-efficient when operated close to their saturation point.

In this chapter it is shown that the NLD is an issue to be carefully considered when using OFDM in cooperative systems. A possible non-optimal solution is also presented. However, having more information of the behaviour of the signals before and after the amplification, could help to improve the performance of the system in a more optimal way. The next chapter analyses the statistical behaviour of the NLD noise at the relay and proposes a solution for mitigating the effects of the NLD.

Chapter 6

The Relay and the Nonlinear Distortion Noise

In this chapter, an analysis of the amplification performed in the relay is presented. The input and output of the amplifier is analysed from the statistical point of view to establish if there is certain predictable behaviour in the NLD. In addition, based on the results of this analysis, a technique to improve the performance of the cooperative system is presented.

6.1 Statistical behaviour of the NLD noise term

As explained in chapter 3, the OFDM symbol has a Gaussian behaviour. Unlike the case analysed in that chapter, in this case the signal is then distorted by a channel and thermal noise before being amplified. The statistical behaviour of such signal may give additional information on how to approach the problem of NLD in cooperative systems. In principle, if the input of the amplifier is Gaussian, then the NLD can be represented as additive Gaussian noise [15]. If this is the case, existing techniques could be applied to improve the performance of the system.

The input of the amplifier y_{SR} can be expressed as:

$$y_{SR} = h_{SR} * x_{SR} + n_{SR} \tag{6.1}$$

It is known that x_{SR} and n_{SR} are complex Gaussian variables with known variance. It is also known that h_{SR} can be modelled as a Rayleigh fading channel, which in principle can be considered as a complex Gaussian variable as well, with known parameters.

The behaviour of y_{SR} depends on how fast h_{SR} changes. Therefore, an analysis of two cases is presented:

- A slow fading channel, where the conditions of h_{SR} between adjacent OFDM symbols also vary randomly but close to the conditions of the previous state.
- A very fast fading channel, where the conditions of h_{SR} between adjacent OFDM symbols change randomly and independently from its previous state.

From the simulations point of view, this means that for the first case, a set of strongly correlated channels are generated for all the symbols to be transmitted, whereas in the second case a new channel is generated every time a new symbol is transmitted. The slow fading channel has a Doppler spread $f_c \approx 10Hz$, resulting from the following parameters in the simulations:

- delay profile: 1, 2, 3, 4 chips
- power profile: 0, -1, -3, -9
- channel impulse response length: 4
- terminal velocity: 5 km/h
- carrier frequency: 2 400 MHz
- bandwidth: 6 000 Hz

In figure 6.1 the pdf of y_{SR}, for a slow fading h_{SR}, is shown together with a Gaussian distribution with the same variance.

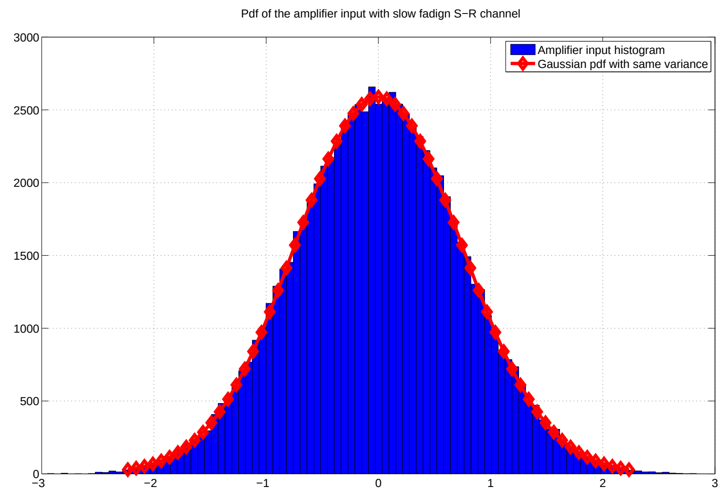

Figure 6.1: PDF of the amplifier input, slow fading case.

The input of the amplifier can be modelled as a Gaussian behaviour. As presented in chapter 3, this means that the NLD can be represented as an additive Gaussian noise. Therefore the effects of the NLD can be measured and mitigated with techniques similar to those used for AWGN.

Figure 6.2 presents the pdf of y_{SR}, for a fast fading h_{SR}, together with a Gaussian distribution with the same variance.

In this case, the behaviour is close but not equal to the Gaussian case. Strictly, the NLD may not be modelled as an additive Gaussian noise. In figure 6.3 the complementary cdf of y_{SR} is compared to that of a Gaussian.

It can be seen that y_{SR} presents a behaviour similar to a Gaussian distribution for values close to the mean but differs significantly for higher values.

When the coherence time of h_{SR} is much larger than the OFDM symbol duration, the channel can be considered to have almost deterministic behaviour. Therefore,

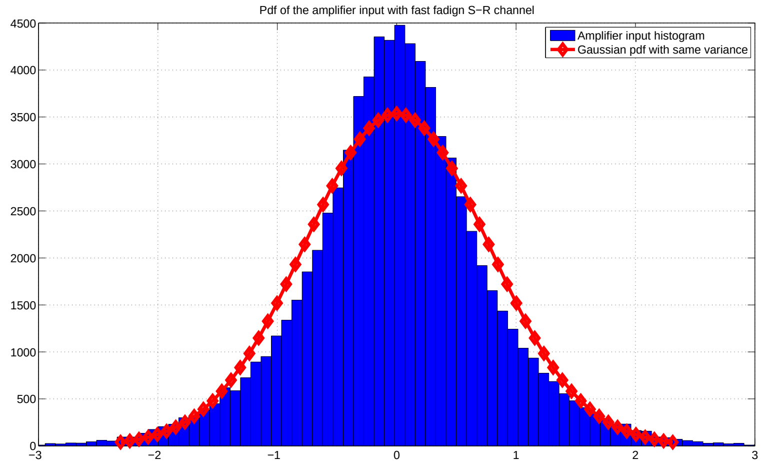

Figure 6.2: PDF of the amplifier input, fast fading case.

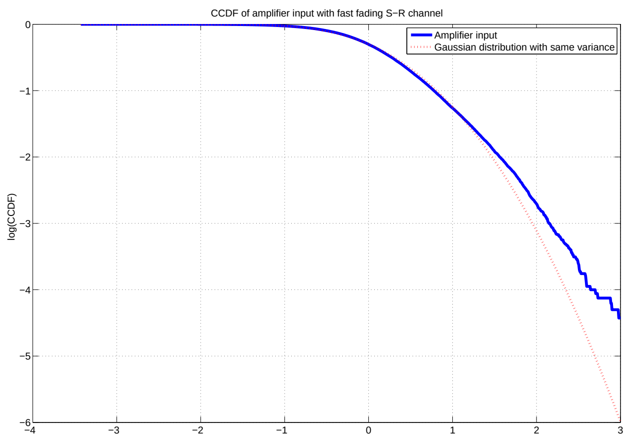

Figure 6.3: CCDF of the amplifier input, fast fading case.

y_{SR} is basically the addition of two Gaussian variables, one of which is much smaller, resulting in a Gaussian variable as well.

This is not the case for a fast fading h_{SR}. The OFDM symbol is affected by a random variable that changes for each OFDM symbol. As a result, y_{SR} is the result of a convolution of two Gaussian variables plus a third Gaussian variable. The distribution of this variable may be difficult to obtain. Furthermore, the distribution of the output of a nonlinear amplifier with this input may be even more complicated. Finding a new model for the NLD noise under these circumstances may therefore be very challenging. However, since the distribution does not differ dramatically from that of a Gaussian, it is worth to evaluate the results of considering it Gaussian.

6.2 The Position of the Relay

For AF cooperative systems, when the NLD is not considered, the best position of the relay is at half of the distance between the source and the destination. If there is an NLD term at the relay, this situation may change. In this section we analyse the performance of the cooperative system varying the distance between the source and the relay.

Two cases are considered. The path-loss model considered is $d^{-\alpha}$ where $\alpha = 4$ is used to describe the loss in a wireless environment.

The first case considers the relay to be always between the source and the destination. Figure 6.4 describes this case.

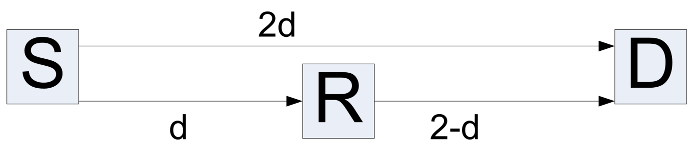

Figure 6.4: Distances between S and R, in-line case.

In figure 6.5 the BER curves for different distances between S and R. When the relay is far from the destination, the performance is very poor. The performance improves as the relay comes closer to the destination. In this case, the best performance is achieved when the relay is in the middle between S and D.

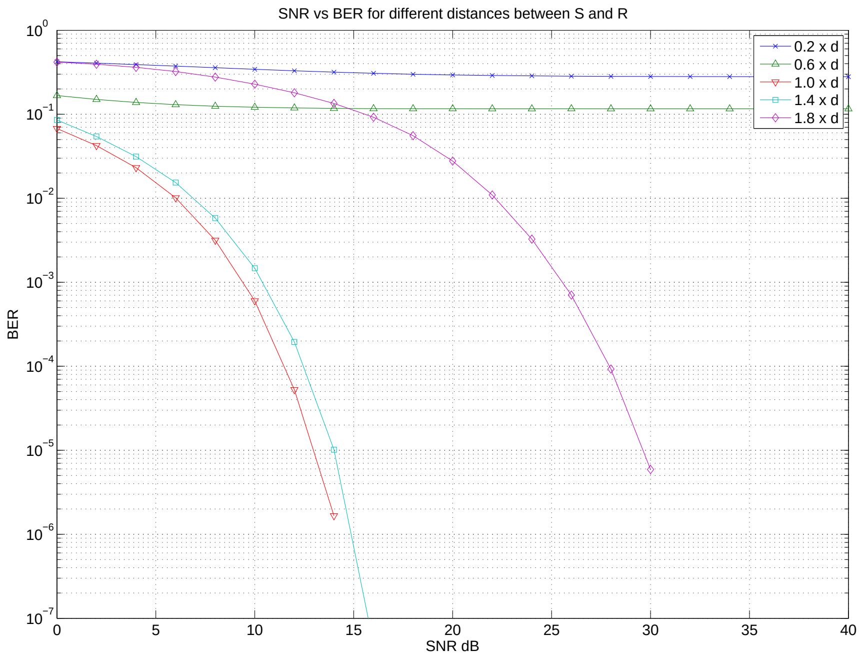

Figure 6.5: BER for different distances between S and R, in-line case.

The second case considers the relay to be somewhere between the source and the destination. This means, in practice, that a triangle is formed among R, S and D. Figure 6.6 describes this case.

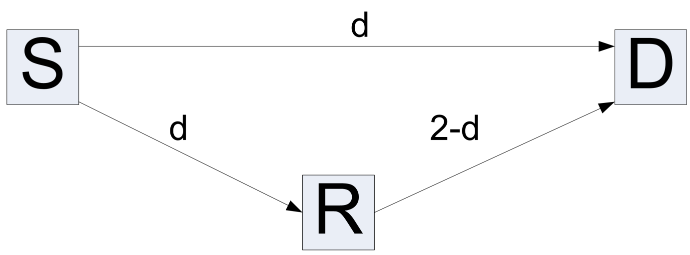

Figure 6.6: Distances between S and R, general case.

As seen from figure 6.7, the performance is better when the relay and the destination are closer. In this case, since the overall S-R-D distance is larger compared to the S-D distance, the received signal from the S-R-D branch has suffered more fading. It can be seen from these results that the system is in general more sensitive to fading after the amplification than to fading before the amplification.

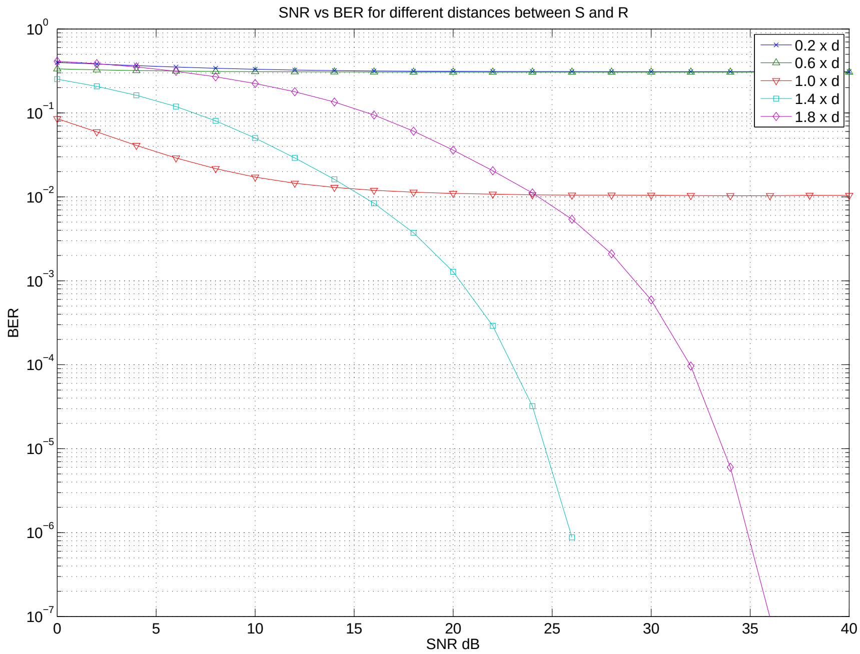

Figure 6.7: BER for different distances between S and R, general case.

In this example, the best performance is also achieved when the distance between R and D is 0.6 of the distance between S-D. Larger distances from the destination impact the performance significantly.

6.3 The NLD noise at the receiver

From the information presented in previous sections, it is possible to identify two possible actions to increase the performance of the cooperative system: to consider the NLD distortion as an AWGN and compensate its effects at the destination or to control the choice of a relay so that it is in the zone where the performance is optimised.

The choice of the relay could be implemented by a control mechanism for the whole system, that may require additional signalling or specific protocols. This is beyond the scope of this thesis and therefore will not be investigated. In [40] a discussion on this issue is presented.

Considering the NLD as an AWGN provides a possibility of compensating the effects of it at the receiver using traditional tools. The following section shows how the NLD noise can be used to optimise the MRC.

6.3.1 Optimisation of the MRC

In section 6.1, it is explained how the NLD noise can be modelled as a Gaussian variable in the case of the slow fading channel and it is close to a Gaussian variable in the case of a fast fading channel. In this section a method for improving the MRC, by considering the NLD noise at the receiver, is presented.

The input of the amplifier can be expressed as:

$$y_{SR} = h_{SR} * x_{SR} + n_{SR} \tag{6.2}$$

The output of the amplifier is then:

$$\begin{aligned} x_{RD} &= F\Big(y_{SR}\Big) && (6.3)\\ &= F\Big(h_{SR} * x_{SR} + n_{SR}\Big) && (6.4) \end{aligned}$$

From equation (3.15), the output of the amplifier can be modelled as the scaled version of the input plus a noise term:

$$
\begin{aligned}
x_{RD} &= K y_{SR} + d(y_{SR}) && (6.5)\\
&= K\Big(h_{SR} * x_{SR} + n_{SR}\Big) + d(y_{SR}) && (6.6)\\
&= K h_{SR} * x_{SR} + K n_{SR} + d(y_{SR}) && (6.7)
\end{aligned}
$$

Performing the change of variable:

$$\hat{n}_{SR} = K n_{SR} + d(y_{SR}) \tag{6.8}$$

The signal sent by the relay to the destination can be expressed as:

$$x_{RD} = K h_{SR} * x_{SR} + \hat{n}_{SR} \tag{6.9}$$

This new distortion term can be applied at the MRC (5.10) presented in chapter 3, in a system with $K = 1$. As a result the new MRC is:

$$Y_D = \frac{H^*_{SD}}{\sigma^2_{SD}} Y_{SD} + \frac{H^*_{SRD}}{\hat{\sigma}^2_{SRD}} Y_{SRD} \tag{6.10}$$

where

$$
\begin{aligned}
\hat{\sigma}^2_{SRD} &= |H_{RD}|^2 \hat{\sigma}^2_{SR} + \sigma^2_{RD} && (6.11)\\
\hat{\sigma}^2_{SR} &= \sigma^2_{SR} + \sigma^2_{NLD} && (6.12)
\end{aligned}
$$

The term

$$\sigma^2_{NLD} = N_c \sigma^2_{nld} \tag{6.13}$$

is the variance of the distortion noise. σ_{nld} is the variance of the time-domain distortion term $d(y_{SR})$ and can be calculated from simulations. This is done by calculating the covariance for several repetitions (which can be considered to be a training sequence of uniformly distributed random bits) of:

$$d(y_{SR}) = x_{RD} - y_{SR} \tag{6.14}$$

That is basically subtracting the input of the amplifier from the output of the amplifier.

6.3.2 Performance of the Optimised MRC

In figure 6.8 the results of this method are presented for the case of a slow fading channel, using 512 OFDM symbols as training sequence. The doppler spread of the channel is $f_d \approx 10$ Hz.

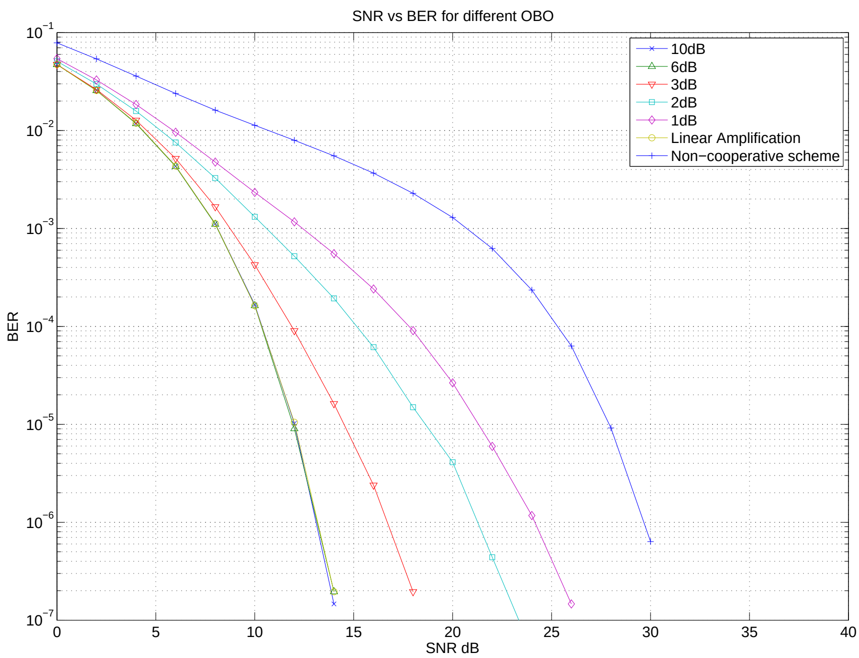

Figure 6.8: BER for different OBO with optimised MRC and slow fading channel.

As it can be seen from the figure, the performance of the cooperative system is significantly enhanced by the use of this method. Since the NLD noise is considered additive, the MRC uses this additional term to define the weight of the S-R-D branch in a more realistic way.

In the case of the slow fading channel, the NLD noise has a Gaussian distribution, resulting in effects equivalent as having AWGN. Therefore, the use of the MRC considering the NLD noise term results in an optimal combiner.

In figure 6.9 similar results are presented for the fast fading channel case. In this case the channel varies randomly between OFDM symbols but remains the same within one OFDM symbol. This figure can be compared to figure 5.8 in chapter 5.

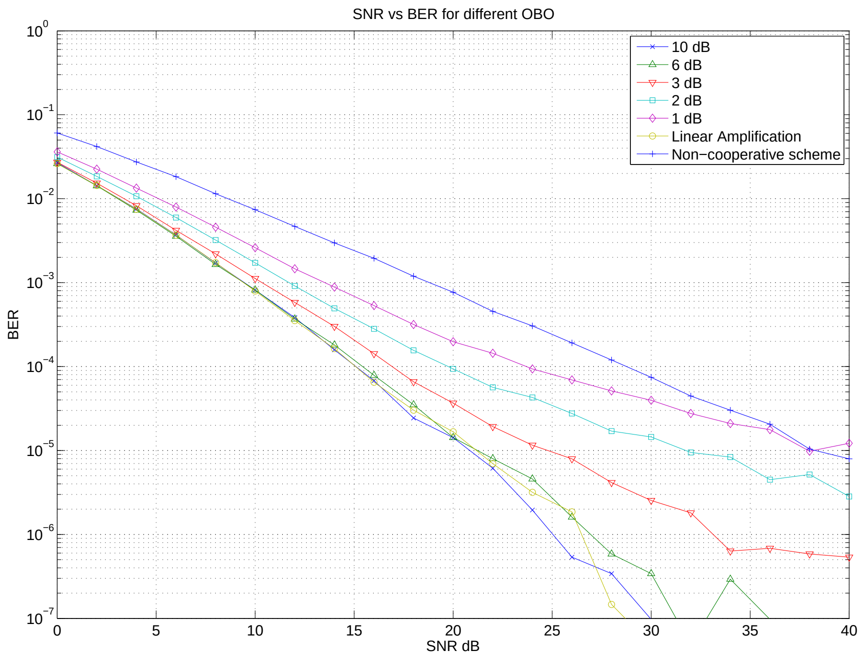

Figure 6.9: BER for different OBO with optimised MRC and fast fading channel.

It is important to consider that for the case of a fast fading channel, the input of the amplifier has, close to, but not Gaussian behaviour. However, the results show that making the assumption of the NLD as an additive Gaussian term improves the performance of the cooperative system.

This could mean that the model presented in [15] may be generalised to other cases than just the Gaussian input, but not necessarily. It could also mean that, even if the NLD term is not exactly Gaussian, assuming that it is provides the MRC with a tool to assign the weight to the S-R-D branch in a better way. This means, that the MRC considers that there is additional distortion on that branch and therefore reduces its weight. This improves the performance of the system significantly, but is not clear if it is possible to obtain a better combiner by modelling the NLD noise in a different way.

The results, in both cases, are significantly better for all levels of OBO, but drastically better for lower OBO levels. As the OBO increases, the NLD noise term decreases, making its consideration less relevant at the receiver. In any case, this method allows decreasing the OBO (i.e. a more power-efficient system) and still achieving better results compared to the non-cooperative scheme.

The results presented here show that considering the NLD noise as an additive Gaussian term, and considering this term at the MRC, improve the performance of the system significantly, compared with the case where the MRC is performed assuming a linear amplification.

6.3.3 Estimation of the NLD Variance

In the previous section in was shown that calculating σ_{NLD} from a training sequence, and using that value to optimise the MRC, improves the performance of the communication system. However, the introduction of a training sequence for doing so represents certain degradation in the system rate efficiency and increases the processing load at the relay. In addition, calculating σ_{NLD} at the relay also implies that its value has to be somehow forwarded to the destination. An alternative could be to calculate this value at the destination which, combined with the channel estimation, may represent an important challenge. In this section, results for different length of training sequences are presented in order to evaluate if the improvement in performance is worth the consequences of having this additional calculation. The assumption is that σ_{NLD} is calculated at the relay and that it is possible to send it to the destination.

Figure 6.10 shows the BER of the cooperative system for different number of training OFDM symbols. This process is performed for two level of OBO: 1 dB and 6 dB.

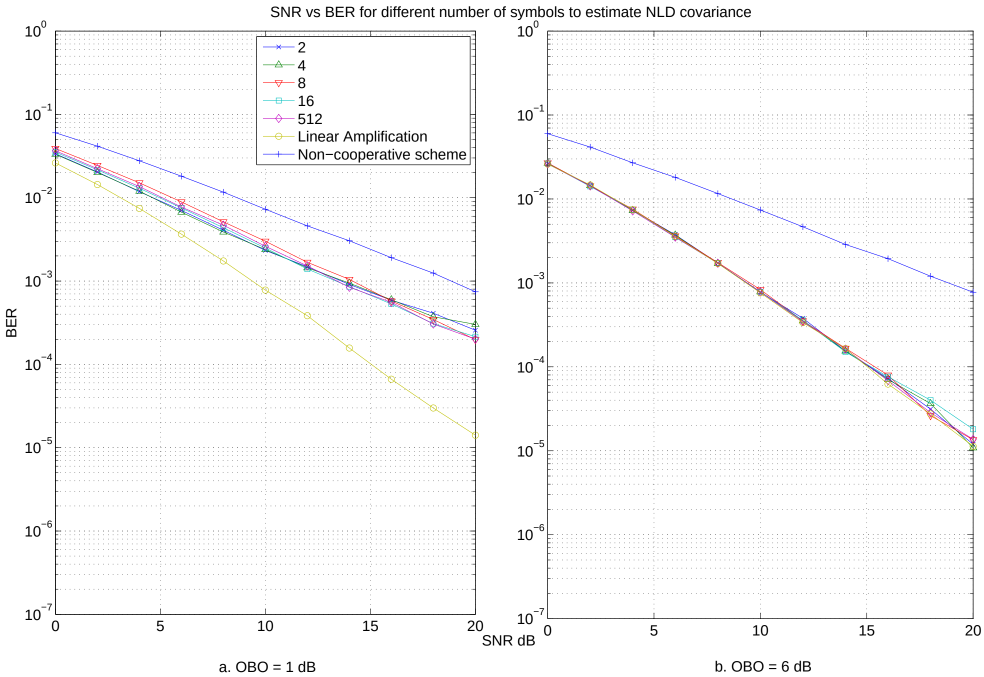

Figure 6.10: BER for different lengths of σ_{NLD} training sequence.

The results are good even for short training sequences. In addition, increasing the length of the training sequence improves the performance only marginally. A short training sequence represents a less negative impact in the resources of the system. The optimised MRC outperforms the regular MRC without the need of long training sequences, minimising the effects in the system rate and the processing load in the relay.

6.4 Summary of the Chapter

In this chapter the role of the relay is analysed. The first section provides a statistical analysis of the amplifier input signal. According to the literature, if the input is Gaussian, the output can be modelled as a scaled version of the input plus an additive Gaussian noise term. The input is shown to be Gaussian for a low fading channel and close to Gaussian for a fast fading channel.

Some analysis about the position of the relay is also presented. Results show that the position of the relay also affects the performance of the amplifier. Unlike linear systems - where the optimal position is at the middle between the relay and the destination - the optimal position is slightly closer to the destination. Choosing the best relay according to the position requires additional control and signalling processes.

Considering the NLD as an additive Gaussian noise, and including it in the MRC, improve the performance of the system significantly. Only short sequences are required to calculate σ_{NLD}, so the effects on system rate and processing load of the relay are minimised. In the case of the low fading channel, it could be considered to be the optimal combiner. In the case of the fast fading channel, the performance is greatly improved but it cannot be considered the optimal combiner. This, because it is not clear if with other model for the NLD, the performance could be improved.

This chapter proposed a method for optimising the combiner and improve the performance of nonlinear cooperative systems. The following chapter introduces a variation to the PANC, presented in chapter 3, for cooperative systems.

Chapter 7

Power Amplifier Nonlinearity Cancellation in Cooperative Communications

In chapter 6 it is shown that optimising the MRC improves the performance of cooperative systems with NLD at the relay. In chapter 3 the PANC technique is described for the case of a non-cooperative system. In this chapter, a modified PANC for cooperative systems is introduced. In addition, the performance of the system with and without PANC is compared. To do so, a relay with saturation voltage of 1.4 V and OBO = 1dB is used.

7.1 The PANC for a Cooperative System

In the system model used for this thesis, the destination receives two signals: only one of those is distorted by NLD. This means that the PANC has to be applied only to the S-R-D branch. In addition, the input of the amplifier is not the source symbol but a channel-distorted version of it. This also has to be considered when implementing the PANC. Figure 7.1 presents a block diagram of the modified PANC for this system.

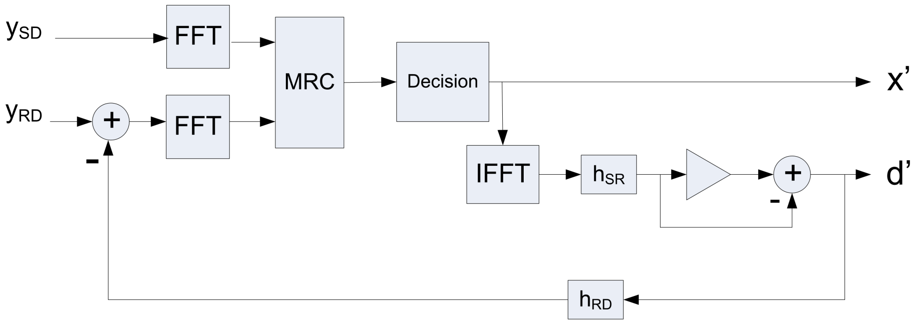

Figure 7.1: PANC model for a cooperative system.

The algorithm of this modification of the PANC is presented in table 7.1.

Modified PANC for Cooperative Communications
For i = 1 to number of iterations.
1. Demap received symbol.
2. Transform to time domain.
3. Convolve with channel.
4. Amplify using the amplifier model.
5. Compute the distortion term.
6. Subtract the distortion term from y_{RD}.
7. Transform to frequency domain and combine.
End

Table 7.1: The modified PANC algorithm

The first important difference is that the estimation of the distortion term d' is subtracted only from y_{RD}. The second big difference is that the estimation of the received symbol x' is convolved with h_{SR} after the transformation, before sending it to the amplifier model. Like the PANC presented in chapter 3, the channels and the model of the amplifier are assumed to be known at the destination.

The PANC uses x' to estimate the distortion d' that would have been generated if the sent symbol was in fact x'. After subtracting d' from y_{RD} a better estimation of the sent symbol is obtained. When this process is performed several times, x' is then closer to the actual sent symbol.

7.2 System with PANC and a regular MRC

In chapter 6 a method for optimising the MRC was presented. This required certain additional processing by the relay and a method for transmitting this information to the destination. In this section, PANC is used for an MRC that is not considering the NLD. In other words, an MRC that is optimised for a linear cooperative system. The idea is to evaluate if the performance is improved without having to estimate σ_{NLD} at the relay. Figure 7.2 shows the BER curves for this case.

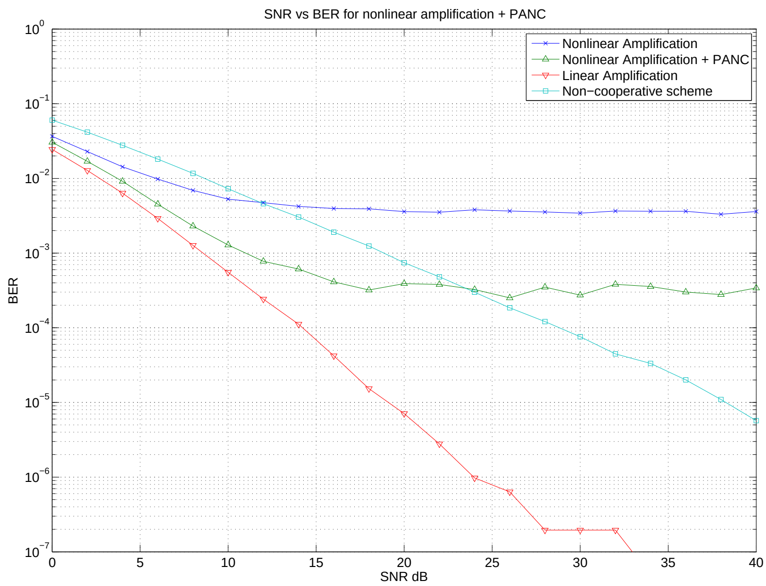

Figure 7.2: Performance of PANC with a regular MRC.

The performance is improved significantly. However, there is a lower bound where after a 15 dB SNR, the performance does not improve any more. This case could be an option if the cooperative system wants to be kept as simple as possible. Yet, it is hard to tell if the performance gain justifies the increased complexity with respect to the non-cooperative scheme.

7.3 System with PANC and an optimised MRC

In this section, PANC is applied to a cooperative system with an optimised MRC. That is the MRC presented in chapter 6. The optimised MRC improves the performance of the cooperative system under NLD. The idea here, is to evaluate if that performance can be improved further. In figure 7.3 the BER curves for this case are presented.

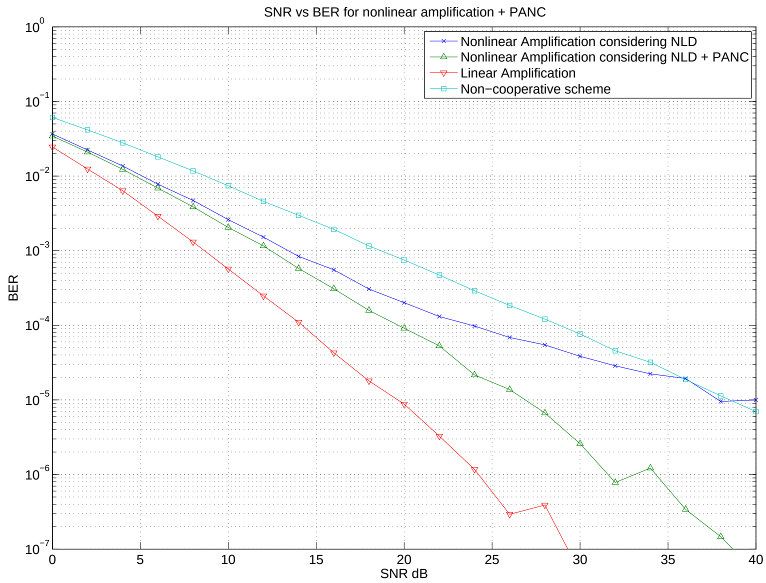

Figure 7.3: Performance of PANC with an optimised MRC.

The curves show a marginal improvement for low SNRs and a good improvement for higher SNR. If we consider the 20 dB SNR a common operating point, the improvement in performance is not that significant.

There is one important disadvantage of combining the optimised MRC with the PANC technique. The optimised MRC considers the NLD to weight the received signals but the PANC technique objective is to remove the NLD term from the received signal. Therefore the optimised MRC is optimal for the reception of the signal y_{RD} but not for $y_{RD} - d'$.

7.4 System with PANC and an intelligent MRC

The PANC and the optimised MRC do not perform extremely well together. However, the improvement in performance, though marginal, gives room to explore this possibility further. To do so, two steps have to be considered.

The first step is when reception takes place. In this case, the optimised MRC has proved to be a good choice.

The second step is when the PANC is being performed. The sample of the received signal is somehow stored during this time. Since after the first iteration, the estimated distortion term is removed from the received signal, the optimised MRC is no longer valid for this case.

One option would be to re-estimate the parameters of the MRC for every iteration. This is computationally demanding and technically very complicated.

The second option is to consider the operation of the PANC. After the first iteration, the PANC is subtracting a d' already close to the actual distortion term. Now, assuming that it is not just close but equal, the MRC to use in this case is the same as the regular MRC (or the MRC optimised for the linear case).

As a result an intelligent or hybrid MRC that uses the optimised version of the MRC on rececption, but the regular MRC for the PANC process, can be implemented. In figure 7.4 the BER curves for this case are presented.

In this case the performance is improved significantly by the use of PANC. The BER curve is not just close to the linear case but it also shows a similar behaviour. In principle this means that both elements are doing their part. The intelligent MRC, when working in the optimised mode, gives results in a first estimation of the sent symbol. This estimation is good enough for the PANC to make an accurate estimation of the NLD term and subtract it from the received symbol. Once this is done, the symbol can be considered almost distortion-free and the intelligent MRC, working in the regular mode, performs the combination accordingly.

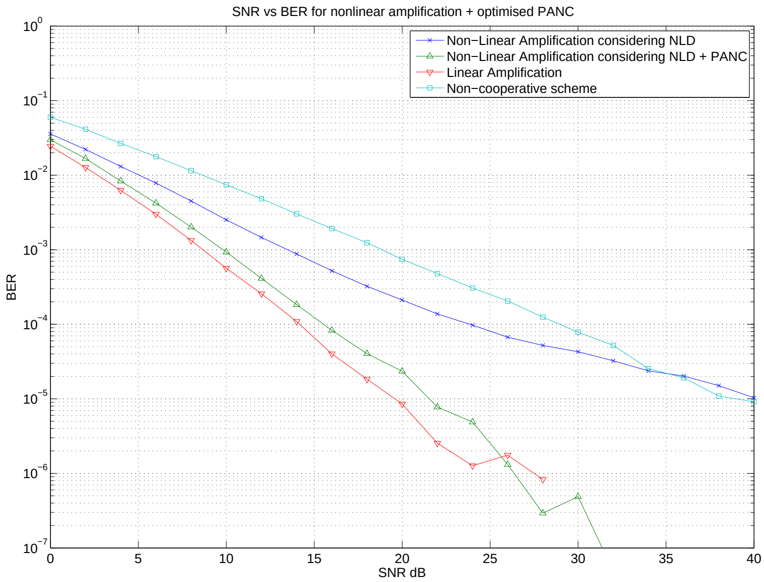

Figure 7.4: Performance of PANC with an intelligent MRC.

In figure 7.5 the performance of the three different options of PANC are presented. Here it is possible to appreciate that PANC with intelligent MRC performs better for all cases. Between the other two, the normal MRC + PANC shows better results for low SNR and the optimised MRC + PANC performs better for higher SNR. The difference in processing load between the optimal MRC and the intelligent MRC is minimal. Therefore the choice on how to apply PANC depends on whether the information about σ_{NLD} is available at the destination. In the cases when it is, the intelligent MRC should be used when performing PANC.

7.5 Summary of the Chapter

In this chapter a modification to the PANC technique is introduced for cooperative systems. This modified technique is tested using simulations. The PANC improves the performance of the system in all cases, however the improvement in performance when a regular MRC combiner is used is not very significant. On the other hand,

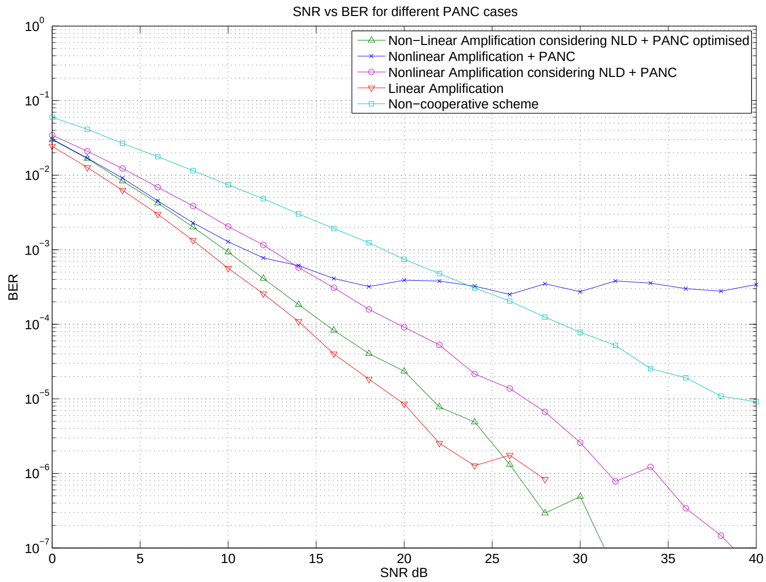

Figure 7.5: Performance of the three possible PANC modes.

combining the optimised MRC with the PANC technique requires re-considering the MRC. An intelligent MRC, that works as an optimised MRC during reception and as a regular MRC during PANC, should be used.

Chapter 8

Practical Considerations

The solutions and methods presented in previous chapters consider the knowledge of certain information in the system. This information includes the channel state information (CSI), the parameters of the nonlinear amplifier and the variance of the thermal noise. In practice, most of these factors are unknown and must be estimated. Particularly, the estimation of the CSI and the amplifier parameters in a cooperative system is relatively more complex than in the case of a one-to-one system. This chapter presents an overview of the challenges and options for the estimation of the different variables.

8.1 Channel Estimation at the Destination

The destination must either have the knowledge or perform the estimation of all the channels involved in the system. The link which represents a bigger challenge is the S-R-D link.

Channel estimation in an OFDM system with the presence of a nonlinear amplifier is described in [43]. In this case, the coherent time of the channel is an important factor to consider.

For slow fading channels, it is possible to use some OFDM training symbols inserted in regular intervals. This is also known as a training symbol based approach [44]. The advantage of having a full OFDM symbol as a training sequence is that it is possible to control its PAPR. Therefore, the training symbol is chosen to have a low PAPR to minimise the effects of the nonlinear amplifier so the channel can be accurately estimated.

For fast fading channels, having full OFDM symbols as training sequence may significantly decrease the rate of the system. In these cases, training symbols are inserted in certain carriers of each OFDM symbol, creating a set of pilot carriers. In this pilot-tone-multiplexed approach [44], the channel estimation of the data carriers (those that are not pilot carriers) is performed by interpolation. For example, in [45, 46] interpolation is performed using truncated DFT matrices.

In cooperative systems the problem of the amplifier is more complex given the presence of a channel before the signal is amplified. This has to be carefully considered. Following a similar approach than before, providing training sequences with low PAPR will minimise the distortion caused by the amplifier. In [47] the channel estimation problem for an OFDM AF cooperative system (without nonlinear distortion) is discussed.

An option worth considering is that each element in the cooperative system estimates the channel towards its closest cooperation partner and then uses a control network or some signalling protocol to forward this information to the rest of the elements. In a first approach this may seem significantly inefficient, since it requires that each element performs the estimation, and additional spectral resources to transmit this information. However, considering a fully cooperative scheme, a mobile device can be a relay and a destination almost simultaneously. This means that the effort in terms of processing load is not increased. In addition, having several elements performing estimation could mean less frequent training sequences, compensating the need of additional resources for signalling and control. It is also important to evaluate if, since the channel has been already estimated at each element, the option of decode and forward is a better alternative. Although it is not clear whether this option is viable, it is worth exploring it further.

8.2 NLD Variance Estimation

It is clear that the easiest place to estimate the NLD variance is at the relay. This is, because at the relay it is only required to subtract the the input from the amplifier from the output of the amplifier, and then calculate the variance of that. However, this means that this information should be forwarded to the destination so it can be applied in the MRC. Provided that there would be already a signalling protocol, used for example to transmit CSI information, this solution is viable. But implementing a signalling or control solution just for transmitting this information is not practical. In any case, it is also important to consider the quantisation factor of the estimation. Using more bits for the quantisation of the NLD variance provides more precise results but reduces the rate of the system.

If the NLD variance must be estimated at the destination, the process may be more complex. However, from the results presented in chapter 6 it is clear than a rough estimation is sufficient to have a good performance. A possible solution would be to calculate the variance of the overall noise power on the S-R-D branch and use that value directly as the total noise on that branch. This can be done using a training sequence and after the channels have been estimated. This option is worth exploring further since in addition to simplifying the calculation of the NLD variance, it also could also provide a more efficient way for implementing the MRC.

8.3 Estimation of Additional Information Needed for PANC

The implementation of the PANC requires two factors to be known and which estimation is quite complex. These are the parameters of the amplifier and the

CSI of the S-R link. The estimation of these factors is quite complex because of the structure of the S-R-D link.

In the previous section, the issue of estimating the CSI was presented for the S-R-D link. There it was discussed how to estimate the equivalent h_{SRD}. However, the PANC process requires the value of only h_{SR} in addition to the parameters to model the amplifier.

The problem is reduced to the identification of a system consisting of h_{SR} followed by a nonlinear function $F(\cdot)$ followed by h_{RD}. This can be modelled as a Wiener-Hammerstein model [48] and it is a problem comparable to the identification of nonlinear amplifiers with memory. The channel estimation of h_{SRD} can be performed using a low PAPR sequence. If h_{SRD} is compensated at the destination, the system could be simplified to a Wiener-Hammerstein model where the second linear filter is the inverse of the first (h_{SR}). Some method, like the one presented in [49] could then be used to identify the model and obtain the desired estimates.

8.4 Summary of the Chapter

In this chapter, some practical issues concerning the implementation of the methods proposed in this thesis are presented. These methods consider that the CSI of all the channels as well as the characteristics of the amplifier and its distortion are known by the destination node. In practice this is not the case. Some possibilities for the estimations are discussed, and some previous work in the subject is introduced. This chapter can be considered a good framework for future work on the subject.

Chapter 9

Conclusions and Future Work

The implementation of OFDM systems results very efficient due to the use of IFFT and FFT algorithms. By exploiting the properties of the Fourier transform and the addition of the cyclic prefix, it is possible to implement very reliable systems with high immunity to ICI and ISI. Its spectral efficiency and higher immunity to multipath environments make OFDM a strong candidate for a number of future applications. However there are some issues to be considered e.g. the need of very accurate frequency and time synchronisation. One of the main complication of OFDM is its high PAPR which makes it very sensitive to nonlinearities in the amplifier.

OFDM systems are highly sensitive to NLD because of the high PAPR inherent to this modulation technique. Unfortunately, HPAs are more efficient when operated close to their saturation point, where the nonlinearities increase. The impact on the performance has to be evaluated to find a good trade-off between efficiency and performance. The NLD term can be modelled as an additive Gaussian noise in the system. Several techniques exist for mitigating the effects of NLD noise, specifically the power amplifier nonlinearity cancellation (PANC) technique can be applied at the receiver in cases where other techniques are not available at the transmitter.

Cooperative communications are being studied as a source of diversity in cases where having multiple antennas on a single device is not practical. The factors that define how a cooperative system works are the cooperation method, the modulation and demodulation method and the combining method. In this thesis we studied a coherent AF cooperative system with a single relay and MRC. The combination at the destination must be done considering that the S-R-D branch has two sources of noise. Therefore the branches are not equally weighted.

When using OFDM in cooperative systems, the NLD becomes an issue that has to be considered carefully. In this thesis it is demonstrated that the NLD in these systems cause a serious degradation in the BER. In addition, the main cause of distortion has been identified. The S-R channel plays an important role in the resulting NLD of the system. Increasing the OBO of the amplifier improves the performance in terms of BER but it makes the system very inefficient from the power resources point of view.

After studying the behaviour of the relay in more detail, it was shown that the input of the amplifier at the relay is Gaussian for slow fading channels but it is not clearly defined for fast fading channels. The fact of having a Gaussian input allows the assumption of the NLD as an additive Gaussian noise. This approach is also tested for the fast fading channels as the behaviour is not Gaussian but it is somehow close to it.

A new option for implementing the MRC at the destination was proposed. In this option, the variance of the NLD, considered as an additive Gaussian noise, is included in the MRC, improving the performance of the system significantly. In addition it was demonstrated that this option requires only short sequences to calculate this variance, so the effects on system rate and processing load of the relay are minimised. In the case of the low fading channel, it could be considered to be the optimal combiner. In the case of the fast fading channel, the performance is greatly improved but it is not clear if it is the optimal combiner.

In addition, a modification to the PANC technique for cooperative systems is introduced. The PANC improves the performance of the system in all cases. It was also shown that to maximise the performance of the system, an intelligent MRC, that works as an optimised MRC during reception and as a regular MRC during PANC, should be used.

Some practical issues regarding the proposed methods were also discussed. The procedures presented in this thesis were made assuming the knowledge of channels and amplifier model, and also assuming that the information available at the relay can be somehow forwarded to the receiver. Since this is usually not true in real systems, a discussion of possible techniques to estimate these variables was presented.

The discussion presented in chapter 8 can be considered as a framework for further work. The estimation of channels and amplifier model are, without doubt, an issue worth to be investigated. Another branch of future work could be concentrated on modelling the distortion term in cases of fast fading channels; having a better model may give way to new techniques to mitigate its effects. Other situation to be considered is the study of nonlinearity effects when several relays are used, in parallel or in series.

The results presented in this thesis are relevant and promising. Including the nonlinearity issues in the study of cooperative communications brings them closer to reality. In this thesis it was demonstrated that, although the NLD causes a severe degradation in the performance of cooperative systems (specifically in the BER), it is possible to mitigate the effects of the NLD using known techniques and adapting them to the cooperative case. Furthermore, applying these tools results in a performance close to the linear case.

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