Contribution to the characterization and evaluation of multiple antenna systems for communications: prototyping, propagation and antenna perspective

TESIS DOCTORAL

Laura García García
Ingeniera de Telecomunicación
2007
Departamento de Señales, Sistemas y Radiocomunicaciones
Grupo de Radiación

ESCUELA TÉCNICA SUPERIOR
DE INGENIEROS DE TELECOMUNICACIÓN

Contribution to the characterization and evaluation of multiple antenna systems for communications: prototyping, propagation and antenna perspective

TESIS DOCTORAL

Autora: Laura García García
Ingeniera de Telecomunicación

Director: Leandro de Haro Ariet
Doctor Ingeniero de Telecomunicación

Madrid, 2007
ABSTRACT

The advent of new services and applications in mobile and wireless communications, most of them based on high data rates, is helping to reactivate the Telecommunication market. Moreover, current trends seem to point towards the preference for wireless connections to Internet instead of the traditional wired ones, towards the growing of data exchanges (for example with web 2.0 and peer-to-peer applications) and towards the exponential increase in the number of devices connected to Internet (as in the proposed future “Internet of things”). A clear consequence of all these trends is the significant increase in data rate requirements for wireless and mobile communications. As a result, research on new techniques that allow to increase data rate and system performance has been boosted by the market needs.

Having intensively utilized frequency, time and even code diversity resources, spatial diversity offered by the radio channel has been proposed as an alternate method of great interest to be exploited. Advances in signal processing platforms have improved the computational capacities in digital signal processing, opening new options for signal processing for antenna arrays. The main objective of this thesis has been to investigate the possibilities for radiocommunication systems that are based on the use of multiple antennas to take advantage of the available spatial diversity in the channel. Moreover, current open issues and problems that prevent multi-antenna systems from being widely adopted have been thoroughly analysed, with special emphasis on real implementation issues. Novel approaches to evaluate multi-antenna systems have been proposed, and results regarding real prototypes have been presented and discussed.

Multi-antenna systems for wireless and mobile communications is a very wide topic, and it would be impossible to cover every aspect on this topic. Therefore, the work in this thesis has been focused on four main aspects, which cover a quite multidisciplinary research area but are somehow interlinked. Those are: 1) adaptive antennas for mobile systems, 2) rapid prototyping for MIMO, 3) propagation aspects of MIMO systems and 4) antenna aspects of MIMO systems.

The first step in exploiting spatial information and diversity in the channel consists in using multiple antennas at one end of the link (typically the base station) together with beamforming signal processing, forming the so-called adaptive antennas. Despite their undeniable advantages, there are some issues that make it difficult the adoption of adaptive antennas in real networks. We have analysed these issues, showing that a new simplified approach of adaptive antenna that allows a transparent operation from the network point of view is possible, at the cost of a small reduction in the adaptive antenna performance. A prototype based on the software-radio concept and with real-time operation has been implemented and its features have been evaluated. A new methodology to easily evaluate adaptive antennas has also been presented. Finally, a set-up module that allows seamless integration of the implemented
adaptive antenna with a real 3G-UMTS network has been designed, implemented and tested under realistic circumstances.

The natural next step in the spatial diversity world is using multiple antennas not only at one link end but at both transmitter and receiver. These systems are usually called multiple-input multiple-output (MIMO) systems, and can be based on different signal processing approaches, as spatial diversity, spatial multiplexing, etc. Taking advantage of the experience obtained from the development of an adaptive antenna, we have design and implemented a novel MIMO testbed. Many MIMO prototypes and testbeds have been presented in recent years. In spite of this, there are still some open issues regarding MIMO channel characterization and effect of realistic impairments, as well as the effect of antenna array configurations in the system. The main interest of our MIMO testbed is that it has been designed in order to allow studying these three aspects in an easy way for the user. Some results obtained from a measurement campaign carried out the the new MIMO testbed have been evaluated and discussed, being of special interest the comparison of single- and multi-polarized MIMO systems and the analysis of effect of channel normalization in capacity computation.

In order to gain a better insight into other propagation characteristics of the MIMO channel, a complete measurement campaign was carried out. This time the main objective was to compare indoor and outdoor-to-indoor MIMO channel characteristics, since these scenarios are of great interest for low-mobility applications in offices (wireless access) and other indoor scenarios. We have included a detailed analysis of parameters as correlation, coverage or received power and available capacity for a 4×4 system. This study has complemented the previous one by including different combinations of base stations when one or two base stations are used simultaneously. Advantages and drawbacks of each scenario and base station locations have been drawn, which may help system deployments.

Finally, in order to cover all the elements in the MIMO system (channel, base station and user equipment), realistic user devices have been studied. In order to overcome problems in size and weight, we have proved that compact arrays may be a suitable solution. A novel multi-element antenna for handheld devices has been designed and implemented. Several methods to evaluate its performance have been proposed and their results compared, such as standardized MIMO channels, computation of antenna covariance and measurements with a MIMO testbed. The results in terms of accuracy of results and computation or time cost can be extrapolated to the evaluation of any antenna array for MIMO systems.

The contributions of this thesis have been presented in several articles published in international technical journals, as well as in the most important international and national conferences in antennas and signal processing for communications.
RESUMEN

La continua propuesta de nuevos servicios y aplicaciones en comunicaciones móviles e inalámbricas, en gran parte con requisitos de altas tasas binarias de datos, está impulsando el crecimiento del mercado de las telecomunicaciones. Por otra parte, las tendencias actuales apuntan a la preferencia por conexiones inalámbricas a Internet en vez de las tradicionales conexiones por cable, al incremento de intercambio de datos (por ejemplo con la web 2.0 y las aplicaciones peer-to-peer) y al aumento exponencial en el número de los dispositivos conectados a Internet (como la propuesta de la futura "Internet de cosas"). Una consecuencia clara de todas estas tendencias es el aumento significativo en los requisitos de tasa de datos que las comunicaciones inalámbricas y móviles deben ser capaces de ofrecer. Como resultado de esto, la investigación de nuevas técnicas capaces de ofrecer ese aumento en la tasa de datos y la mejora en el funcionamiento del sistema ha sido y está siendo impulsada en los últimos años.

Tras haber aprovechado al máximo los recursos de diversidad en frecuencia, tiempo e incluso código, la diversidad espacial ofrecida por el canal radio ha sido propuesta como un método alternativo a ser explotado de gran interés. Los avances en las plataformas de procesado de señal han mejorado la capacidad computacional del procesado, abriendo las nuevas opciones para el proceso de señal aplicado a arrays de antenas. Así, el objetivo principal de esta tesis ha sido investigar las posibilidades de los sistemas de la radiocomunicación basados en el uso de antenas múltiples para aprovechar la diversidad espacial disponible en el canal. Además, se han analizado aspectos no resueltos y problemas que impiden que los sistemas de múltiples antenas puedan ser desplegados y adoptados de forma masiva. Se ha puesto un especial énfasis en aspectos de implementación real. También se han propuesto nuevos enfoques para evaluar los sistemas multiantenna, y se han presentado y comentado los resultados obtenidos de los prototipos y medidas en sistema real.

El estudio de sistemas de múltiples antenas para comunicaciones inalámbricas es un tema muy amplio y sería imposible cubrir todos sus aspectos en una tesis. Por eso se ha centrado el trabajo de esta tesis en 4 puntos principales, que han sido seleccionados tras un primer trabajo de análisis de estado del arte. Estos son: 1) antenas adaptativas para sistemas móviles, 2) prototipado rápido para MIMO, 3) aspectos de propagación de los sistemas MIMO y 4) aspectos de antena de los sistemas MIMO.

El primer paso en explotar la información y la diversidad espaciales en el canal consiste en usar múltiples antenas únicamente en uno de los extremos del enlace (típicamente en la estación base) junto con el procesado de conformación correspondiente, dando lugar a las antenas adaptativas. A pesar de sus innegables ventajas, hay ciertos aspectos que dificultan la adopción global de las antenas adaptativas, como son su complejidad y alto coste. En esta tesis se han analizado estos aspectos, mostrando que un nuevo modelo simplificado de antena adaptativa es posible, de modo que se permita su funcionamiento de forma transparente a la
Resumen

estación base y la red real. El coste que hay que pagar es un empeoramiento de las prestaciones de la antena, que a pesar de ello sigue mejorando significativamente las prestaciones de una antena sectorial convencional. Se ha implementado un prototipo basado en el concepto de software-radio, con funcionamiento en tiempo real, y se han evaluado sus características. Además, y puesto que los métodos de evaluación de antenas convencionales no son válidos en general para antenas adaptativas, se ha propuesto una nueva metodología de evaluación fácil de llevar a cabo. Finalmente, se ha diseñado e implementado un módulo de set-up para la integración transparente de la antena adaptativa en una red UMTS real, se ha probado su funcionamiento y se han presentado sus puntos fuertes y también sus deficiencias.

El paso siguiente en el objetivo de aprovechar la diversidad espacial consiste en utilizar múltiples antenas no sólo en un lado del sistema, sino en ambos (transmisor y receptor): es lo que típicamente se denominan sistemas MIMO (Multiple-input multiple-output). Éstos se pueden basar en distintos esquemas de procesado, como la multiplexación espacial, los denominados de diversidad espacial, etc. Aprovechando la experiencia obtenida del desarrollo de la antena adaptativa, se ha diseñado y desarrollado un nuevo testbed MIMO. Existen muchos testbeds MIMO presentado en la literatura recientemente; sin embargo, el interés del presentado en esta tesis radica en que permite estudiar múltiples aspectos de los sistemas MIMO de forma fácil y cómoda para el usuario. En concreto, se ha hecho hincapié en el diseño del MIMO testbed en que éste incluyese efectos no ideales de sistemas reales, que permitiese evaluar distintas configuraciones de array y que ayudase a probar fácilmente algoritmos MIMO y a realizar medidas de canal MIMO. En este bloque de trabajo se han presentado algunos resultados obtenidos de una campaña de medidas realizada con este nuevo testbed MIMO, cuyos resultados se han evaluado y comentado. De especial interés es la comparación que se ha realizado entre sistemas con una y con varias polarizaciones, así como el análisis del efecto la normalización del canal en el cálculo de la capacidad.

Para comprender mejor otras características de propagación del canal MIMO, la campaña de medidas previa se ha complementado con otra campaña de medidas muy completa, que se ha enfocado al estudio y comparación de entornos indoor con los entornos outdoor-indoor. Hemos incluido un análisis detallado de parámetros como correlación, cobertura o capacidad disponible para un sistema 4×4. Además, se han estudiado diversos emplazamientos para la estación base y posibles combinaciones con y sin información del canal en el transmisor, proponiendo soluciones realistas para un despliegue en interiores. Se han presentado las ventajas y desventajas de cada escenario y emplazamiento de antena, tratando de presentar consejos de despliegue en líneas lo más generales posible.

Finalmente, para cubrir todos los elementos en el sistema MIMO (canal, estación base y equipo de usuario), se han estudiado dispositivos multiantena de usuario que fuesen realistas. Para cumplir los requisitos de tamaño y peso, se han presentado las soluciones de los arrays
compactos. En concreto se ha diseñado un array compacto para dispositivo de usuario tipo PDA. Se ha validado su buen funcionamiento para sistemas MIMO mediante varios métodos propuestos para evaluar antenas para MIMO. Se han comparado dichos métodos en cuanto a precisión en resultado y requisitos de tiempo y computacionales, mostrando así sus ventajas y desventajas de modo que éstas se puedan extrapolar para la evaluación de antenas para MIMO en general.

Las contribuciones de esta tesis han sido plasmadas en varios artículos publicados en múltiples revistas científico-técnicas de reconocido prestigio. Además, los principales resultados se han presentado en congresos internacionales y nacionales de renombre en el área de las antenas y el procesado de señal para comunicaciones móviles e inalámbricas.
SECTION I – RESUMEN AMPLIADO (EXTENDED ABSTRACT IN SPANISH)

I.1 Introducción y objetivos........................................................................................................xxiii
  I.1.1 Introducción al ámbito de la tesis.................................................................................. xxiii
  I.1.2 Motivación .................................................................................................................... xxiii
  I.1.3 Objetivos ...................................................................................................................... xxiv
  I.1.4 Resumen de principales aspectos novedosos de la tesis............................................. xxv
  I.1.5 Plan de desarrollo y metodología y seguidos ........................................................... xxvi
  I.1.6 Organización del documento de la tesis ................................................................. xxvii

I.2 Antena adaptativa para sistemas de comunicaciones móviles: diseño, implementación y evaluación de los módulos de procesado de señal ....................xxix
  I.2.1 Introducción .............................................................................................................. xxix
  I.2.2 Descripción general del sistema de antena adaptativa.......................................... xxix
  I.2.3 Arquitectura hardware ............................................................................................ xxxii
  I.2.4 Evaluación de la antena adaptativa.......................................................................... xxxiii
  I.2.5 Set-up para la antena adaptativa ............................................................................. xxxiv
  I.2.6 Principales resultados ............................................................................................. xxxiv

I.3 Prototipado rápido para MIMO: diseño, implementación y medidas con un demostrador MIMO para WLAN .....................................................xxxvi
  I.3.1 Introducción ............................................................................................................... xxxvi
  I.3.2 Arquitectura y descripción modular del demostrador MIMO............................... xxxvi
  I.3.3 Medidas de canal y resultados ................................................................................. xxxvi

I.4 Aspectos de propagación en sistemas MIMO: análisis con múltiples emplazamientos de estación base y múltiples escenario ........................................xl
  I.4.1 Introducción ................................................................................................................ xl
  I.4.2 Sistema y entornos de medida................................................................................... xl
  I.4.3 Medidas y resultados .................................................................................................. xlii

I.5 Aspectos de antena en sistemas MIMO: diseño, implementación de array de antenas y métodos de evaluación para MIMO .............................................xliv
Table of Contents

I.5.1 Introducción ............................................................................................................. xliv
I.5.2 Diseño de un array compacto para MIMO ............................................................. xliv
I.5.3 Métodos de evaluación de arrays de antenas para MIMO ........................................ xlv

I.6 Conclusiones y líneas futuras .............................................................................. xlvii
I.6.1 Conclusiones ............................................................................................................ xlvii
I.6.2 Líneas futuras ............................................................................................................ xlix

SECTION II – MAIN DOCUMENT

1. INTRODUCTION ....................................................................................................... 1
   1.1 Motivation ........................................................................................................... 4
   1.2 Objectives and general methodology of the thesis ........................................... 8
   1.3 Outline of the thesis .......................................................................................... 11

2. THEORETICAL BACKGROUND AND STATE OF THE ART ......................... 13
   2.1 Introduction and terminology: smart antennas, MIMO systems,
      beamforming and signal processing for spatial diversity ............................. 16
      2.1.1 Introduction ................................................................................................ 16
      2.1.2 Used terminology in this thesis ................................................................. 17
   2.2 A first step in multi-antenna techniques: smart antennas based on
      beamforming ...................................................................................................... 18
      2.2.1 Basis concepts on smart antennas and beamforming ......................... 18
      2.2.2 Beamforming and adaptive algorithms ............................................... 19
      2.2.3 Prototyping for smart antennas ............................................................ 24
   2.3 Making the most of spatial diversity: multi-antenna systems at transmitter
      and receiver (MIMO) ......................................................................................... 26
      2.3.1 MIMO: basic concepts ............................................................................ 26
      2.3.2 Narrowband and wideband characterization of the MIMO radio channel 28
      2.3.3 MIMO capacity ..................................................................................... 38
      2.3.4 MIMO signal processing: transceiver schemes and algorithms .......... 41
      2.3.5 MIMO channel measurements ............................................................. 45
      2.3.6 MIMO channel models .......................................................................... 48
      2.3.7 Antenna aspects in MIMO systems ....................................................... 54
      2.3.8 Prototyping for MIMO systems .............................................................. 60
      2.3.9 Summary on state of the art in MIMO: what is new, what is lacking? .... 64
3. ADAPTIVE ANTENNA FOR MOBILE SYSTEMS: design, implementation and evaluation of signal processing modules

3.1 Introduction and motivation ................................................................. 70
3.2 Beamforming and 3G, theoretical study ................................................. 73
  3.2.1 Introduction .................................................................................. 73
  3.2.2 Analysis and theoretical study of beamforming methods ............... 73
  3.2.3 Beamforming for 3G systems: advantages and drawbacks ........... 83
3.3 An adaptive antenna for 3G systems: general description and operation . 84
  3.3.1 Introduction .................................................................................. 84
  3.3.2 System operation: the “plug and play” concept ....................... 84
  3.3.3 General description and main features ........................................ 87
  3.3.4 Simulation results and expected performance .............................. 93
3.4 Signal processing modules: design and implementation for real-time operation .................................................................................. 96
  3.4.1 General architecture of signal processing modules ...................... 96
  3.4.2 Hardware platform ....................................................................... 98
  3.4.3 Signal processing modules for adaptive antenna ....................... 107
  3.4.4 Code optimization and load distribution ...................................... 120
  3.4.5 Analysis of impairments due to non-ideal effects and proposed solutions for their mitigation .......................................................... 124
3.5 Novel methodology to evaluate adaptive antennas ............................. 130
  3.5.1 Introduction .................................................................................. 130
  3.5.2 Evaluation under controlled scenario: 2-steps measurements in an anechoic chamber ................................................................. 130
  3.5.3 Evaluation under realistic scenario ............................................... 133
3.6 Set-up stage and modules to operate in a real 3G network ................. 135
  3.6.1 Need for set-up in the plug-and-play adaptive antenna .............. 136
  3.6.2 Call establishment in UMTS ......................................................... 136
  3.6.3 Proposed set-up for adaptive antenna in a UMTS network .......... 137
  3.6.4 Real-time DSP implementation .................................................... 139
  3.6.5 Test results of set-up module ...................................................... 141
3.7 General conclusions, contributions and further research on adaptive antennas for mobile system .......................................................... 143
4. RAPID PROTOTYPING FOR MIMO: design, implementation and measurements with a WLAN MIMO testbed........................................................................................................147

4.1 Introduction and motivation ................................................................. 149
4.2 Design and implementation of a MIMO testbed .......................... 150
  4.2.1 Design for multi-purpose scenarios and application ............... 150
  4.2.2 General architecture ............................................................... 152
  4.2.3 Modular description .............................................................. 154
  4.2.3.1 Antenna modules ............................................................. 154

4.3 MIMO channel measurements: Indoor and Outdoor-to-Indoor channel characterization including polarization effect ................................. 172
  4.3.1 Motivation and progress beyond the state of the art ............... 172
  4.3.2 Measurement set-up ............................................................. 172
  4.3.3 Approaches for capacity computation considering channel normalization ................................................................. 175
  4.3.4 Power computation and coverage maps ................................. 177
  4.3.5 Capacity results for indoor and outdoor-to-indoor scenarios ...... 179

4.4 General conclusions, contributions and further research on MIMO prototyping and dual-polarized MIMO channel measurements. ........ 182

5. PROPAGATION ASPECTS OF MIMO SYSTEMS: Analysis with multiple base station locations and multiple scenarios ........................................ 187

5.1 Introduction and motivation ................................................................. 190
5.2 Measurement system and considered scenarios ............................ 193
  5.2.1 MIMO testbed used in the measurement campaign ............... 193
  5.2.2 Data processing and channel estimation procedure ................ 197
  5.2.3 Measurement environment .................................................... 201

5.3 Evaluation of measurements and analysis of results: general channel characteristics and single-user analysis .............................................. 204
  5.3.1 General model ................................................................. 204
  5.3.2 Analysis of spatial correlation in MIMO systems for multiple scenarios ................................................................. 205
  5.3.3 Path loss, fading and coverage analysis ..................................... 211
  5.3.4 Capacity analysis and different methods of combining base stations .... 215

5.4 Multi-user-oriented evaluation of measurements and analysis of results 221
  5.4.1 Multi-user MIMO system model and sum capacity problem ........ 221
  5.4.2 Evaluation of multi-user MIMO at downlink: studied scenarios and capacity analysis ................................................................. 224
Table of Contents

5.4.3 Evaluation of multi-user MIMO at uplink: studied scenarios and capacity analysis .............................................................................................................228

5.5 General conclusions, contributions and further research on propagation aspects for MIMO systems .............................................................................................................232

6. ANTENNA ASPECTS OF MIMO SYSTEMS: design, implementation of antenna array and evaluation methods for MIMO .................................................................235

6.1 Design of a compact antenna array for MIMO systems: a solution for realistic user terminals .................................................................238

6.1.1 Design of the antenna element .................................................................................................................................................................238

6.1.2 Antenna array: low mutual coupling and compact design ..................................................................................................................249

6.1.3 Final array characteristics and prototype implementation ..................................................................................................................252

6.1.4 Reference antenna array .................................................................................................................................258

6.1.5 Conclusions .................................................................................................................................................................................262

6.2 Evaluation of antenna arrays for MIMO .................................................................264

6.2.1 Evaluation using measured radiation patterns ......................................................................................................................................264

6.2.2 Evaluation using measured radio channel responses ..................................................................................................................281

6.2.3 Evaluation with a reverberation chamber .................................................................................................................................289

6.2.4 Comparison of methods for evaluation of MIMO antenna arrays and conclusions ..................................................................................................................292

6.3 General conclusions, contributions and further research on antenna aspects for MIMO systems .............................................................................................................299

7. GENERAL CONCLUSIONS, FUTURE WORK AND PUBLICATIONS ............303

7.1 General conclusions and contributions ....................................................................................................................................................306

7.2 Future work ......................................................................................................................................................................................309

7.3 Publications ..................................................................................................................................................................................310

APPENDIX I: Antennas for measurements ................................................................................................................................................319

BIBLIOGRAPHY ........................................................................................................................................................................325
LIST OF FIGURES

RESUMEN EXTENDIDO - EXTENDED ABSTRACT IN SPANISH

Figura Resumen 1. Metodología general seguida durante la tesis y tipo de resultados obtenidos en cada paso. ................................................................. xxvii
Figura Resumen 2. Esquema modular del prototipo de antena adaptativa............................xxx
Figura Resumen 3. Prototipo de antena adaptativa desarrollado y evaluado ....................xxxiii
Figura Resumen 4. Esquema del demostrador MIMO propuesto ........................................xxxvii
Figura Resumen 5. Módulo receptor del sistema de medidas MIMO utilizado ....................xlii
Figura Resumen 6. Plano de planta donde se realizó la campaña de medidas, junto con los emplazamientos de transmisores estudiados ............................................................... xlii
Figura Resumen 7. Array compacto desarrollado, para sistema MIMO................................xlv
Figura Resumen 8. Array de referencia para comparación con array compacto....................xlv

CHAPTER 1

Figure 1-1. User data rate and scenarios for different wireless access techniques...........4
Figure 1-2. General methodology followed during thesis and type of result from each step. ...10

CHAPTER 2

Figure 2-1. Switched array vs adaptive array........................................................................18
Figure 2-2. Narrowband linear beamformer .........................................................................20
Figure 2-3. Smart antenna arrays base stations for WiMax, from ArrayComm [44] .............25
Figure 2-4. 12-element adaptive antenna, for wireless local loop system Super WLL, by Kyocera ................................................................................................................25
Figure 2-5. The MIMO concept ..........................................................................................26
Figure 2-6. The three aspects of benefit offered by MIMO....................................................27
Figure 2-7. MIMO techniques and channel propagation characteristics ............................28
Figure 2-8. Representation of Tx and Rx spatial correlation ..................................................35
Figure 2-9. Illustration of parallel subchannels for an NR NT MIMO system, where K is the rank of the MIMO channel.................................................................37
Figure 2-10. Water-filling scheme for CSI at Tx (optimum Tx power allocation) ..............39
Figure 2-11. Space-Time scheme (spatial diversity) .............................................................42
Figure 2-12. Example of beamforming for MIMO. Buildings are depicted in red, relevant paths in yellow and the optimum patterns in blue for 4-element antennas in Tx and Rx............43
Figure 2-13. Spatial Multiplexing architecture .....................................................................44
Figure 2-14. Conventional MIMO techniques ....................................................................45
Figure 2-15. Geometrically-based MIMO channel models with elliptical distribution of scatterers ..................................................................................................................51
Figure 2-16. Illustration of the one-ring and the two-ring models ........................................52
Figure 2-17. Scatterers distribution in the Distributed Scattering MIMO model..................52
Figure 2-18. Antenna representation in a MIMO system.....................................................55
Figure 2-19. Equivalent circuitual representation of the antenna in transmission and reception..56

CHAPTER 3
Figure 3-1. Schematic representation of simulated UMTS system to study NLSM and RLS algorithms.................................................................75
Figure 3-2. Iterative computation of beamforming weights for theoretical study of algorithms NLSM and RLS........................................................................76
Figure 3-3. Example of radiation pattern obtained from simulation in an UMTS scenario, with an adaptive antenna using NLMS and two values of step size μ .................................................................77
Figure 3-4. Evolution of one of the weights, for two different values of μ ........................78
Figure 3-5. Evolution of mean square error with NLMS algorithm, for two values of step size μ .................................................................78
Figure 3-6. Evolution of instantaneous SINR for adaptive antenna with NLMS algorithm, for two different values of step size .................................................................79
Figure 3-7. Mean square error for RLS, with two different values of forgetting factor, λ₀ .................................................................80
Figure 3-8. Comparison of NLMS and RLS performance in a UMTS scenario. Radiation pattern.................................................................80
Figure 3-9. Evolution of SINR obtained with adaptive antenna with NLMS and RLS adaptive algorithms........................................................................81
Figure 3-10. BER of data channel for UMTS with adaptive antenna, cases of Table 3-II ....82
Figure 3-11. Conventional structure of a digital adaptive antenna for W-CDMA systems ......85
Figure 3-12. General architecture of adaptive antenna (uplink), with the plug-and-play concept. ........................................................................86
Figure 3-13. General architecture of adaptive antenna (downlink), with the plug-and-play concept .................................................................86
Figure 3-14. General diagram of the adaptive antenna ..........................................................88
Figure 3-15. Antenna module of the adaptive array prototype .............................................89
Figure 3-16. Diagram of RF module, receiver ......................................................................90
Figure 3-17. Diagram of RF module, transmitter.................................................................90
Figure 3-18. Radiofrequency modules in the overall structure of the prototype..................91
Figure 3-19. RF-IF modules .............................................................................................91
Figure 3-20. Normalized array factor obtained with total interference cancellation and partial interference cancellation methods.................................................................94
Figure 3-21. Uplink SINR increase as a function of the number of users............................95
Figure 3-22. Downlink SINR increase seen by the mobile user .............................................95
Figure 3-23. SDR architectures ..........................................................................................97
Figure 3-24. General diagram of signal processing modules, uplink.....................................98
Figure 3-25. General diagram of signal processing modules, downlink...............................98
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>3-26</td>
<td>Technology situation for ADCs in terms of vertical resolution and sampling rate (taken at the moment of ADAM design)</td>
<td>100</td>
</tr>
<tr>
<td>3-27</td>
<td>The 6235 2-channel digital receiver plus A/Ds</td>
<td>102</td>
</tr>
<tr>
<td>3-28</td>
<td>The 6229 2-channel digital upconverter plus DACs</td>
<td>102</td>
</tr>
<tr>
<td>3-29</td>
<td>The 4292 4-DSP QUAD</td>
<td>104</td>
</tr>
<tr>
<td>3-30</td>
<td>General hardware structure</td>
<td>106</td>
</tr>
<tr>
<td>3-31</td>
<td>Hardware modules of ADAM prototype and test equipment</td>
<td>106</td>
</tr>
<tr>
<td>3-32</td>
<td>Adaptive antenna prototype</td>
<td>107</td>
</tr>
<tr>
<td>3-33</td>
<td>Block diagram of coarse synchronization</td>
<td>110</td>
</tr>
<tr>
<td>3-34</td>
<td>Block diagram of fine synchronization</td>
<td>111</td>
</tr>
<tr>
<td>3-35</td>
<td>Uplink demodulator diagram</td>
<td>112</td>
</tr>
<tr>
<td>3-36</td>
<td>Downlink demodulator and modulator diagrams</td>
<td>112</td>
</tr>
<tr>
<td>3-37</td>
<td>Diagram for computation of total weights with method aggregated weights</td>
<td>114</td>
</tr>
<tr>
<td>3-38</td>
<td>Diagram for computation of total weights with method aggregated reference</td>
<td>114</td>
</tr>
<tr>
<td>3-39</td>
<td>Beamformer structure: Uplink</td>
<td>116</td>
</tr>
<tr>
<td>3-40</td>
<td>Beamformer structure: Downlink</td>
<td>118</td>
</tr>
<tr>
<td>3-41</td>
<td>Diagram of A/D converter and IQ demodulator for one channel</td>
<td>119</td>
</tr>
<tr>
<td>3-42</td>
<td>Representation of spectrum of sampled signal for two possible sampling frequencies (upper figure: fs=61.44=16•R, lower figure: fs=30.72=8•R)</td>
<td>119</td>
</tr>
<tr>
<td>3-43</td>
<td>D/A converter and IQ modulator</td>
<td>120</td>
</tr>
<tr>
<td>3-44</td>
<td>Optimization steps flow diagram</td>
<td>122</td>
</tr>
<tr>
<td>3-45</td>
<td>Load distribution</td>
<td>124</td>
</tr>
<tr>
<td>3-46</td>
<td>Radiation patterns obtained by simulation of signal instability, with high (left) and low (right) variation</td>
<td>125</td>
</tr>
<tr>
<td>3-47</td>
<td>Calibration procedure used in the uplink for correcting amplitude and phase differences between the antenna elements</td>
<td>126</td>
</tr>
<tr>
<td>3-48</td>
<td>Method to compute the calibration matrix in downlink for correcting amplitude and phase differences between the antenna elements and RF-IF branches</td>
<td>127</td>
</tr>
<tr>
<td>3-49</td>
<td>Scheme of receiver channel: the frequency mismatch produces a phase error.</td>
<td>127</td>
</tr>
<tr>
<td>3-50</td>
<td>Demodulated DPCCH bits, with constant ∆f, with no interferences considered (left) and considering an interfering user (right)</td>
<td>128</td>
</tr>
<tr>
<td>3-51</td>
<td>Radiation pattern simulated for different quantized phase values in the phase track module</td>
<td>128</td>
</tr>
<tr>
<td>3-52</td>
<td>First step of procedure for evaluating adaptive antenna</td>
<td>131</td>
</tr>
<tr>
<td>3-53</td>
<td>Second step of procedure for evaluating adaptive antenna</td>
<td>132</td>
</tr>
<tr>
<td>3-54</td>
<td>Measured radiation pattern with interfering users in 18° and 10° azimuth angle</td>
<td>133</td>
</tr>
<tr>
<td>3-55</td>
<td>Measurement scenario for the realistic environment</td>
<td>133</td>
</tr>
<tr>
<td>3-56</td>
<td>Set-up module and relation with other signal processing modules</td>
<td>135</td>
</tr>
<tr>
<td>3-57</td>
<td>Simplified messages scheme in a UMTS call</td>
<td>136</td>
</tr>
<tr>
<td>3-58</td>
<td>Transport channel multiplexing structure</td>
<td>140</td>
</tr>
</tbody>
</table>
CHAPTER 4

Figure 4-1. Main scheme and modular architecture of the MIMO testbed.................................153
Figure 4-2. TX monopole array, to be used as one possible antenna module with variable spacing among elements.......................................................................................................................................156
Figure 4-3. Reflection coefficient for monopole arrays at TX (left) and RX (right). ............156
Figure 4-4. Coupling level for different spacing between antenna pairs in the monopole array. ...................................................................................................................................................................................157
Figure 4-5. Slant dual-polarized antennas based on crossed dipoles. ........................................158
Figure 4-6. S parameters of TX antennas..................................................................................158
Figure 4-7. S parameters of RX antennas .................................................................................159
Figure 4-8. Scheme of RF module at TX .................................................................................160
Figure 4-9. Photographs of the RF module at TX....................................................................160
Figure 4-10. Scheme of RF module at RX ..............................................................................161
Figure 4-11. Photographs of the RF module at RX.................................................................161
Figure 4-12. Scheme of signal processing module at TX, implemented over DSPs and FPGAs. ...................................................................................................................................................................................163
Figure 4-13. Scheme of signal processing module at RX, implemented over DSPs and FPGAs. ...................................................................................................................................................................................164
Figure 4-14. Main window of the client application in Matlab, for channel sounding........166
Figure 4-15. Window dialog for estimation of H matrix (Channel Sounder application)........167
Figure 4-16. Main window of the client application in Matlab, for MIMO algorithms testing.169
Figure 4-17. Example of used frame structure for algorithm testing ....................................170
Figure 4-18. Constellation of received signal before (upper figure) and after (lower figure) Alamouti demodulation.................................................................................................................................170
Figure 4-19. UMAT transmitter (Integration of TX modules)...................................................171
Figure 4-20. UMAT receiver (Integration of RX modules).....................................................171
Figure 4-21. Floormap of measured scenario. TX locations are shown in red (letters), RX routes are depicted in blue (numbers)...................................................................................................................................................................................175
Figure 4-22. Tx viewpoint for indoor (left) and outdoor (right) location for different antenna arrays. ...................................................................................................................................................................................175
Figure 4-23. Indoor TX location (A) for office-type environment .............................................176
Figure 4-24. Received power computed from the estimated H matrices (squared Frobenius Norm). ...................................................................................................................................................................................179
Figure 4-25. Capacity with instantaneous power normalization (fixed received SNR).........181
Figure 4-26. CDF of capacity for the four sets of measurements. The H matrix is normalized to a fixed value so the mean SNR is 20 dB for the 4 sets...................................................................................................................................................................................182
Figure 4-27. CDF of capacity for office-like scenario (Tx at location A, Rx at location 3), for different antenna modules. ...................................................................................................................................................................................182
LIST OF FIGURES

Figure 5-1. Illustration of hardware transmitter modules for the used narrowband testbed. ....193
Figure 5-2. Illustration of hardware receiver modules for the used narrowband testbed. .......193
Figure 5-3. Photograph of transmit modules. The 4 output channels were mounted together for
the outdoor location of base station. .......................................................................................194
Figure 5-4. Photograph of receiver module with 8 receive inputs. The system was mounted on a
trolley so it can be easily moved along indoor routes. ..........................................................195
Figure 5-5. Antennas used for the TX side, for indoor (left) and outdoor (right) BS location.
Although three 2-port antennas were available for the indoor location, only 2 were used (one for
each TX group). ..........................................................................................................................196
Figure 5-6. RX antenna on the trolley used to carry out the measurement campaign. ...............197
Figure 5-7. Frequencies used for each TX output (antenna) .....................................................198
Figure 5-8. Frequency allocation for the 4 sinewaves that were transmitted (one for each TX
output) and actual received signal, where frequency drift and the filter effect can be observed
...................................................................................................................................................198
Figure 5-9. Data processing to estimate H matrix from radio measurements ..........................200
Figure 5-10. Example of some elements of the estimated H matrix for one NLOS route ..........201
Figure 5-11. Floor plan (4th floor) and locations for base stations 1 and 2 for the 4 cases in the
measurement campaign. ...........................................................................................................202
Figure 5-12. Scenario seen from TX antennas for indoor (left) and outdoor (right) locations .203
Figure 5-13. CDF of the spatial power correlation coefficients for transmit antennas. 4 locations
are considered for the 2 base stations: indoor co-location (A), indoor medium spatial separation
(B), indoor opposite location with larger separation (C) and outdoor location (D). ...............207
Figure 5-14. CDF of the spatial power correlation coefficients for receiver antennas. 4 locations
are considered for the 2 base stations: indoor co-location (A), indoor medium spatial separation
(B), indoor opposite location with larger separation (C) and outdoor location (D). ...............209
Figure 5-15. CDF of spatial correlation coefficient for all the RX antenna and all the routes, as a
function of the TX location (case) ............................................................................................210
Figure 5-16. Path loss averaged over the 16 channel elements. Example of one route (NLOS,
case A) ......................................................................................................................................212
Figure 5-17. Overall fading and remaining slow fading after filtering with sliding window of
1m ..................................................................................................................................................213
Figure 5-18. Path loss averaged over each 1m, for cases A-D with different locations for base
stations ........................................................................................................................................214
Figure 5-19. Capacity maps including path loss in the H matrices. The transmit power is chosen
so that the average SNR = 10 dB for the whole floor, in the four studied cases ......................217
Figure 5-20. Capacity CDF for Option 1 (a fixed 2 × 4 system) and Option 2 (selection between
two 2 × 4 systems) ..................................................................................................................218
Figure 5-21. Capacity CDF for Option 3 (full 4 × 4 system) and Option 2 (selection between
two 2 × 4 systems) ..................................................................................................................219
Figure 5-22. Comparison of water-filling (Full CSI at TX) and no CSI at TX for a 4×4 MIMO
case at a local average SNR = 10dB. ......................................................................................219
Figure 5-23. Scenario for downlink MU-MIMO with intercell interference, for BS location case
1 ...................................................................................................................................................223
Figure 5-24. CDF of capacity for a multiuser MIMO system with two BSs, low SNR case (SNR=2dB, SIR=10dB). ........................................................................................................... 225

Figure 5-25. CDF of capacity for a multiuser MIMO system with two BSs, medium SNR and SIR case (SNR=SIR=10dB)......................................................................................................226

Figure 5-26. CDF of capacity for a multiuser MIMO system with two BSs, high SNR and SIR case (SNR=SIR=30dB). ............................................................................................................226

Figure 5-27. Example of case for uplink MU-MIMO with co-channel interference, for BS location case 1 ........................................................................................................................................... 227

Figure 5-28. CDF of single user capacity for desired user, without (SU) and with (MU) intracell interference........................................................................................................................................... 229

Figure 5-29. CDF of sum-capacity for uplink MIMO. 2 users in the system. SNR=10dB........230

CHAPTER 6

Figure 6-1. General structure of a PIFA....................................................................................239

Figure 6-2. Front view of PIFA with two substrates....................................................................240

Figure 6-3. Front view of a PIFA with two dielectric substrates and folded top patch..............241

Figure 6-4. Proposed structure of the element for a compact array for MIMO .........................242

Figure 6-5. S11 as a function of patch length, L2. ......................................................................244

Figure 6-6. S11 as a function of shorting pin position, L2..........................................................244

Figure 6-7. S11 for two different locations of element in the PDA mockup..............................245

Figure 6-8. The two studied position for the element on the selected area..............................245

Figure 6-9. Radiation patterns (azimuth and elevation planes) for the two studied element positions. The same dimensions and element characteristics were used for the two positions.245

Figure 6-10 S11 parameter and Smith Chart for the final element design. ..............................247

Figure 6-11. Radiation pattern representation for the final PIFA element (simulation results).247

Figure 6-12. Radiation pattern, cuts for θ=90º (left) and φ=90º (right). .....................................247

Figure 6-13. Final structure of the 1-element design (CST Microwave Studio view) ...............248

Figure 6-14. Main array configurations that were studied. The grey zone represents the 75 × 75 selected area for the elements placement. Big and small dots at the corner and side of each element show the position of feeding and shorting pins, respectively. ..............................250

Figure 6-15. S parameters for the 4 studied configurations. .....................................................251

Figure 6-16. Selection of higher coupling coefficients (zoom) for the working frequency (f=1766.6 MHz), for the four studied configurations.................................................................251

Figure 6-17. Structure of final designed array ..........................................................................252

Figure 6-18. Simulated reflection coefficient for final array....................................................253

Figure 6-19. 3D representation of radiation patterns (far field) for each element in the final array design. The radiation patterns are depicted over each respective element, for intuitive representation purposes. ........................................................................................................................................... 254

Figure 6-20. Simulated gain for final array design. ..................................................................254

Figure 6-21. Implemented compact antenna array...................................................................256
Figure 6-22. Reflection (left) and coupling (right) coefficients for the implemented prototype. Dashed lines represent the previous results (simulation) for the same design. ........................................256
Figure 6-23. Measured gain for the compact array .................................................................257
Figure 6-24. Simulated element for the reference array ..........................................................259
Figure 6-25. Photograph of the implemented reference array (4-monopole array) ...............260
Figure 6-26. Measured S parameters for the reference monopole array. ...............................260
Figure 6-27. Measured 3D radiation pattern for the reference monopole array .....................261
Figure 6-28. Measured gain (2-D) for the reference array .....................................................262
Figure 6-29. Variance as a function of incidence region in elevation, computed from the radiation patterns and eqs. ( 6-11) and ( 6-12). .................................................................268
Figure 6-30. Covariance as a function of incidence region in elevation, computed from the radiation patterns and eqs. ( 6-11) and ( 6-12). .................................................................268
Figure 6-31. Variance as a function of incidence region in elevation, computed from the radiation patterns and eqs. ( 6-11) and ( 6-13) .................................................................269
Figure 6-32. Covariance as a function of incidence region in azimuth, computed from the radiation patterns and eqs. ( 6-11) and ( 6-13) .................................................................269
Figure 6-33. Calculated MIMO average capacity using the Kronecker model with 4 uncorrelated antennas at the TX and the measured patterns of REF and AUT arrays at RX .................................270
Figure 6-34. CDF of capacity for SNR=10 dB, for the AUT and the reference array. Kronecker model and measured radiation patterns are used. .................................................................271
Figure 6-35. Average capacity for different SNR values, for the AUT and the reference array. Kronecker model and measured radiation patterns are used. .................................................................272
Figure 6-36. Geometrical description of Spatial Channel Model, from 3GPP-3GPP2 Group...273
Figure 6-37. CDF of capacity for the AUT and the reference array, obtained with 3GPP2 SCM and measured radiation patterns. AoDmax = 180º .................................................................279
Figure 6-38. Average capacity for different SNR values, for the AUT and the reference array. The 3GPP spatial channel model and measured radiation patterns are used. .................................................................280
Figure 6-39. Average capacity for different SNR values, for the AUT and the reference array. The 3GPP spatial channel model and measured radiation patterns are used. .................................................................281
Figure 6-40. The two arrays under study, mounted on a trolley (MS) ......................................282
Figure 6-41. Comparison of CDF of capacity for the two arrays with and without TX CSI using all the measurement routes. .................................................................................................284
Figure 6-42. Average capacity for different SNR values, for the AUT and the reference array. measured channel response is used. .................................................................................................284
Figure 6-43. CDF of eigenvalues (top) of the channel covariance matrix, HH'H, and eigenvalue dispersion (bottom), for all the measured routes .................................................................................................285
Figure 6-44. CDF of capacity obtained in the route on 4th floor, north corridor, for indoor (dashed line) and outdoor (solid line) TX location, for the two antenna arrays. The REF is shown in red and the AUT in blue. .................................................................286
Figure 6-45. Floorplan and route through the north corridor, 4th floor ............................... 287
Figure 6-46. Capacity as a function of local RX position, for AUT and reference array. Indoor TX location (LOS case) and SNR = 10 dB. 2D color map represents power ratio REF/AUT (in dB). .................................288
Figure 6-47. Capacity as a function of local RX position, for AUT and reference array. Outdoor TX location (NLOS case) and SNR = 10 dB. 2D color map represents power ratio REF/AUT (in dB)............................................................................................................................ 288

Figure 6-48. Drawing of a reverberation chamber and measurement scheme .......................... 290

Figure 6-49. Calculated capacity for a 3×4 MIMO system using the channel data from a reverberation chamber, and comparison with ideal i.i.d. 3×4 channel (no CSI at TX)............. 291

Figure 6-50. Studied arrays mounted in the reverberation chamber where measurements were gathered....................................................................................................................... 291

Figure 6-51. Comparison of methods to evaluate antenna arrays for MIMO systems. Capacity as a function of SNR, 4×4 MIMO. Blue has been used for the reference array, and red for the AUT. .................................................................................................................. 293

Figure 6-52. Comparison of methods to evaluate antenna arrays for MIMO systems. Capacity as a function of SNR, 3×4 MIMO. Blue has been used for the reference array, and red for the AUT. .................................................................................................................. 294

Figure 6-53. Comparison of required time to evaluate the two antenna arrays with the four proposed methods........................................................................................................ 297

CHAPTER 7

Figure 7-1. Topics investigated in this thesis and synopsis of main contributions. ................. 306
LIST OF TABLES

CHAPTER 2
Table 2-I. Main parameters in characterization of radio channels .............................................32
Table 2-II. MIMO measurement campaigns and their main characteristics ..............................47
Table 2-III. Characteristics of some relevant existing prototypes ..............................................64

CHAPTER 3
Table 3-I. Parameters used in simulation to study $\mu$ in NLMS ..................................................76
Table 3-II. Considered cases for study comparing NLMS and RLS ...........................................82
Table 3-III. Required number of operations for one loop (iteration) of NLMS and RLS ..........83
Table 3-IV. Beamforming of uplink physical channels .............................................................87
Table 3-V. Beamforming of downlink physical channels ...........................................................87
Table 3-VI. Main features of smart antenna ADAM .................................................................93
Table 3-VII. Main features of selected hardware for the adaptive antenna prototype ..........93
Table 3-VIII. Number of clock cycles and acquisition time for coarse synchronization algorithm ...........................................................109
Table 3-IX. Reduction of the number of clock cycles in the coarse synchronization module ...122
Table 3-X. Computational load of synchronization, demodulation and beamforming modules. ...........................................................123
Table 3-XI. Measured values for increase in gain and SIR when using the adaptive antenna ...134
Table 3-XII. Required set-up parameters in the adaptive antenna ..............................................138
Table 3-XIII. Maximum execution time per module ...............................................................141
Table 3-XIV. Acquired set-up values, for one test ...............................................................142

CHAPTER 4
Table 4-I. Summary of UMAT main features .................................................................172
Table 4-II. Summary of measurement setup for each of the six scenarios under analysis ....174

CHAPTER 5
Table 5-I. Baseband sinewave frequency for each TX ..........................................................198
Table 5-II. Main characteristics of measurement setup ........................................................203
Table 5-III. Mean values of spatial correlation at the transmitter (TX) .................................210
Table 5-IV. Mean values of spatial correlation at the receiver (RX) ......................................210
Table 5-V. Average path loss for all measured routes and the four BS locations (A to D) ....214
CHAPTER 6

Table 6-I. Dimensions of final element design .................................................................246
Table 6-II. Dimensions of final array design .................................................................253
Table 6-III. Simulated antenna efficiency for compact array ........................................258
Table 6-IV. Measured antenna efficiency for compact array ...........................................258
Table 6-V. Measured antenna efficiency for reference array ............................................262
Table 6-VI. Distribution function for AoD calculation in SCM simulations ..................277
Table 6-VII. Summary of parameters used for simulation with 3-GPP SCM ..................278
Table 6-VIII. Comparison of required time to evaluate two 4-element arrays for MIMO application, with different methods .................................................................296
Table 6-IX. Main characteristics of the 4 proposed methods to evaluate antenna arrays for MIMO ..................................................................................................................298
ACRONYMS

3G: Third Generation (of mobile communications)
ACF: autocorrelation function
ADC: Analog to Digital Converter
AoA: angle of arrival
AoD: angle of departure
AP: Access point
ASIC: application-specific integrated circuit
BLAST: Bell LAbs Space-Time scheme
BER: Bit Error Rate
BS: Base Station
CDF: Cumulative Distribution Function
CG: Conjugate Gradient
CSI: channel state information
DAC: Digital to Analog Converter
DoA: direction of arrival
DoD: direction of departure
DSP: Digital Signal Processor
FPGA: Field Programmable Gate Array
HSDPA: High Speed Downlink Packet Access
HSPA: High Speed Packet Access
HSUPA: High Speed Uplink Packet Access
i.i.d: independent and identically distributed
IF: intermediate frequency
ISI: inter symbol interference
LAN: Local Area Network
LMS: Least Mean Squares
LO: Local oscillator
LOS: Line Of Sight
MBER: Minimum Bit Error Rate
MIMO: Multiple Input Multiple Output
MISO: Multiple Input Single Output
MMSE: Minimum Mean Square Error
NLMS: Normalized LMS
NLOS: Non-Line Of Sight
OSTBC: Orthogonal Space-Time Block Codes
OVSF: Orthogonal variable spreading factor
PAS: power angular spectrum
PIFA: Planar Inverted-F antenna
PDP: power delay profile
QOSTBC: Quasi Orthogonal Space-Time Block Codes
RF: radio frequency
RX: receiver
RLS: Recursive Least Squares
RVM: Relevant Vector Machine
SIMO: Single Input Multiple Output
SM: Spatial Multiplexing
SMI: Simple Matrix Inversion
SNR: Signal to Noise Ratio
STBC: Space-Time Block Codes
STTC: Space-Time Trellis Codes
SVD: Singular Value Decomposition
SVM: Support Vector Machine
TCM: Trellis-coded Modulation
TX: transmitter
UE: User Equipment
UMTS: Universal Mobile Telecommunications System
VSWR: Voltage Standing Wave Ratio
WDCM: Wideband Directional Channel Model
WLAN: Wireless Local Area Network
SECTION I

RESUMEN AMPLIADO
(EXTENDED ABSTRACT IN SPANISH)
Resumen ampliado

- XX -
# TABLA DE CONTENIDOS DEL RESUMEN AMPLIADO

**I.1 Introducción y objetivos**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.1.1 Introducción al ámbito de la tesis</td>
<td>xxiii</td>
</tr>
<tr>
<td>I.1.2 Motivación</td>
<td>xxiii</td>
</tr>
<tr>
<td>I.1.3 Objetivos</td>
<td>xxiv</td>
</tr>
<tr>
<td>I.1.4 Resumen de principales aspectos novedosos de la tesis</td>
<td>xxv</td>
</tr>
<tr>
<td>I.1.5 Plan de desarrollo y metodología y seguidos</td>
<td>xxvi</td>
</tr>
<tr>
<td>I.1.6 Organización del documento de la tesis</td>
<td>xxvii</td>
</tr>
</tbody>
</table>

**I.2 Antena adaptativa para sistemas de comunicaciones móviles: diseño, implementación y evaluación de los módulos de procesado de señal**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.2.1 Introducción</td>
<td>xxix</td>
</tr>
<tr>
<td>I.2.2 Descripción general del sistema de antena adaptativa</td>
<td>xxix</td>
</tr>
<tr>
<td>I.2.3 Arquitectura hardware</td>
<td>xxxii</td>
</tr>
<tr>
<td>I.2.4 Evaluación de la antena adaptativa</td>
<td>xxxiii</td>
</tr>
<tr>
<td>I.2.5 Set-up para la antena adaptativa</td>
<td>xxxiv</td>
</tr>
<tr>
<td>I.2.6 Principales resultados</td>
<td>xxxiv</td>
</tr>
</tbody>
</table>

**I.3 Prototipado rápido para MIMO: diseño, implementación y medidas con un demostrador MIMO para WLAN**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.3.1 Introducción</td>
<td>xxxvi</td>
</tr>
<tr>
<td>I.3.2 Arquitectura y descripción modular del demostrador MIMO</td>
<td>xxxvi</td>
</tr>
<tr>
<td>I.3.3 Medidas de canal y resultados</td>
<td>xxxix</td>
</tr>
</tbody>
</table>

**I.4 Aspectos de propagación en sistemas MIMO: análisis con múltiples emplazamientos de estación base y múltiples escenario**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.4.1 Introducción</td>
<td>xl</td>
</tr>
<tr>
<td>I.4.2 Sistema y entornos de medida</td>
<td>xl</td>
</tr>
<tr>
<td>I.4.3 Medidas y resultados</td>
<td>xlii</td>
</tr>
</tbody>
</table>

**I.5 Aspectos de antena en sistemas MIMO: diseño, implementación de array de antenas y métodos de evaluación para MIMO**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.5.1 Introducción</td>
<td>xlv</td>
</tr>
<tr>
<td>I.5.2 Diseño de un array compacto para MIMO</td>
<td>xlv</td>
</tr>
<tr>
<td>I.5.3 Métodos de evaluación de arrays de antenas para MIMO</td>
<td>xlv</td>
</tr>
</tbody>
</table>

**I.6 Conclusiones y líneas futuras**

<table>
<thead>
<tr>
<th>Sección</th>
<th>Página</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.6.1 Conclusiones</td>
<td>xlvii</td>
</tr>
<tr>
<td>I.6.2 Líneas futuras</td>
<td>xlix</td>
</tr>
</tbody>
</table>


I.1 Introducción y objetivos

I.1.1 Introducción al ámbito de la tesis

El paradigma de la Sociedad de la Información y las Comunicaciones es una comunidad en la que los usuarios estén siempre on-line, puedan acceder a información de lo más heterogénea en el menor tiempo posible y puedan moverse libremente sin perder la conectividad aprovechando distintas tecnologías de comunicaciones según el entorno o el escenario. El primer paso hacia este ideal de comunicaciones “para todos, en todas partes” es el uso de las comunicaciones inalámbricas o wireless, que permiten al usuario cambiar de localización sin perder la comunicación. No sólo es necesario hacer realidad la tecnología, sino que ésta debe permitir ofrecer servicios atractivos para el usuario y a un coste asequible. Son muchos los esfuerzos de investigación y desarrollo a realizar.

Es un hecho que los nuevos servicios ofrecidos en comunicaciones inalámbricas requieren cada vez una mayor tasa binaria y velocidad de transmisión de información. Y no sólo se trata de mayor oferta, sino que a su vez la demanda de estos servicios de banda ancha (videoconferencia en el teléfono móvil, alta velocidad de transferencia a través de WiFi, televisión digital de alta definición…) es cada vez mayor. Este crecimiento supone un reto tecnológico, que ha llevado a la búsqueda de nuevas soluciones para mejorar la capacidad y eficiencia de las comunicaciones inalámbricas, tanto móviles (telefonía móvil) como itinerantes (WiFi, WiMax) o fijas (DVB-T, DVB-T2). Los sistemas de múltiples antenas o sistemas MIMO, son una de las apuestas de futuro con más fuerza en las futuras redes de nueva generación, junto con otras técnicas como la tecnología ultrawideband, o la multiplexación ortogonal en frecuencia OFDMA. Sin embargo, aún queda un importante esfuerzo científico y de investigación por hacer para llevar al usuario final las técnicas MIMO en su sentido más amplio, aunque el camino ya se ha empezado. La tesis que se presenta aquí contribuye a recorrer un trozo más de dicho camino.

El objetivo principal de esta tesis es aportar conceptos y resultados novedosos en aspectos de gran interés para impulsar el desarrollo de la tecnología de múltiples antenas y técnicas MIMO en sistemas de comunicaciones de nueva generación. En concreto, y puesto que los sistemas MIMO son un tema de estudio muy amplio, el trabajo se centra principalmente en aspectos de propagación, radiación y diseño de antena, junto a aspectos prácticos de implementación como el desarrollo de prototipos.

I.1.2 Motivación

Habiendo aprovechado al máximo el espectro de frecuencia y la multiplexación en tiempo y código, los sistemas de comunicaciones futuros se encuentran con el desafío de encontrar otro método de mejorar la eficiencia espectral. La solución que se está considerando en muchos casos es el uso de la diversidad espacial ofrecida por los sistemas con múltiples
antenas. Ya en sus primeras versiones, los estándares de 3GPP para UMTS consideraban el uso de diversidad espacial, aunque sólo en la estación base y con 2 antenas. Algunos trabajos de investigación y desarrollo han sido llevados a cabo tanto por el mundo científico como por el empresarial para estudiar y desarrollar prototipos de antenas inteligentes en los sistemas de telefonía móvil de tercera generación, si bien el esfuerzo aún no ha llevado su implantación masiva. En la actualidad, el interés por incluir técnicas MIMO en los nuevos estándares de comunicaciones inalámbricas es innegable, siendo WiMax, el estándard estadounidense para comunicaciones inalámbricas metropolitanas, el primer ejemplo ya en marcha de ello. Éste será previsiblemente seguido por WiFi (con el borrador de estándar 802.11n para redes inalámbricas locales), DVB-T2 para televisión digital terrestre de alta definición, y las nuevas versiones de telefonía celular de tercera generación europeas 3GPP. Para asegurar el éxito de las técnicas MIMO y con ello de estos sistemas de comunicaciones, son necesarios un gran número de estudios, simulaciones, prototipos y medidas. Esta tesis pretende contribuir a este esfuerzo investigador.

I.1.3 Objetivos

El objeto principal de la tesis es aportar aspectos novedosos al estudio y caracterización de los sistemas de múltiples antenas en transmisión y recepción, en especial desde una perspectiva de propagación, antena e implementación real. El mismo se concreta en varios objetivos a cumplir durante el desarrollo de la tesis, que se resumen a continuación:

· OBJETIVO 1 (tema: antenas adaptativas para despliegues en redes reales): Desarrollo de los módulos de procesado de un prototipo de antena inteligente para comunicaciones móviles de tercera generación, basada en el concepto “plug and play”. Diseño de la nueva metodología de medida y caracterización del prototipo, así como la implementación de los módulos necesarios para su integración con una red real UMTS.

· OBJETIVO 2 (tema: prototipado rápido y sistema de medidas MIMO para múltiples aplicaciones): Implementación de un prototipo MIMO para caracterización de canal, prueba rápida de algoritmos e inclusión de efectos de antena. Se pretende con este objetivo contar con una herramienta que permita caracterizar el canal MIMO, y en concreto estudiar el canal MIMO teniendo en cuenta múltiples polarizaciones, así como facilitar a los desarrolladores de algoritmos MIMO la prueba y evaluación de los mismos incluyendo un canal real y efectos de implementación real.

· OBJETIVO 3 (tema: aspectos de propagación y configuración de estaciones base para MIMO): Caracterización del canal MIMO para un entorno de interiores considerando antenas con múltiples polarizaciones, e incluyendo varios posibles escenarios, como indoor-indoor e indoor-outdoor, tipo oficina y tipo pasillo, con y sin línea de visión directa, etc. Además, se realiza el análisis de varios emplazamientos de estaciones base (BS) para MIMO mediante la realización de campañas de medida, considerando aspectos de correlación entre BS y capacidad.
Se incluye el estudio de varias posibilidades en la combinación de la señal en las estaciones base y evaluación de los resultados considerando varias opciones en cuanto a la información disponible en transmisor y receptor, así como el análisis de la capacidad considerando no sólo un usuario en el sistema, sino múltiples usuarios y el efecto de las interferencias.

· OBJETIVO 4 (tema: arrays de antenas para un mundo MIMO realista): Diseño e implementación de un array de antenas compacto y realista, para un dispositivo tipo de futuras aplicaciones MIMO, como las PDAs. Investigación y propuesta de nuevos métodos de medida para caracterizar las antenas desde el novedoso punto de vista de sistemas MIMO.

I.1.4 Resumen de principales aspectos novedosos de la tesis

Desde la publicación de los artículos pioneros en técnicas MIMO de Foschini y Telatar, en 1996 y 1999 respectivamente, el interés por estos sistemas ha crecido exponencialmente, llevando a un incremento sustancial en el número de trabajos de investigación y de aspectos relacionados con sistemas multiantena. Si bien las bases teóricas han sido ampliamente tratadas, las implementaciones realistas son escasas, y existen aspectos importantes olvidados o tratados muy por encima, hasta ahora.

En cuanto a las antenas inteligentes, su base teórica es bien conocida, y existen algunas implementaciones reales de las mismas. Sin embargo, convencionalmente se ha supuesto su diseño en conjunto con la estación base, lo que limita la flexibilidad en su implantación. En este proyecto de tesis se propone un nuevo concepto de antena adaptativa “plug and play”, junto con su implementación real.

Por otro lado, los sistemas MIMO han sido típicamente estudiados desde el punto de vista de sistema: algoritmos, teoría de la información…Sin embargo, aspectos como el efecto de la antena como elemento real o las características de propagación en canales MIMO y su efecto en los bloques de procesado son temas no totalmente claros. En esta tesis se han realizado varios prototipos MIMO para caracterizar varios aspectos del canal tanto en banda ancha como en banda estrecha, facilitando a la vez la prueba de algoritmos en entornos realistas de tipo indoor e indoor–outdoor. También se estudia el efecto de la antena mediante la implementación de un array compacto adecuado para un dispositivo realista (tipo PDA), proponiéndose métodos de caracterización y medida de antenas para MIMO. Gracias a dichas implementaciones, se presentan varios resultados de caracterización y medidas considerando aspectos realistas, como efectos no deseados debidos a la implementación con elementos reales, efectos del canal, etc.

Los resultados obtenidos durante la tesis han dado lugar a una serie de artículos publicados en revistas científico-técnicas de conocido renombre, así como a una gran cantidad de artículos presentados en conferencias nacionales e internacionales. En concreto, los resultados se concretan en:

· 4 artículos publicados en revistas científico-técnicas.
- 21 artículos presentados en conferencias internacionales.
- 10 artículos presentados en conferencias nacionales.

Además, el trabajo realizado durante la tesis ha dado lugar a una patente internacional. También se han dado tres conferencias invitadas en congresos internacionales, en las que se han presentado algunos de los resultados de esta tesis. Por último, cabe decir que en el marco de esta tesis la doctoranda ha dirigido dos proyectos fin de carrera.

I.1.5 Plan de desarrollo y metodología y seguidos

La tesis se ha desarrollado siguiendo 4 líneas principales de trabajo de investigación en las que se han presentado contribuciones. Previo a dicho trabajo de investigación, se realizó una labor de estudio pormenorizado del estado del arte en cada uno de los temas trabajados, que ha dado lugar a un capítulo (Capítulo 2 de la tesis) de documentación, estudio de estado del arte y evaluación de aspectos novedosos aún por investigar o resolver. Además, no sólo se estudió el estado del arte al comienzo de la tesis, sino que se ha continuado realizando un seguimiento del mismo a lo largo de la tesis.

Los bloques o líneas de trabajo son los siguientes:

- Diseño, realización y evaluación de los módulos de procesado de una antena adaptativa para UMTS y análisis de efectos de implementación real.
- Realización de un demostrador MIMO como plataforma para estudio de configuraciones de antenas y aspectos de propagación y su efecto en algoritmos, y estudio del canal MIMO con múltiples polarizaciones en entornos indoor e indoor-outdoor.
- Caracterización de canal MIMO en entornos de interiores incluyendo múltiples emplazamientos de estación base multiantena y diversas configuraciones de transmisión, así como el efecto de la interferencia en sistemas multiusuario MIMO.
- Diseño y evaluación de antenas realistas para sistemas MIMO en entornos de interiores.

En general, en cada una de estas líneas de trabajo se ha seguido la misma metodología, con algunas pequeñas variaciones en función del trabajo que se desarrolló en cada una:

- Realización de un estudio del arte del área de investigación a tratar.
- Definición del problema no resuelto y los aspectos sobre el que se va a trabajar.
- Diseño del sistema o prototipo que se va a implementar para evaluar o analizar los aspectos objetivo del tema a tratar (algoritmos, modelos, arquitectura...).
- Implementación del sistema o prototipo a medir o a utilizar para la evaluación y estudio.
- Realización de las medidas y pruebas correspondientes.
· Evaluación de los datos obtenidos: procesamiento de los datos en bruto y evaluación de los mismos, obteniendo las conclusiones del estudio y proponiendo nuevas líneas de acción para el futuro.

La Figura Resumen 1 representa esquemáticamente el esquema seguido.

**Figura Resumen 1. Metodología general seguida durante la tesis y tipo de resultados obtenidos en cada paso.**

### I.1.6 Organización del documento de la tesis

Esta tesis se ha organizado en 7 capítulos:

· El _capítulo 1_ realiza una introducción al ámbito de la tesis, presenta los objetivos de la misma y comenta la metodología seguida durante la tesis. Por último, realiza un repaso a la organización del documento de la tesis.

· El _capítulo 2_ introduce los principales aspectos teóricos sobre antenas inteligentes y sistemas MIMO, y ofrece un detallado estudio del arte sobre ambos temas. Además, se subrayan los principales puntos no resueltos en ambas áreas de trabajo.

· El _capítulo 3_ presenta el diseño, la implementación y la evaluación de los módulos de procesado de señal de una antena adaptativa para comunicaciones móviles. Se incluyen aspectos varios en cuanto a la implementación en una plataforma de software-radio, como restricciones de funcionamiento en tiempo real en el diseño de los módulos de procesado, o el compromiso entre complejidad y resultados. Además, se explica un nuevo método para medir y evaluar la
antena adaptativa en escenarios controlados y también realistas, y se comentan los resultados obtenidos para el prototipo de antena adaptativa implementado.

· El capítulo 4 describe la implementación de un demostrador MIMO para prueba de algoritmos, medida de configuraciones de antenas y caracterización de canales MIMO en escenario de interiores. Se presentan las diversas opciones que permite el demostrador: múltiples tipos de elementos de antena, tasa binaria y algoritmos de transmisión-recepción MIMO. Se muestran ejemplos de funcionamiento y medidas obtenidos con el mismo.

· El capítulo 5 analiza varios aspectos de los sistemas MIMO desde un punto de vista de propagación. Se comparan resultados de sistemas MIMO en escenarios indoor y outdoor-indoor, usando datos obtenidos a partir de una campaña de medidas. También se analizan diferentes emplazamientos para los transmisores (actuando como estaciones base). Finalmente se incluye en el capítulo un estudio sobre el efecto de las interferencias en el sistema MIMO, incluyendo un análisis con múltiples usuarios basado en las medidas de canal realizadas.

· El capítulo 6 se centra en aspectos de antena para sistemas MIMO. Se estudian las posibles opciones de arrays MIMO realistas y se presenta el diseño de un array compacto novedoso, que es adecuado para su uso en terminales de usuario. También se presentan aquí algunos métodos para evaluar arrays de antenas para sistemas MIMO. Estos son comparados mostrando los resultados obtenidos al evaluar el array compacto previamente presentado, de modo que se muestren claramente las ventajas e inconvenientes de cada método.

· Finalmente, el capítulo 7 resume los resultados obtenidos y las conclusiones más importantes, y presenta los trabajos y líneas de acción futuras propuestas.

Este resumen presenta un esquema similar, con la salvedad de no incluir un apartado con un resumen del estado del arte (se considera más interesante consultar el propio estudio del arte realizado en el capítulo 3). Los apartados que resumen los capítulos con carga técnica (del 3 al 6 incluidos) resumen de forma muy escueta el trabajo realizado, pero no pueden incluir todos los resultados obtenidos por razones de brevedad. Se remite a los lectores interesados a que consulten los capítulos correspondientes para mayor información.
I.2 Antena adaptativa para sistemas de comunicaciones móviles: diseño, implementación y evaluación de los módulos de procesado de señal

I.2.1 Introducción

Las antenas inteligentes se presentan como una opción de gran interés dentro del campo de las comunicaciones móviles. Bajo el término antena inteligente se agrupan varios tipos de antenas, desde las antenas de haces conmutados hasta las antenas adaptativas, siendo en este último tipo en el que se ha centrado el trabajo en esta tesis.

Los beneficios de las antenas adaptativas en los sistemas de comunicación móviles, y en concreto en los sistemas celulares basados en CDMA como UMTS son innegables. Los de mayor relevancia son la mejora en la relación señal a interferencia, y por tanto el incremento de la capacidad del sistema, aumentando con ello el número de usuarios permitidos en el sistema. La coexistencia de UMTS y GSM hace de gran interés la implementación de antenas adaptativas multi-estándar.

A pesar de los beneficios que ofrecen las antenas adaptativas, éstas no han sido adoptadas de manera amplia en el mercado. Una de las principales razones es el alto coste de implantación y mantenimiento, ya que típicamente estas antenas realizan el procesado a nivel de bit y esto supone que deban ser específicas para cada estación base de fabricante. Por ello resulta de gran interés una solución que permita la independencia entre la antena adaptativa y la estación base, mediante la utilización de interfaces estandarizados como es el interfaz radio Uu.

En esta tesis se presenta una antena adaptativa realizada siguiendo el anterior concepto, que se ha denominado antena “plug-and-play”. El trabajo de la tesis se centró en los módulos de procesado de la antena adaptativa, así como en los métodos de evaluación a utilizar. Se resume a continuación de forma muy breve el trabajo llevado a cabo. Más información puede consultarse en el capítulo 3 de la tesis.

I.2.2 Descripción general del sistema de antena adaptativa

Funcionamiento y esquema modular general

La antena adaptativa que se presenta se divide en dos módulos bien diferenciados: el módulo de radiofrecuencia, en el que se trata la señal de RF y FI, y el módulo de procesado de señal. En este último se realiza todo el procesado necesario para la conformación de la señal, partiendo de la señal de frecuencia intermedia ofrecida por el módulo de radiofrecuencia correspondiente. La Figura Resumen 2 muestra un esquema simplificado de la antena adaptativa desarrollada. En la arquitectura completa se distingue entre la arquitectura modular para el enlace ascendente y la correspondiente al enlace descendente. El trabajo de la tesis se ha centrado en los módulos de procesado, por lo que no se detallan los módulos de RF.
La antena adaptativa desarrollada se sitúa entre la estación base del operador y los equipos de usuarios móviles, ocupando por tanto el lugar ocupado actualmente por las antenas convencionales. El sistema propuesto consta de cuatro elementos en el array.

Para facilitar su rápido despliegue y la independencia de la estación base, la antena adaptativa se basa en un diseño que se ha denominado de “plug and play”. La señal recibida se procesa de manera que la comunicación con la estación base utiliza el interfaz Uu. Así, la señal recibida en uplink se pasa de radiofrecuencia (RF) a frecuencia intermedia (FI) y ahí se digitaliza. Los módulos de procesado la convierten a banda base, y tras su procesado la señal ya conformada es de nuevo convertida a analógica y trasladada a radiofrecuencia para ser enviada a la estación base. El proceso inverso es realizado en downlink. De este modo, el funcionamiento de la antena adaptativa es transparente para la estación base y el terminal móvil.

La conformación se ha basado en referencia temporal, utilizando la información del canal dedicado de control del usuario. Sólo se ha considerado el funcionamiento de la antena en estado conectado y para la Release 99 de 3GPP, dejándose su posible evolución para el funcionamiento con canales de datos no dedicados y con HSPA como posible línea futura de trabajo.

**Módulos de procesado**

El procesado de señal o software radio se ha dividido en varios módulos, que se pueden agrupar en dos grandes bloques: módulos de sincronismo y MODEM, y módulos de conformación.

- Los **módulos de sincronismo y MODEM** de la antena adaptativa se encargan de llevar a cabo las etapas de sincronización de la señal recibida para poder demodular adecuadamente los datos de usuario necesarios para la conformación, así como de realizar la modulación y demodulación cuando es necesario. Para la aplicación deseada de este sistema, en el enlace ascendente se realiza únicamente la demodulación del canal dedicado de control (DPCCH) y en el enlace descendente se realiza la modulación y la demodulación de los canales dedicados de datos (DPDCH) y control (DPCCH). Independientemente de la técnica de espectro ensanchado que
estemos utilizando, es imprescindible la información temporal de la trama transmitida para poder demodular la señal recibida. El bloque de sincronismo se ha dividido en dos bloques: el de adquisición y el de seguimiento. El bloque adquisición o sincronización gruesa permite lograr que la señal recibida y el código de scrambling que la demodula tengan sólo un pequeño desfase relativo, menor que un chip. La función del bloque de seguimiento o sincronismo fino es obtener una alineación más precisa y mantener la sincronización durante toda la recepción, para lo que se aprovechan el sobremuestreo que se realiza (ya que la señal se muestrea a cuatro muestras por chip).

- Los módulos de conformación de la antena adaptativa tienen como objetivo tratar la señal recibida, de modo que la señal obtenida a la salida del mismo presente una mejor relación señal a interferencia C/I. Para ello se conforma la señal multiplicándola por ciertos pesos, que maximizan la potencia de la señal en la dirección del usuario deseado y minimiza la potencia en la dirección de las señales interferentes que puedan aparecer en el sistema. El diagrama de radiación cambia de forma adaptativa, variando los pesos para obtener los valores que optimizan la relación C/I. De este modo se logra seguir en tiempo real tanto las variaciones de los usuarios deseados como de las posibles interferencias que pueden afectar al sistema y que se desean reducir. Para realizar la conformación, se han estudiado teóricamente diversos algoritmos adaptativos, tanto convencionales como novedosos, y se han seleccionado dos de ellos para ser estudiados en detalle para el caso específico de conformación para una antena adaptativa en un sistema UMTS: el algoritmo NLMS (normalized least mean squares) y el algoritmo RLS (recursive least squares). Se han analizado los resultados obtenidos con ambos algoritmos para diversos números de usuarios e interferencias, así como sus requisitos computacionales y su capacidad de adaptación a los cambios en el entorno y el canal. Finalmente, se ha seleccionado el algoritmo NLMS para su implementación en el prototipo dada su menor carga computacional y sus resultados finales casi tan buenos como los de RLS, aunque presenta una velocidad de convergencia menor. Un aspecto de interés es que para reducir la carga computacional, y teniendo en cuenta que la señal ofrecida a la estación base o nodo B debe ser una única señal (y no una señal por usuario), se realiza la conformación para todos los usuarios deseados de forma simultánea, obteniéndose unos pesos globales que conforman toda la señal recibida. Esto equivale a realizar una cancelación parcial de interferencias en Uplink.

Para el enlace descendente se han analizado varias posibilidades para calcular los pesos a partir de los pesos obtenidos en el enlace ascendente. Puesto que las frecuencias de ambos enlaces son distintas, los canales observados no serán iguales, pero las direcciones de llegada de las señales de los usuarios sí. Por ello una opción es utilizar unos pesos corregidos en fase para tener en cuenta las diferencias de frecuencia. Para reducir la carga computacional, se ha analizado también el resultado cuando se utilizan los mismos pesos en ambos enlaces, obteniéndose resultados aceptables para el caso considerado (sólo 4 elementos de array). Esta fue la solución final adoptada. En cuanto a la conformación, ésta sí se realiza a nivel de bit en
downlink, por lo que la cancelación sigue el esquema que hemos denominado “cancelación total”, en contraposición del de cancelación parcial. Como resultado, la conformación es más sencilla y requiere mucha menos carga computacional, pero se necesita tanto modular como remodular los canales de datos de usuario (cuando en uplink sólo era necesaria la demodulación del canal de control de cada usuario, sin modulación posterior).

I.2.3 Arquitectura hardware

La implementación y desarrollo del software radio se ha llevado a cabo en procesadores digitales comerciales. Esta implementación del procesado de la señal en DSPs permite realizar actualizaciones y mejoras del sistema con gran sencillez, dando una gran flexibilidad al sistema. La selección de los procesadores ha tenido en cuenta que el sistema de antena adaptativa debe trabajar en tiempo real. Puesto que la señal UMTS es de banda ancha y se trabaja con una tasa de señal relativamente alta (3.84 Mcps) la capacidad de procesado para implementar el software radio debe ser grande. Por ello se han utilizado varios procesadores digitales de gran capacidad de cálculo, de última generación en el mercado cuando se realizó el desarrollo, trabajando en paralelo. Se han elegido procesadores comerciales de aritmética de punto fijo, ya que ésta permite obtener una mayor velocidad de cálculo que la aritmética de coma flotante. En concreto, para la implementación práctica se han utilizado los procesadores de Texas Instruments TMS320C6203, que trabajan con un reloj a 300 MHz y permiten obtener hasta 2400 millones de instrucciones por segundo (MIPS). Se ha trabajado con placas de Pentek (QUADs) con cuatro de estos procesadores en paralelo, para obtener una mayor velocidad de cálculo. Junto a estas placas se han incluido cinco receptores digitales (incluyendo conversores analógico-digitales) y cinco transmisores digitales (que incluyen conversores digital-analógico) que reciben y transmiten las señales ofrecidas por los módulos de RF-FI. Puesto que se contaba con varios procesadores trabajando en paralelo, fue necesario realizar un reparto de tareas entre estos, asociando los diversos módulos a distintas tarjetas QUADs y distintos DSPs dentro de las mismas, tratando de optimizar al máximo la capacidad total del sistema. También el código obtenido fue optimizado para tratar de aprovechar al máximo la capacidad de los procesadores, ya que una mayor capacidad permite trabajar con un mayor número de usuarios. Tras las medidas de carga de los distintos módulos se observa que se pueden conformar simultáneamente hasta 3 usuarios con el sistema hardware presentado.

Se ha trabajado con aritmética de punto fijo. Esto permite aumentar la velocidad de cálculo para procesos lineales. Para obtener un código más optimizado ha sido necesario convertir los algoritmos utilizados a punto fijo, lo que lleva a utilizar un algoritmo de adaptación LMS en punto fijo.

La Figura Resumen 3 muestra una fotografía de los módulos de antena (izquierda) y procesado (derecha) del prototipo de antena inteligente desarrollado.
I.2.4 Evaluación de la antena adaptativa

Tras la implementación del sistema de antena inteligente, se hace necesario evaluar su funcionamiento. Como parte de esta tesis se estudia la medida de la misma en entornos controlados y realistas. Puesto que el diagrama de radiación de una antena adaptativa varía en función de las características del entorno, los métodos de medida de antena convencionales no son aplicables a este tipo de antenas, por lo que se hace necesario desarrollar nuevos métodos. En la tesis se ha propuesto un método de medida basado en 2 pasos: en un primer paso se calculan los pesos de conformación en la antena adaptativa mientras que en un segundo se realiza la medida del diagrama de radiación con el módulo de cálculo de pesos desconectado. Se realiza la caracterización de la antena adaptativa en diversos entornos: entorno controlado (cámara anecóica) y entorno realista (outdoor).

Además, se incluye en el análisis de las medidas los efectos debidos a aspectos no ideales de la implementación real. El hecho de realizar un prototipo real permite caracterizar aspectos que generalmente no se incluyen en simulación, y que sin embargo aparecen en prototipos reales, debido a no idealidades del sistema. En concreto, los efectos más importantes que se han estudiado son los siguientes efectos:

- Diferencias de fase y amplitud entre las cadenas de radiofrecuencia, y pequeñas inestabilidades en las mismas. Se ha estudiado su efecto, principalmente en el diagrama de radiación obtenido, y se ha propuesto el uso de un módulo de calibración para mitigar su efecto.

- Errores debidos a las diferencias de frecuencia entre transmisor y receptor. Se estudió el error que introduce el hecho de trabajar con distintos cristales en transmisor y receptor para generar las frecuencias de mezcla y el reloj de los conversores de muestreo, efecto muy normal en un sistema real. Se ha presentado un módulo de recuperación y
Resumen ampliado

seguimiento de frecuencia que ha sido optimizado para tener en cuenta que en el sistema habrá no sólo usuarios deseados sino también interferencias. Además, se reduce la carga computacional al máximo y se tiene en cuenta que la implementación debe ser óptima para aritmética de punto fijo.

I.2.5 Set-up para la antena adaptativa

Para permitir que la antena adaptativa funcione de manera independiente respecto de la estación base a la que se conecte, se hace necesario realizar un módulo que en esta tesis hemos denominado de set-up. Dicho módulo es capaz de reconocer que un nuevo usuario ha entrado en el sistema y obtiene los datos asociados a dicho usuario necesarios para que la antena adaptativa lo conforme adecuadamente. Este módulo, si bien conlleva poca carga computacional para el sistema, sí es de una gran complejidad conceptual, debido a los muchos niveles y mensajes a tener en cuenta.

Para el funcionamiento transparente de la antena adaptativa se requiere conocer ciertos datos especificados por la red cuando se establece una portadora radio (“Radio Bearer”), como por ejemplo el código de ensanchado utilizado por el usuario, el formato de trama, etc. Para ello, en esta tesis se ha propuesto e implementado un módulo de set-up simplificado para su funcionamiento en redes UMTS típicas. Consta de cuatro pasos:

1. Obtención del canal físico secundario S-CCPCH del interfaz radio, a partir de los datos indicados en las recomendaciones de 3GPP.
2. Lectura de la cabecera de nivel MAC, que contiene información acerca del usuario y del canal lógico que transporta.
3. Lectura de la cabecera de nivel RLC, que recoge un número de secuencia (para el correcto reensamblado de los segmentos) y varios campos que indican la longitud de los datos de nivel superior.
4. Entrega de los paquetes que componen el canal DCCH al nivel RRC y decodificación de la información de dichos paquetes, que se encuentran condificados en lenguaje ASN-1.

Para comprobar su funcionamiento, este módulo fue probado con terminales de prueba de UMTS junto con la antena adaptativa, mostrando su correcto funcionamiento para detectar la entrada de nuevos usuarios y sus parámetros asociados, permitiendo así el correcto funcionamiento de la antena adaptativa.

I.2.6 Principales resultados

Como resultado de este bloque de la tesis, se presentan varios tipos de resultados. A continuación se resumen los más importantes.

En primer lugar, se ha hecho un estudio teórico de la mejora que se obtiene con el uso de la antena adaptativa utilizando los dos algoritmos seleccionados, es decir, NLMS y RLS. Se
han llevado a cabo numerosas simulaciones a nivel de enlace, con distinto número de usuarios deseados e interferentes. Se presentan gráficas de BER (Bit Error Rate) en función del número de usuario y del algoritmo utilizado, y se demuestra que la antena adaptativa ofrece una importante mejora comparada con una antena sectorial convencional. Los valores exactos se pueden consultar en el capítulo 3, si bien cabe decir que con la antena adaptativa de 4 elementos se logran mejoras de BER de hasta 10 veces menor que con la antena sectorial. En cuanto a la comparación entre ambos algoritmos adaptativos, si bien RLS logra mejores resultados en general que LMS, las diferencias son mínimas, funcionando incluso mejor NLMS en escenarios de alta movilidad del usuario que RLS. Sin embargo, el incremento en complejidad de RLS es significativo, requiriéndose hasta 20 veces más operaciones que con NLMS para un array de 4 elementos.

A lo largo de la tesis, también se presentan resultados de carga computacional en los DSPs. Estos son importantes en lo referido a la implementación real, ya que permiten asegurar el funcionamiento en tiempo real de la antena adaptativa, y muestran qué modulos suponen la mayor carga. Además, es necesario para realizar un correcto reparto de tareas en el sistema de procesado paralelo utilizado. Se puede destacar que el módulo de conformación en uplink es uno de los que implican mayor carga computacional, ya que ha de realizar la conformación a nivel de chip. También los módulos de sincronismo suponen una importante carga.

Por último se presentan los resultados de las medidas del sistema de antena adaptativa. Se presentan tanto medidas de diagrama de radiación en cámara anecoica como medidas de mejora de ganancia y de SIR (Signal to Interference Ratio) en entornos realistas. También se incluyen pruebas del módulo de set-up en una red de operado de telefonía real. Con todo ello se muestra la mejora que ofrece la antena adaptativa y que el método de medida en dos pasos para antena adaptativa es adecuado.
I.3 Prototipado rápido para MIMO: diseño, implementación y medidas con un demostrador MIMO para WLAN

I.3.1 Introducción

Tras el estudio y evaluación de una antena adaptativa, que supone diversidad espacial en sólo uno de los extremos del enlace, se aborda en la tesis el estudio de sistemas con múltiples antenas en transmisión y recepción. Para llevar a cabo un estudio realista, es necesario contar con un sistema que permita realizar medidas y pruebas en entornos reales. Por ello, y partiendo de la experiencia obtenida durante la tarea anterior, se ha realizado un demostrador MIMO, cuyas principales características son la flexibilidad y versatilidad, de modo que se facilite la caracterización de diversos aspectos de propagación en sistemas MIMO, así como el estudio y evaluación de técnicas MIMO considerando varias configuraciones de antena.

Existen un gran número de prototipos y demostradores testbed en la literatura, cada uno de ellos diseñado con un propósito específico: channel sounders para realizar medidas de canal precisas, prototipos para evaluar los resultados obtenidos con un determinado algoritmo, etc. El demostrador presentado aquí no representa un sistema completamente novedoso y rompedor en el ámbito de los demostradores MIMO, sino que su novedad radica en la multifuncionalidad que permite obtener y en su diseño con un triple propósito:

- Obtención de medidas de canal MIMO para su caracterización en entornos indoor y outdoor-indoor, con capacidad de fácil desplazamiento por rutas indoor.
- Evaluación sencilla y rápida de algoritmos y esquemas MIMO incluyendo canales reales que se pueden variar según el escenario en que se quiera probar el algoritmo.
- Caracterización de diversas configuraciones de array de antenas para MIMO, con varios módulos de antena implementados (con una o doble polarización, y posibilidad de variar fácilmente la separación entre antenas, además de permitir incluir nuevos módulos futuros).

El demostrador es especialmente adecuado para propósitos de investigación y educacionales, ya que permite realizar múltiples pruebas y medidas de características de los sistemas MIMO de forma sencilla.

I.3.2 Arquitectura y descripción modular del demostrador MIMO

El demostrador propuesto consta de varios módulos, que se resumen en la Figura Resumen 4. Se ha propuesto un esquema de funcionamiento off-line para los algoritmos y esquemas que se deseen probar o medir, mientras que el envío y recepción de datos se realiza on-line; de este modo se simplifica en gran medida la implementación, puesto que no es necesario considerar restricciones de tiempo real durante el procesado de datos, que puede por
tanto ser más complejo. Pero a la vez, el esquema anterior permite incluir efectos reales del canal y considerar la realización de los módulos más sencillos en tiempo real.

![Esquema del demostrador MIMO propuesto.](image)

El testbed ha sido implementado para su funcionamiento en la banda inferior de WLAN (centrada en 2.45 GHz), ya que uno de sus objetivos es la caracterización de canales MIMO indoor. Además, la banda inferior de WLAN también ha sido considerada recientemente para nuevas versiones de 3G, por lo que la caracterización y estudio del canal MIMO en esta frecuencia se hace aún más interesante. El demostrador se ha realizado para un número de antenas de hasta 4 en transmisión y 4 en recepción (MIMO de hasta $4 \times 4$), con la posibilidad de variar la configuración de antenas utilizadas.

Se pueden identificar varios módulos que constituyen el testbed MIMO implementado: módulos de pre- y post-procesado, módulos de RF-FI y módulos de antena. Además, un PC de control en transmisor y otro en recepción permiten controlar el intercambio de información y revisar el estado del sistema.

**Módulos de procesado digital**

En la parte del transmisor, el procesado offline consiste en un PC donde se generan en Matlab las señales a enviar. El procesado online se encarga de recibir dichas señales, convertirlas de digital a analógico y enviarlas a las cadenas de RF por las cuatro salidas correspondientes en el mismo instante de tiempo. Para ello, se utiliza una plataforma Software-Radio similar a la utilizada para el desarrollo de la antena adaptativa presentada anteriormente. Se basa en una placa DSP Quad Pentek 4292 (TMS320C6203) y dos placas upconverters digitales donde se implementan las cuatro cadenas transmisoras digitales. En el lado correspondiente a la recepción, las cuatro cadenas receptoras consisten en otra placa DSP Quad Pentek 4292 (TMS320C6203) y dos placas digitales downconverters.
Módulos de RF-FI

Los conversores superior e inferior generan digitalmente la frecuencia intermedia seleccionada de 40 MHz. La frecuencia de muestreo para los conversores digital-analógico (DACs) y analógico-digital ADCs) provienen de los osciladores de cristal internos, utilizándose la misma señal para todas las cadenas en cada lado del enlace, si bien siendo distintos en transmisión que en recepción. De este modo se incluye en el demostrador el efecto del posible error de frecuencia entre transmisor y receptor, que puede aparecer en sistemas reales. Por otro lado, los osciladores locales utilizados para la conversión de frecuencia (IF a RF) se obtienen de generadores de señal comerciales y sus frecuencias se eligen para conseguir una frecuencia de RF de 2.45 GHz. De este modo se pretende que el sistema sea lo más realista posible, ya que en un sistema de comunicaciones real el transmisor y el receptor poseen distinta referencia de frecuencia. En transmisión y recepción la etapa de RF está formada de cuatro cadenas realizadas mediante componentes comerciales. La potencia de transmisión máxima es de +25 dBm, con lo que el demostrador es aplicable a escenarios picocelda y microcelda.

Módulos de antena

Se han implementado dos módulos de la antena para caracterizar los sistemas MIMO teniendo en cuenta dos posibilidades en polarización: antenas con una única polarización y antenas con doble polarización.

El primer tipo de módulo de antena consiste en un array de monopolos $\lambda/4$ sobre un plano de masa con varias posibles posiciones de conector de antena, para soportar distintas configuraciones de antena en base al espaciado entre los elementos. Las distancias posibles para varían entre 0.1 y 1 veces la longitud de onda.

El segundo tipo de módulo de antena consiste en un array de dos elementos formados cada uno por dos dipolos cruzados, con polarización lineal de ±45°. Los dos elementos están separados una distancia fija de $\lambda/2$.

La versatilidad del testbed desarrollado permite la fácil integración de otras configuraciones y otros tipos de elementos.

Interfaz de usuario y post-procesado

Se ha desarrollado un interfaz de usuario fácil de utilizar e intuitivo, para trabajar con el demostrador tanto como medidor de canal como de probador de algoritmos. El interfaz está basado en Matlab, y permite representar gráficamente algunos de los resultados medidos, así como facilitar al usuario la corrección del error de frecuencia y el cálculo de las matrices de canal de forma automática. Se ha definido una estructura de la señal a enviar al demostrador, de modo que se facilite la comunicación de posibles programas del usuario con algoritmos MIMO y el demostrador.
I.3.3 Medidas de canal y resultados

Como aplicación del demostrador desarrollado, y también para cubrir uno de los objetivos planteados en la tesis, se ha realizado una campaña de medidas con el demostrador, en entornos indoor y outdoor-indoor y comparando las medidas obtenidas con ambos módulos de antena (con polarización simple y con doble polarización).

Atendiendo al análisis de propagación para canales MIMO multipolarizados, se han realizado medidas en diferentes escenarios. El primero consiste en un escenario de propagación tipo pasillo. Aunque la mayoría de los trabajos previos se han centrado en escenarios tipo oficina, el escenario pasillo es también importante, especialmente para comunicaciones inalámbricas de baja movilidad dentro de los edificios (usuario caminando por el pasillo). Se consideraron dos diferentes posiciones para el transmisor, para tener en cuenta casos de interiores (indoor) y exteriores (outdoor). En segundo lugar, se hicieron varias medidas también en entorno de oficinas. Todas las medidas fueron llevadas a cabo teniendo en cuenta las dos configuraciones de array: monoplos y diplos cruzados.

Como primer resultado, se presenta en la tesis un análisis del efecto del método de normalización de canal que se utilice para calcular la capacidad disponible en el sistema. Se muestra que la normalización típicamente utilizada (normalización según la potencia instantánea recibida en promedio en el array) no siempre es la más adecuada, especialmente cuando se quiere comparar las prestaciones de distintos arrays, ya que dicha normalización elimina la posible mejora por ganancia que pueda introducir cada array. Por otro lado, la normalización a un cierto valor promedio para un determinado escenario permite estudiar la variabilidad de la potencia recibida, así como la cobertura ofrecida para un cierto valor de potencia transmitida.

También se ofrecen resultados de cobertura, es decir, mapas de potencia media recibida. Se muestra que la potencia es mucho menos variable en entornos outdoor-indoor que en entornos indoor, ya que se evitan los cambios de zonas con visión directa y sin visión directa.

Por último, la campaña de medidas realizada permitió comparar los resultados obtenidos en un sistema donde todas las antenas utilizan la misma polarización (sistema con diversidad espacial únicamente) con un sistema donde se utilizan antenas con doble polarización cruzada (incluye diversidad espacial y de polarización). Tras analizar las medidas, se muestra que la capacidad obtenida es mayor cuando se utilizan antenas con doble polarización en el entorno tipo pasillo y oficinas medido. En conclusión, el sistema con doble polarización ofrece mejores prestaciones para el escenario tipo pasillo medido. Este resultado es razonable, ya que el entorno medido presenta poca riqueza en diversidad espacial (pocos dispersores) y por tanto el uso de diversidad en polarización incrementa la descorrelación entre antenas y por tanto mejora la capacidad para el caso estudiado de no CSI.

Más resultados interesantes obtenidos a partir de estas medidas se presentan en el documento de la tesis.
I.4 Aspectos de propagación en sistemas MIMO: análisis con múltiples emplazamientos de estación base y múltiples escenario

I.4.1 Introducción

La selección del mejor emplazamiento para situar una estación base para sistemas MIMO es un aspecto de gran importancia desde el punto de vista del operador. A pesar de las muchas campañas de medidas MIMO presentadas en la literatura el estudio de cuáles son las características y los parámetros a tener en cuenta para seleccionar el emplazamiento de las estaciones base no ha sido ampliamente considerado.

Por otro lado, la mayoría de las campañas de medida previas que se encuentran en la literatura consideran sistemas MIMO con un módulo de transmisores y otro de receptores, por lo tanto en el sistema se supone una única estación base (BS). Sin embargo, algunos estudios teóricos han considerado también la posibilidad de que las estaciones base cooperen entre sí, por lo que el análisis de medidas para este caso también resulta interesante.

Por ello hemos llevado a cabo otra campaña de medidas, diferente a la presentada en la sección anterior, en la que se ha puesto especial énfasis en evaluar diferentes emplazamientos de estación base, comparar entornos indoor con entornos outdoor-indoor para un número de puntos de medida considerable (no principalmente en pasillos como fue el caso de los ejemplos de la sección anterior) y también analizar diversos métodos de combinación y cooperación de estaciones base.

I.4.2 Sistema y entornos de medida

Se ha llevado a cabo una campaña de medidas en el campus de la universidad KTH, en Estocolmo. Tras ser procesadas, las medidas obtenidas se han utilizado para evaluar diversos parámetros y características de los entornos medidos. A continuación se presenta brevemente el sistema MIMO que se utilizó para obtener las medidas de canal, así como los entornos de medida considerados.

Sistema de medida

Para obtener las medidas de canal posteriormente utilizadas en el análisis, se utilizó un demostrador MIMO con 4 cadenas transmisoras y hasta 8 cadenas receptoras. Para obtener los resultados presentados en esta sección se utilizó un sistema 4×4. Se trata de un sistema de banda estrecha basado en el uso de procesadores digitales para el procesado de señal. El transmisor se consideró como estación base o punto de acceso (estático) mientras que el receptor o terminal de usuario representó el lado móvil del enlace.

La frecuencia portadora utilizada para las medidas fue 1766.6 MHz. Las cadenas de RF se construyeron a partir de componentes de propósito general conectorizables. Con la intención
de estudiar diversas configuraciones de estaciones base en el sistema, las cadenas transmisoras se agrupan en bloques de dos, estando cada una de ellas controlada por un ordenador. De este modo, se trabaja con dos grupos o estaciones base de 2 transmisores cada uno. Los módulos de recepción se montaron sobre una plataforma móvil, que se alimentó mediante una batería de 12 V. Así, se permitió incluir movilidad en el receptor para poder desplazarlo por varias rutas de interiores en diferentes plantas del edificio. En cuanto al módulo de antena, se utilizaron distintos tipos de elementos para transmisión y recepción. En transmisión, se utilizaron dos antenas planas comerciales con doble polarización (polarización a \( \pm 45^\circ \)), de Huber-Suhner, para las medidas realizadas con emplazamientos de interiores. Para el emplazamiento en exteriores se utilizaron dos arrays comerciales para estación base, de banda ancha y también doble polarización a \( \pm 45^\circ \), con mayor ganancia y haz más directivo. En recepción, el módulo de antena utilizado para las medidas analizadas en este artículo fue un array de monoplos separados una distancia \( d=\lambda/2 \).

**Entornos medidos**

Las medidas consideran varias rutas del receptor, siempre en entornos de interior, dentro del edificio de oficinas y laboratorios. Con el fin de analizar diversos entornos de aplicación del sistema MIMO con varias estaciones base, la campaña de medidas realizada abarcó varios escenarios, haciendo especial hincapié en los posibles emplazamientos y configuraciones de las estaciones base o grupos transmisores. Así, se han dividido las medidas realizadas en 4 casos o escenarios en función de la posición de los transmisores:

- Los 4 transmisores (ambas estaciones base) se sitúan al final de la planta analizada del edificio, lado este, por tanto sin separación espacial relevante.

- Ambas estaciones base se sitúan en el lado este del edificio, pero al final de distintos pasillos. Se incluye por tanto separación espacial entre estaciones base.

- Cada estación base se situó en lados opuestos de la planta del edificio (separación entre estaciones base máxima para emplazamientos *indoor* en la planta).

- Los 4 transmisores (ambas estaciones base) se sitúan en la azotea del edificio colindante al edificio estudiado, apuntando hacia el mismo.

Para los 4 casos de emplazamiento de los transmisores, los receptores siguieron varias rutas en el interior del edificio, que incluyen entornos de oficinas, pasillos y laboratorios. Las medidas se tomaron a velocidad peatonal (aproximadamente 1 m/s). Se tomaron medidas de canal en 3 plantas diferentes, aunque las tres de características similares.

En la Figura Resumen 5 se presenta una foto del receptor MIMO, mientras que en la Figura Resumen 6 se presenta el plano de una de las plantas medidas y las posiciones de los transmisores para cada uno de los casos considerados.

- xli -
I.4.3 Medidas y resultados

A partir de las medidas obtenidas en la campaña realizada, se han estudiado diversos parámetros y se han analizado varias características del sistema MIMO desde un punto de vista de propagación. A continuación se resumen los resultados obtenidos.

En primer lugar, se ha evaluado la correlación espacial obtenida entre antenas de un mismo lado del enlace (transmisor o receptor) para todos los puntos medidos. Desde un punto de vista de sistema MIMO, las propiedades de correlación espacial del canal son de primordial importancia. Se ha mostrado que las matrices de correlación espacial de transmisión y recepción se pueden usar para estimar la matriz de correlación \( R \) del canal en determinados casos. Ésta, a su vez, ofrece información sobre la diversidad espacial y la capacidad que se puede obtener en el entorno MIMO estudiado. Tras calcular el coeficiente de correlación para cada punto medido y cada pareja de antenas, se presentaron los resultados en forma de función de probabilidad acumulada (CDF), para cada emplazamiento medido. Como cabía esperar, la correlación es mucho menor para los casos con BSs separadas. Como resultado de mayor interés, cabe destacar que para el caso outdoor sin separación entre BSs se obtiene una correlación espacial algo menor que para el caso equivalente indoor, lo que puede ser debido a la menor separación entre antenas. Sin embargo, es especialmente interesante notar que en el caso outdoor la reducción en correlación obtenida al usar distinta polarización es más importante que la obtenida debido a la separación espacial entre arrays.

En segundo lugar, se ha realizado un estudio de las pérdidas de propagación, del desvanecimiento de señal y de los mapas de cobertura obtenidos. Principalmente se observa que el emplazamiento en exteriores para la estación base implica unas pérdidas de propagación mucho mayores (de unos 30dB respecto al escenario indoor), como cabe esperar, ya que la señal debe atravesar la pared del edificio.

También se ha llevado a cabo un análisis de la capacidad del sistema para distintas combinaciones de señales en las estaciones base y distintos supuestos para el CSI en...
transmisión. Se han obtenido diversas conclusiones interesantes, pero quizá se puede destacar la que muestra que un esquema MIMO 2×4 donde el grupo de antenas transmisoras (BS) se selecciona usando un algoritmo sencillo de mayor potencia recibida, logra obtener unas prestaciones (en cuanto a capacidad) casi tan buenas como un esquema MIMO 4×4. Esto es muy interesante, ya que el uso del sistema 2×4 reduciría significativamente la complejidad del sistema.

Finalmente, se han aprovechado las medidas anteriores para analizar el efecto de las interferencias (tanto de usuario intracelda como de señales intercelda). Para ello se han tomado algunas de las medidas realizadas secuencialmente como datos tomados “virtualmente a la vez”. Esta suposición es válida en la medida en que todos los puntos de medida fueron repetidos de la misma manera para cada emplazamiento de estación base. Como resultados, se presenta la capacidad con y sin interferencia tanto para uplink como para downlink. Los resultados y conclusiones se pueden consultar en el cuerpo de la tesis.
I.5 Aspectos de antena en sistemas MIMO: diseño, implementación de array de antenas y métodos de evaluación para MIMO

I.5.1 Introducción

Uno de los problemas con los que se encuentran las técnicas MIMO para ser implementadas en sistemas móviles inalámbricos es (entre otros) precisamente que requieren que haya múltiples antenas no sólo en la estación base o punto de acceso, sino también en el terminal de usuario. Esto implica una serie de requisitos de espacio, peso, etc, que hace que el diseño del sistema MIMO sea complejo y deba incluir al array de antena como parte a optimizar también.

Como tarea final de la tesis se ha complementado el estudio del canal MIMO mediante la caracterización y medida del efecto de antenas realistas en el sistema en comparación con el uso de un array de referencia. Para ello se ha llevado a cabo el diseño e implementación de un array de antena compacto.

Los parámetros convencionales de antena (impedancia de entrada, diagrama de radiación, pérdidas de retorno…) no son generalmente la mejor manera de caracterizar su comportamiento en sistemas MIMO. Por ello uno de los puntos de mayor interés en esta tesis es la propuesta de diversos métodos de evaluación de antenas para MIMO y la comparación de sus resultados.

I.5.2 Diseño de un array compacto para MIMO

En la tesis se ha diseñado y posteriormente implementado un array de elementos especialmente pensado para su posible integración en un equipo de usuario realista para ello se han tenido en cuenta las posibles dimensiones de un dispositivo real con aplicación en futuros sistemas MIMO, como puede ser una agenda electrónica personal PDA. Se propone el uso de elementos tipo PIFA, que se caracterizan por su compactos y de fácil fabricación. La posición de los elementos en el array se ha seleccionado para buscar el menor acoplamiento entre elementos posible. En la Figura Resumen 7 se muestra el prototipo desarrollado. Inicialmente, éste se ha caracterizado mediante los parámetros típicos de caracterización de antenas (Sij, diagrama de radiación, eficiencia de radiación…). Las características de este array se comparan con una antena de referencia, que consiste en un array de monoplos separados $\lambda/2$ (Figura Resumen 8).
I.5.3 Métodos de evaluación de arrays de antenas para MIMO

Si bien los parámetros convencionales de antena también son de interés en sistemas MIMO y deben optimizarse, éstos no ofrecen suficiente información (o al menos no directamente) para saber si un array de antenas presenta buenas prestaciones en este tipo de sistemas. Esto se debe a que los sistemas MIMO se basan en propiedades como la diversidad espacial, la riqueza multitrayecto, etc, y por lo tanto no basta con conocer las características típicas de propagación, como la ganancia en una cierta dirección, ya que se esperan rayos procedentes de múltiples direcciones.

Por ello, en esta tesis se ha hecho un esfuerzo por analizar y proponer posibles métodos para evaluar arrays de antenas desde el punto de vista de sistemas MIMO. En concreto, hemos considerado cuatro métodos distintos:

- Evaluación de la capacidad utilizando la covarianza de la antena (obtenida a partir de los diagramas de radiación medidos) y el modelo de Kronecker.
- Evaluación de la capacidad utilizando los diagramas de radiación medidos y el modelo de canal estandarizado por la 3GPP.
- Evaluación utilizando medidas de canal tomadas con un demostrador MIMO en entornos de interior.
- Evaluación mediante medidas en cámara reverberante.

Para cada método, se han analizado los resultados en términos de capacidad disponible, y finalmente se han comparado los cuatro métodos en cuanto a la calidad y precisión de los resultados, el detalle y el tipo de información que se puede obtener y finalmente la complejidad de la evaluación (en cuanto a tiempo necesario y requisitos HW y SW).

Se han obtenido resultados muy interesantes sobre los métodos de evaluación y su comparación, que se pueden consultar en detalle en la tesis. Aquí destacamos que si bien la evaluación mediante medidas de canal en un entorno real con un testbed es el método que
requiere mayor esfuerzo (de tiempo y de recursos), también es el que más se ajusta a la realidad. Por tanto es el más interesante para estudiar emplazamientos específicos, aunque puede no ser el más óptimo para estudios más generales. En cuanto a los demás métodos, cabe decir que el primero es el más sencillo y ofrece resultados bastante cercanos a los obtenidos mediante medida real, si bien implica hacer una suposición sobre la distribución de las señales en el espacio. Las medidas en cámara reverberante son un límite superior de la capacidad que se obtendrá con el array de antenas, ya que asumen que el canal MIMO es ideal (incorrelación total entre elementos, gran riqueza en multitrayectos). Finalmente, el canal de la 3GPP es en general más pesimista que las medidas reales, porque sólo tiene en cuenta componentes en 2 dimensiones.
I.6 Conclusiones y líneas futuras

A continuación se resumen las conclusiones y líneas futuras de la tesis. En el documento se puede encontrar una explicación más detallada de ellas, junto con la justificación de las mismas.

I.6.1 Conclusiones

El trabajo desarrollado en esta tesis ha sido dividido en cuatro bloques o áreas principales, que se corresponde con los resultados presentados entre los capítulos 3 y 6, y de acuerdo con las principales líneas de investigación y asuntos en desarrollo para sistemas multi-antena identificados en el capítulo 2. Para cada bloque se presentan resultados y conclusiones particularizadas al final de cada capítulo. En este resumen se presentan brevemente las ideas y conclusiones principales obtenidas, si bien se pueden consultar de forma más detallada en cada capítulo correspondiente de la tesis.

La mayor parte de los resultados presentados en esta tesis han sido obtenidos mediante medidas y evaluaciones de sistemas reales y de prototipos. También se han llevado a cabo nuevas campañas de medida con el fin de analizar las características de propagación. El uso de datos completamente reales obtenidos de medidas en lugar de usar únicamente simulaciones significan un paso adelante en la obtención de resultados, incluyendo otros aspectos como los efectos de los sistemas no ideales.

Como conclusión general de la tesis, los sistemas multiantena que han sido mostrados ofrecen unas mejoras significativas en los sistemas inalámbricos, especialmente en lo que se refiere a la reducción de interferencia y a la capacidad o incremento de la tasa binaria total ofrecida por el sistema. Además, a lo largo de la tesis se han presentado una serie de problemas que las implementaciones reales deberían contemplar cuando se lleven estos sistemas a sistemas y equipos de usuarios reales. En esta tesis se analizan algunos de estos problemas y también las posibles soluciones para mitigar estos efectos no deseados, encontrándonos que, incluso incluyen efectos no ideales, las antenas adaptativas y los sistemas MIMO presentan mejoras muy interesantes.

En esta tesis se ha demostrado que un prototipo en tiempo real de una antena adaptativa es viable. Se ha diseñado e implementado, y a lo largo de este documento se mencionan algunos problemas debidas a los efectos no ideales en el prototipo real. Aparte de estos efectos, hemos mostrado que la antena adaptativa de cuatro elementos implementada puede ofrecer un incremento en la ganancia de hasta 6dB para operaciones optimas, y se aprecia una reducción de interferencias de aproximadamente 15 dB. Todos los resultados fueron obtenidos a partir de medidas en escenarios controlados y reales. Además, hemos estudiado en detalle las posibilidades de integrar de forma transparente esta antena adaptativa en una red UMTS real y
hemos propuesto y desarrollado un módulo de set-up, analizando su funcionamiento en la primera versión de una red 3G real.

Las antenas adaptativas proporcionarán interesantes resultados en entornos donde el número de multitrayectos es pequeño, como en áreas rurales, o donde las interferencias son fuertes y con dependencia espacial, como en el caso de los hot spots. Sin embargo, cuando los entornos inalámbrico son ricos en multitrayectos con diferencias entre retardos muy pequeñas, donde hay un gran número de elementos dispersores y generalmente no hay línea de visión directa (como en interiores o en áreas urbanas), el procesado MIMO es preferible para sacar provecho de la diversidad espacial. Con el fin de evaluar varios aspectos de los sistemas MIMO hemos desarrollado un nuevo demostrador MIMO, basado en el paradigma de prototipado rápido y el concepto de software radio. En la tesis se muestran sus posibilidades para realizar medidas de canal MIMO, para evaluar diferentes configuraciones de array y para probar distintos algoritmos multiantena. Se han obtenido varias conclusiones de interés a partir de las campañas de medidas llevadas a cabo con este testbed MIMO, como la comparación de resultados en términos de capacidad y pérdidas de propagación para sistemas MIMO con una sola polarización o doble polarización para entornos de interiores.

Se presentan también otros resultados de propagación a partir de otra campaña de medidas, haciendo hincapié en este caso en la comparación entre un escenario de interiores y un escenario interior-exterior. Según los resultados obtenidos, estos últimos ofrecen una diversidad en polarización y espacial mayor y deberían simplificar el desarrollo de las estaciones base para cubrir edificios y otros escenarios de interiores, sin embargo hay que tener en cuenta que requieren transmitir más potencia (hasta 30 dB en los escenarios medidos) incluso si se utilizan antenas directivas (en elevación) en la estación base. Se muestran también otras conclusiones interesantes en relación al emplazamiento de la estación base y la combinación de señales tanto para transmisores sin información de canal (CSI: Channel State Information) como para transmisores con CSI perfecto. Estos resultados se pueden utilizar como recomendaciones para el despliegue de redes inalámbricas en interiores incluyendo esquemas MIMO.

Finalmente, para terminar de cubrir las tres principales partes en el sistema de comunicaciones multiantena estudiado (estación base, canal y equipo de usuario), también hemos investigado el terminal de usuario. La tendencia actual de aumentar el tamaño de los teléfonos móviles (para, por ejemplo, tener mayores pantallas) y de añadir nuevas funcionalidades que requieren mayor velocidad de transmisión (conexiones Wifí, video-conferencias de alta calidad…), hace necesario el diseño y evaluación de nuevos arrays de antenas compactos para dispositivos personales. En esta tesis se presenta un novedoso diseño para un array multielemento con cuatro antenas que puede ser implementado en una PDA real, obteniéndose buenos resultados para los parámetros convencionales de medidas de antenas. Además, se muestra cómo los parámetros convencionales de antenas como los diagramas de...
radiación pueden ser utilizados para calcular otros parámetros de gran interés para evaluar el rendimiento de la antena en sistemas MIMO, como la capacidad bajo determinadas suposiciones. También se presentan y comparan otros métodos que no están directamente basados en parámetros convencionales. Como conclusión de los resultados de la tesis, y en contra de lo que algunas veces se ha asumido, los escenarios de interiores no pueden ser considerados tan ricos en multitrayectos como los canales i.i.d. ideales, y por eso la capacidad medida en cámaras reverberantes para evaluar antenas para MIMO será, generalmente, una aproximación optimista. Por otro lado, los canales MIMO estandarizados a partir de 3GPP sólo consideran canales 2-D, por tanto ignoran las contribuciones de la diversidad de los diagramas de radiación y canales en 3-D. De este modo, los resultados obtenidos a partir de un modelo de canal que incluye un diagrama de radiación 2-D será, generalmente, una aproximación pesimista. Más conclusiones interesantes sobre la comparación de los métodos estudiados para evaluar arrays en sistemas MIMO se pueden consultar en la memoria.

I.6.2 Líneas futuras

Los sistemas multiantena para comunicaciones inalámbricas representan un tema muy amplio con muchas áreas de investigación por cubrir y muchos aspectos sin resolver completamente. A pesar de que algunos de ellos han sido abordados en esta tesis, todavía hay mucho trabajo por hacer. Aquí resumimos brevemente algunas de las líneas más interesantes que deben ser investigadas en el futuro y que están más relacionadas con el trabajo presentado.

Respecto a las antenas adaptativas basadas en conformación, en esta tesis se presenta un prototipo y se evalúa su rendimiento en algunos escenarios. Sin embargo, sería necesaria la validación de las pruebas en múltiples situaciones y con muchos usuarios para evaluar el rendimiento de las antenas adaptativas a nivel de sistema. Además, nuestra solución no considera el cifrado en niveles superiores, que es el habitual el modo de funcionamiento de las redes UMTS actuales. Por lo tanto, las antenas adaptativas comerciales deberían tener en cuenta este aspecto, definiendo un protocolo de nivel superior para entregar la información cifrada necesaria a la antena adaptativa. Aparte de esto, algunos trabajos previos ya han evaluado antenas adaptativas a nivel de sistema en comunicaciones móviles, pero las medidas en redes reales con múltiples antenas adaptativas son escasas.

En cuanto a los sistemas MIMO, son interesantes más medidas de canal y un análisis detallado para comprender mejor el funcionamiento del canal MIMO y para modelar sus características. Algunos ejemplos de aspectos sin resolver sobre esta área son el modelado de grupos de dispersores, la investigación de canales que incluyan polarizaciones en 3D (con tres polarizaciones) o métodos eficientes para modelar el ancho de banda de los canales.

En la tesis analizamos la capacidad disponible para dos supuestos: sin CSI en el transmisor y con CSI perfecto en el transmisor. Sin embargo, en aplicaciones reales es muy probable que sólo se disponga de CSI parcial en el transmisor. Aunque algunos trabajos previos...
se han hecho sobre este asunto, los nuevos algoritmos para CSI parcial son muy interesantes para mejorar el rendimiento de los sistemas en el mundo real. Esta es claramente una línea futura de investigación de gran interés.

En esta tesis se profundiza en las nuevas antenas especialmente diseñadas para sistemas MIMO y la evaluación de las mismas. Sin embargo, hay todavía mucho trabajo que hacer relacionado con este tema. Por ejemplo, son necesarios otros tipos de antenas para dispositivos pequeños de usuario. El trabajo conjunto con fabricantes de equipos será necesario para optimizar el array de antenas incluyendo el efecto del terminal y para ver el comportamiento de terminales multiantena en el mundo real.
SECTION II

MAIN DOCUMENT
INTRODUCTION

1.1 Motivation ....................................................................................................... 4

1.2 Objectives and general methodology of the thesis................................. 8

1.3 Outline of the thesis ...................................................................................... 11
The use of multiple antenna technologies in communication systems has recently been proposed as an interesting solution to improve system performance and increase data rate and overall system throughput. By using the redundancy or diversity that can be obtained when several antennas are used, the aim is to increase bit rate, to minimize interference level and to reduce bit error rate. Previous theoretical works have shown the remarkable advantages of using multi-antenna technology, in only one end of the link or both link ends (transmitter and receiver). As a result, a significant increase in the interest on these technologies has been observed in the research community.

Despite the great research effort that has being carried out, there are still quite a few open issues that should be addressed before the multi-antenna technologies can be considered a mature technology and fully ready for the market, as we will see along this document. This thesis has been performed with the idea of contributing to some of these aspects, with special emphasis to those related with propagation, antenna issues and real implementation.
1.1 Motivation

Recently, novel applications with high data rate requirements have been introduced in mobile communication systems, such as video telephony or real-time video streaming. Due to the expected increase in using these applications among the general public, also a rise in user data rate and throughput will be required in these communication systems.

Moreover, the use of wireless equipments in indoor environments such as those of local area networks (LANs) is also spreading worldwide, which has led to increasing the interest in wireless LAN standards and applications, and their enhanced versions to support higher data rates and to consider some mobility.

As shown in Figure 1-1 [1], multiple technologies may be used for similar scenarios (taking into account mobility and range). For a certain technology, the higher the mobility we would like to consider, the lower the user data rate that can be achieved. The objective of novel developing technologies must be to allow for higher data rates and higher mobile speed, thus moving the boundary of achieved performance to the right in the shown axes.

![Figure 1-1. User data rate and scenarios for different wireless access techniques [1].](image)

Data rate requisites will be more challenging in the near future not only because of the high-performance and real time type of offered (and also demanded) applications, but also the number of users is increasing considerably and is expected to grow rapidly. As an example, the number of 3G/UMTS users worldwide exceeded 50 millions [2]. Thus, the study, development and evaluation of new techniques that can take advantage of different methods to improve system performance and increase user data rate has currently a clear importance for both the research community and the commercial deployment.

Several possibilities may be considered in order to improve quality of service and data rate in a communication system. Some of them consist in using conventional methods based on deployment optimization, such as frequency reuse, reduction of cells or use of micro or
picocells, or sectoring. Other options consider optimized signal processing techniques, e.g. improvement of power control techniques, or cooperative base stations. One of the solutions that is currently generating a great interest in the mobile communications community is the use of new packet-oriented data channels for the second version of 3G networks (the so-called HSPA techniques).

HSPA (High Speed Packet Access) technique mainly involves a software update in current UMTS networks, which allows a straightforward deployment. It is based on the improvement of the radio interface proposed in versions 5 and 6 of 3GPP standards, for both downlink (HSDPA) and uplink (HSUPA). It promises an improvement in the bandwidth offered to the final user, capacity increase for the network operator and interactivity for data applications. These advantages, as well as its compatibility with the first generation of the already deployed UMTS network, makes it especially suitable for offering high data rate services for mobile communications in the near future. As a result, its development and deployment has been sped up, especially in the downlink case, and up to now several telecom operators have included HSDPA [3] in some trial areas. As the main drawback of HSPA solutions, it is envisioned that the core network will have to support higher traffic, thus more nodes and new network dimensioning will likely be required.

So far, most of the considered solutions have taken advantage of frequency diversity, time diversity and code diversity. Recently, novel solutions including antenna arrays and processing techniques aspects have considered the advantage of spatial diversity as an extra plane to improve system performance. In the field of mobile communications, and compatible with the previous HSPA solution, spatial diversity for 3G was considered even in the initial versions of 3GPP standards [4], by including a pilot bit to be used as reference information in transmit-diversity schemes. Up to 2 antennas at the base station were proposed there, combined with several optional methods to make use of the available spatial diversity.

In the same trend of spatial diversity, a solution consists in the use of adaptive or smart antennas based on beamforming, at one end of the link. Roughly speaking, its operational principle is based on dynamically modifying the radiation pattern of the antenna array in order to improve certain parameters. Several research efforts have also been done by 3GPP working groups to include both smart antennas [5] and beamforming [6] in new 3G versions, but they have not been standardized yet. Still, the interest shown by scientific and standardization community gives us an idea of the importance of these techniques in future communication systems.

A further step in spatial diversity is the use of multiple antennas not only at one end of the link, but at both the transmitter and receiver sides. This technique is also known as MIMO (Multiple Input Multiple Output) systems. Since the pioneer works on MIMO presented by Telatar [7] and Foschini [8], there has been an explosion of research works on this topic,
including some standardization efforts, e.g. the 3G recommendations included in new version (releases 5 and 6) such as [9] and [10].

Also wireless LAN communications have been seen as an application where MIMO advantages can be exploited. The increasing demand of these types of systems, as well as the new services that require higher data rates such as high quality audio and video media, has promoted new solutions to improve the performance of WLAN systems. The objectives in data rate for future systems in indoor low-mobility scenarios are expected to be up to 1Gb/s [11], which is presented as a challenging requirement, particularly for systems with limited bandwidth, power and complexity. MIMO systems are shown as an interesting solution, since they allow to exploit the extra spectral efficiency offered by multiple antennas. Regarding this, some efforts in standardization has been realized so far by the IEEE 802.11 task group N, which is currently working, jointly with several partners and research groups, to define the forthcoming IEEE 802.11n MIMO WLAN standard [12]. Moreover, MIMO techniques has also been proposed to be used in larger wireless areas, such as WMAN (wireless metropolitan area networks), in what has been called WiMax solutions, which have already been standardized [13]. These examples show the interest that MIMO systems are arousing also in WLAN world.

The significant capacity enhancement promised by MIMO systems has led to an increasing research effort, so most universities have research groups working on some of the aspects of MIMO techniques, and the number of recent publications and works on this topic is significant. The capacity in a MIMO system does not only depend on the number of antennas, but on the joint effect of several parameters and characteristics of the whole system. Roughly speaking, the achievable performance in a MIMO system depends on three main aspects, namely: the physical structure of the channel, the radiation and configuration characteristics of the antenna array, and finally the MIMO algorithm at transmit and receive modules.

Regarding signal processing and transceiver algorithms for MIMO systems, multiple schemes have been proposed in the literature, some of them aiming at taking advantage of spatial diversity and some others based on spatial multiplexing. Conventionally, the two general cases of either no channel knowledge (CSI: channel state information) or perfect CSI at the transmitter have been considered. Recently, some algorithms and precoders have been proposed to take into account partial channel knowledge at the transmitter, thus allowing a more realistic feedback between transmitter and receiver. There are still some open issues regarding the type of information to be sent as feedback, how the delay and errors affect the feedback channel, etc.

In order to test MIMO schemes, as well as to design some algorithm parameters, it is of great interest to posses accurate channel models for multiple antenna systems. While initially simple MIMO channel models were used, based on simplified statistical parameters, more complex models have recently been presented. Some of them are physical models or are based on geometrical characteristics, while others include some grade of statistical behaviour. Some
aspects are still poorly treated, such as the polarization behaviour, the dense (non separable) multipath components or the elevation components. Moreover, in order to validate the existing models and to characterize certain specific situations, it is of paramount importance to realize channel measurement campaigns in different environments.

Usually, the MIMO schemes and transceivers are designed from a system point of view. However, from an antenna and propagation perspective, it is of paramount importance to account for the array configuration and the antenna characteristics in the whole system. While some theoretical models and also some antenna prototypes for MIMO have been presented in the literature, there is still a lot of work to be done in this field. Also related to this, the use of multipolarized antennas has been proposed as a very interesting way to reduce the array size and to take advantage of the extra polarization diversity.

A final step to test MIMO algorithms, to validate transceiver schemes and to characterize MIMO channels under real conditions, prototypes with multiple antennas are desirable. While the first developed systems were channel sounders with narrowband and offline capabilities, aiming at measuring channel characteristics, novel equipments have been presented to be able to measure the channel impulse response with very small resolution in the delay domain, as well as consider broadband operation. However, few examples have been found which include the option to consider different algorithms or antenna configurations. Recently, some prototypes are in development stage with the objective of implementing complete MIMO WLAN systems.

As shown in previous paragraphs, there are multiple characteristics to be studied in MIMO systems, from theoretical aspects such as capacity issues to modelling and system measurements. While some of them may be considered as quite mature such as conventional MIMO algorithms, others need a better understanding and extra work is desirable, such as antenna effects in MIMO transceivers or thorough analysis of using polarization jointly with MIMO techniques. Next chapter details the existing work for each of the previous aspects in MIMO and highlights some open issues.
1.2 Objectives and general methodology of the thesis

Based on the importance that multi-antenna technology will probable have for future communication systems, this thesis aims at **contributing to the understanding, analysis and improvement of multiple antenna systems by studying and evaluating several aspects, with special emphasis on those of importance for a successful implementation in real systems.** Beginning with the design and implementation of smart antennas, the main aspects of the real implementation and evaluation of an adaptive antenna are presented. As a further step in multi-antenna techniques, and taking advantage of the previous prototyping experience, a MIMO system is developed and some novel aspects of MIMO systems are analysed, including results from measurement campaigns in different situations. As a result, this thesis may help to a better understanding of the capabilities offered by MIMO techniques and may bring closer their use in real communication systems in a near future.

After a previous work of revision of current state of the art and main open issues in the wide research area of multi-antenna systems, four main topics of work have been identified, so that the main objectives of the thesis are wrapped around them. These main objectives are summarized as follows:

- **OBJECTIVE 1 (topic: adaptive antennas for real deployments):** this thesis will **analyse** the main real-implementation issues that prevent the deployment of **adaptive antennas in real networks**, and will propose possible solutions to reduce their complexity and to enhance their operation in real-time and within a real operator network. Several **beamforming algorithms for real implementation** will be studied, and the main **signal-processing blocks** of an adaptive antenna will be implemented and evaluated. In order to evaluate their performance, a **suitable evaluation method for adaptive antennas** will be developed and put into practice. In order to achieve this objective, an adaptive antenna prototype will be designed, implemented and tested, focusing on the signal processing modules.

- **OBJECTIVE 2 (topic: rapid prototyping and multi-purpose measurement system for MIMO):** this thesis will **propose an easy-to-use, low-cost and flexible measurement system for MIMO indoor**, with the purpose to facilitate the evaluation of MIMO channels for different scenarios and **antenna configurations**, as well as to simplify **testing MIMO algorithms** under realistic conditions. Moreover, the performance of MIMO systems will be evaluated and compared for different scenarios and antenna characteristics and conclusions about propagation characteristics for indoor MIMO channels will be drawn. In order to cover this objective, a multi-purpose MIMO testbed will be designed and implemented, and some measurements with different antenna configurations and in different scenarios will be carried out to show the possibilities of
the testbed and to bring some light into some not fully known characteristics of the MIMO channel.

- **OBJECTIVE 3 (topic: issues on propagation and base station configurations for MIMO):** this thesis will study propagation characteristics of MIMO channels focusing on two aspects of interest: outdoor-indoor MIMO channels for low mobility scenarios and enhancement of MIMO capacity by using polarization diversity in different scenarios. Moreover, and as one of the results to be drawn, the thesis will address the question of how to select the best location for base stations in order to deploy a MIMO system to cover indoor scenarios. Also different options for base stations combinations will be studied and suboptimal but realistic solutions will be proposed. As a result, best practices guidelines and recommendations for base stations (at physical layer) to improve indoor capacity for MIMO systems will be given. In order to accomplish this objective a measurement campaign with a MIMO system will be planned and performed for the considered scenarios and base station combinations, and the results will be carefully analysed.

- **OBJECTIVE 4 (topic: antenna arrays for realistic MIMO world):** this thesis will study how to design realistic antenna arrays for handheld devices in order to achieve a good performance in MIMO systems. Thus, a novel antenna array for a user terminal will be designed, considering in the design the required aspects to obtain an optimized performance for MIMO. Moreover, the evaluation of antenna arrays for MIMO systems will be addressed, by proposing several methods to analyse the antenna performance and comparing them. In order to cover this objective, an antenna array for MIMO will be implemented and afterwards evaluated with the proposed methods, so the results can be compared.

As observed from the description of the main objectives of the thesis, we may notice that a similar methodology is followed in the four areas in order to cover each of the objectives. First of all, a thorough revision of the state of the art in the topic of interest is carried out, in order to know the exact situation of research in each of the four areas to be covered. Afterwards, some theoretical studies are done, in order to propose a scheme, architecture or solution to be implemented and tested so that one or some of the open issues are studied. A real implementation and measurements are carried out, in order to obtain data under real conditions. These are a key aspect in this thesis, since we aim to foster the use of multi-antenna systems in real world. Finally, detailed processing of data and analysis of results is done, and the main conclusions for each of the four areas of study are given. Figure 1-2 represents graphically the methodology that has been followed throughout the thesis.
Figure 1-2. General methodology followed during thesis and type of result from each step.
1.3 Outline of the thesis

The rest of this document is organized as follows:

· **Chapter 2** introduces the main theoretical aspects of both smart antennas and MIMO systems. Furthermore, a thorough study of state of the art is given, and the main open issues are highlighted here.

· **Chapter 3** presents the design, implementation and evaluation of the signal processing modules of an adaptive antenna for mobile communications. Several aspects regarding the implementation on a software radio platform are included, such as real time constrains in the design of the signal processing modules and complexity versus performance trade-off. Also a novel method to measure and evaluate the adaptive antenna performance under controlled and realistic scenarios is explained, and results obtained for the realized prototype are included.

· **Chapter 4** describes the implementation of a MIMO testbed for testing algorithms, measuring several antenna configurations and characterizing MIMO channels in indoor scenarios. Several options regarding the type of antenna elements to be used, the data rate and the algorithms to be performed are allowed. Some examples are shown, including measurements and results.

· **Chapter 5** analyzes some possibilities for MIMO systems from a propagation point of view. Indoor to indoor and outdoor to indoor communications are compared, by using the data obtained from a measurement campaign. Also different locations for the transmitters (or base stations) are analysed. After that, the polarization diversity is introduced in the study, and single-polarized antennas are compared to multi-polarized ones. The advantages and drawbacks of multi-polarization systems are shown.

· **Chapter 6** focuses on an antenna perspective of MIMO systems. Some possibilities for realistic MIMO arrays are studied and a novel compact array design is presented, which is suitable for realistic user terminal. Some methods to evaluate antenna arrays for MIMO systems are presented, and by using the compact array and a reference array, they are compared, showing the benefits and drawbacks of each one.

· Finally, **chapter 7** summarizes the results and concluding remarks, and proposes some future research topics.

The most important results and contributions to the state of the art that were obtained along the work of these thesis have been presented and accepted for publication in leading international technical journals, and many conference papers have been presented at the most important international and national conferences in the fields of antenna and signal processing for communications. Specifically, this thesis has given rise to:

· **5 articles** published in technical journals.

· **21 papers** presented at international conferences.
· 10 papers presented at national conferences

Moreover, the work done throughout the thesis has produced one international patent. Besides, three invited talks at international conferences have been given presenting some of the results of this thesis. Finally, two Master Thesis have been supervised by the author of this thesis within this work’s scope. All the details of the contributions of this thesis are given in chapter 7.
THEORETICAL BACKGROUND AND STATE OF THE ART

2.1 Introduction and terminology: smart antennas, MIMO systems, beamforming and signal processing for spatial diversity ........................................ 16
  2.1.1 Introduction ........................................................................................................ 16
  2.1.2 Used terminology in this thesis ........................................................................ 17

2.2 A first step in multi-antenna techniques: smart antennas based on beamforming ......................................................... 18
  2.2.1 Basis concepts on smart antennas and beamforming ..................................... 18
  2.2.2 Beamforming and adaptive algorithms .......................................................... 19
      2.2.2.1 Basics on beamforming ................................................................. 20
      2.2.2.2 Adaptive algorithms: conventional and novel algorithms ................. 21
  2.2.3 Prototyping for smart antennas ...................................................................... 24

2.3 Making the most of spatial diversity: multi-antenna systems at transmitter and receiver (MIMO) ................................................. 26
  2.3.1 MIMO: basic concepts ...................................................................................... 26
  2.3.2 Narrowband and wideband characterization of the MIMO radio channel 28
  2.3.3 MIMO capacity ............................................................................................... 38
  2.3.4 MIMO signal processing: transceiver schemes and algorithms ............... 41
  2.3.5 MIMO channel measurements .................................................................... 45
  2.3.6 MIMO channel models ................................................................................... 48
  2.3.7 Antenna aspects in MIMO systems ............................................................... 54
  2.3.8 Prototyping for MIMO systems ..................................................................... 60
  2.3.9 Summary on state of the art in MIMO: what is new, what is lacking? .. 64
The purpose of this chapter is twofold. Firstly, an overview of the main concepts related with multi-antenna techniques and MIMO systems is given, with special emphasis on the aspects that will be addressed along the thesis. Thus, the general theoretical background required to understand the rest of this document is presented. As a second intention of this chapter, previous work on both adaptive antennas and MIMO systems is detailed, obtained after an extensive search effort in the existing literature. The aim is to give the reader an idea of the state of the art as well as to point out the main open issues and topics of current interest, which have motivated the work of this thesis.

Although both technologies are related, this chapter has been divided into beamforming-based antenna systems and MIMO systems. We refer to the first group as “smart antennas”, and we have used the term “MIMO systems” to refer to those systems based on the use of multiple antennas at both link ends. The use of these two terms in the thesis is clarified below.
2.1 Introduction and terminology: smart antennas, MIMO systems, beamforming and signal processing for spatial diversity

2.1.1 Introduction

Aiming at optimizing the possibilities offered by the radio channel, wireless systems have tried to exploit diversity in all its facets: frequency diversity, time diversity and, more recently in civil communications, code diversity. However, the advent of novel applications and user needs have led to a continuous increase in required capacity and data rate for radio communications. This has boosted the research in other diversity possibilities that may offer an enhanced system performance, without requiring extra frequency or bandwidth allocation (which is a scarce resource). One of the most interesting options is exploiting the spatial diversity inherent in the radio channel, by using multiple antennas in the system. A first step consists in using multiple antennas (or an antenna array) at one end of the link (usually at the base station or access point, due to less restrictive conditions of size and power compared with the user equipment).

Antenna array techniques are not novel: they have been used in radio applications for several decades. Some examples are arrays used for the base station in mobile communications, or conformal arrays used for satellite applications. The radioelectric characteristics of these antenna arrays are typically fixed, and hence they do not allow for radiation pattern adaptability. In these cases, the objective of using multiple elements is generally to achieve more directive radiation patterns than the ones usually obtained with a single element, or to fulfil a certain radiation mask, as well as to increase gain of the overall antenna module.

Contrary to the examples of previous paragraph, the so-called smart antennas based on beamforming techniques are characterized by their ability to change their overall radiation properties (main lobe bandwidth, main lobe angle, second lobe level…) in order to optimize certain parameters (received power, signal to noise ratio…). The main idea behind beamforming techniques is to take advantage of the extra information about the signal that is available thanks to using multiple antennas, mainly information about direction of arrival, so the received signals can be optimally combined. Beamforming techniques typically assume that the received signals are highly correlated, and that the main differences between signals are phase offsets due to wave propagation depending on the direction of arrival. A similar concept can be explained for transmitted signals. Obviously, this concept is applicable only if antenna elements are closely located and scattering and multipath effects are not produced mainly in the surrounding area of the antenna array, so that the direction of arrival can be considered as the same one for all the antenna elements.

When we want to take advantage of a rich multipath environment, as an indoor one with many scatterers present, beamforming is not the optimal technique to be used. Other techniques
based on the exploitation of uncorrelation of signals received at the array can be used, such as
the BLAST scheme for spatial multiplexing, or the Alamouti algorithm for spatial diversity,
both of them presented below. In general, these techniques give a better performance for low
correlated received signals, thus some separation of antennas in the antenna array is preferred.
Most of these schemes for spatial diversity and multiplexing are designed for using multiple
antennas not only at one end of the link, but at the two of them: these are the so-called MIMO
(multiple-input multiple-output) systems. However, we may note that strictly speaking the term
MIMO should refer to general systems with multiple antennas at receiver and transmitter,
independently of the type of signal processing scheme that is used behind the antennas.

2.1.2 Used terminology in this thesis

As outlined in previous paragraphs, the terms “smart antennas” and “MIMO systems”
may be the cause of some controversy when defining what they refer to. From the literal
meaning, MIMO systems should refer to systems where multiple antenna elements are used at
both the transmitter and the receiver, no matter which type of spatial approach (beamforming,
spatial multiplexing…) is used. In this sense, MIMO systems would be a wide term, used to
refer to multi-antenna technology in general (mostly which implies using multiple antennas at
transmitter and receiver). This meaning is used in many books and articles, for example, in [14].

Conversely, the term “smart antennas” is typically used to refer to antenna arrays
based on spatial processing and beamforming, as for example in [15]. Typically, this will
mean using multiple antennas only at one link end, that is, having a MISO (multiple input single
output) or SIMO (single input multiple output) system. In this sense, smart antennas would be a
subgroup of the more general term MIMO systems.

We will use the previous meanings when referring to “smart antennas” and “MIMO
systems”. Our work on smart antennas has focused on antenna array systems based on
beamforming techniques, where only one end of the link uses multiple antennas. After this, we
study the multi-antenna systems in a wider sense and consider multiple elements at both link
ends in what we have called MIMO systems. Although typical MIMO algorithms do not include
beamforming methods, there are also some examples of the application of these techniques
when multiple antennas are used at both sides of the link, as we will see later in the revision of
state of the art.

Nevertheless, we have to mention that other ways to understand these two terms are also
found in the literature. For example, in [16] the authors refer to smart antennas as architectures
with multiple antennas at transmitter and/or receiver, which employ any signal processing
technique to take advantage of spatial diversity, including beamforming, spatial multiplexing,
space-time codes, etc. Other authors (as [17]) use the term MIMO as a subgroup of smart
antennas that consider space-time architectures. We will not use these meanings, but the
previous ones.
2.2 A first step in multi-antenna techniques: smart antennas based on beamforming

2.2.1 Basis concepts on smart antennas and beamforming

Smart antennas (as this term is understood here) are based on the idea of using multiple antennas at one end of the link in order to exploit the benefits of spatial diversity. Although in a general case we could use multiple antennas to transmit (or receive) different data streams simultaneously, we will refer to smart antennas when they are based on the beamforming concept. In this case, the received signals are assumed to be highly correlated, so they can be combined in an optimum way (achieving interference reduction, maximum gain for a certain direction...). In the most advanced case, the combination is done adaptively to the radioelectrical conditions in the environment, while in the most simplified version of smart antennas, a switching between several predefined combinations is allowed to the user. The same concept of combination of signals or changes in the radioelectrical features of the system can be characterized as an adaptive or variable radiation pattern. We may note that beamforming techniques have been used for some time, and their fundamental bases were established some decades ago. However, their application to spatial diversity and mobile communications has recently been proposed and studied.

The smart antenna concept in the sense we use it here covers several solutions. A way to classify them considers the ability to adapt the radiation pattern to the environment. The simplest schemes are the so-called multibeam arrays and phased arrays. Usually based on phase shifters and switches, the first ones allow choosing between several predefined beams or radiation patterns, while the second ones permit the selection of the main lobe direction. However, in order to consider higher degrees of freedom when choosing the radiation characteristics, adaptive antennas are preferred. They not only allow to maximize the power received from the user, but also to reduce interference levels, as depicted in Figure 2-1.

Figure 2-1. Switched array vs adaptive array.
Since smart antennas apply techniques based on spatial selectivity, the best performance is obtained for environments where angular dispersion of multipath components is relatively small [18]. Thus, smart antenna techniques are usually applied in macrocellular scenarios, such as rural or suburban mobile communications. In this case, the multi-antenna array is located in the base station side, where the angular dispersion is typically smaller than the one observed in the user equipment side. Moreover, the size and signal processing restrictions are lower in the base station than in the user equipment, which also suggests restricting the use of smart antennas to the base station. For situations where the angular dispersion in the channel exceeds some upper bound (typically in the case of microcellular or indoor scenarios) the use of smart antennas is discouraged, being of interest the use of MIMO techniques, which will be presented in detail in section 2.3.

The benefits of using adaptive antennas in wireless communications are twofold. Firstly, the link gain in the desired user direction is improved; secondly, a reduction in the interference level may be achieved. Overall, the main advantages of adaptive antenna systems are:

- Link budget improvement, thus increasing the coverage or reducing the number of base stations to cover a certain area, given a transmitted power level.
- Interference level reduction, which allows to improve the offered quality of service in the system.
- The possibility to introduce new services in the system, such as localization or security advantages.

Despite their undoubted advantages, adaptive antennas also present some drawbacks. They can be summarized in a higher complexity in both hardware and software aspects, as well as a higher size required to include multiple antenna elements and the subsequent radiofrequency chains. Also a higher cost for base station may be expected, although the reduction in number of base stations may be accounted in profits studies. Due to the increase in complexity, it is the study of efficient algorithms that offer the possibility of real-time implementation on current software radio platforms are of great interest, as well as simplified schemes for adaptive antennas so they can easily been deployed. Chapter 3 of this thesis is devoted to the design and implementation of the signal processing modules of a real-time adaptive antenna, as well as its evaluation.

### 2.2.2 Beamforming and adaptive algorithms

Adaptive antennas are based on the ability of changing their radiation features as a function of the environmental electromagnetic characteristics. Since these characteristics are expected to change in time, the adaptive antenna system must be capable of adapting its
radiation parameters or figures of merit. In order to do so, adaptive algorithms are typically used, combined with beamforming techniques. Next sections present the basic concepts of beamforming and summarize the state of the art in adaptive algorithms applied to antenna arrays.

2.2.2.1 Basics on beamforming

The conventional beamformer for spatial selectivity performs a linear combination of the received signals in the antenna array, \( x(t) \), by multiplying them by a set of weights, usually referred to as vector of beamforming weights, \( w(t) \). The output signal of the beamformer is then:

\[
y(t) = \sum_{n=1}^{N} w_n^*(t)x_n(t) = w^H(t)x(t)
\]

where \((\cdot)^H\) denotes the hermitic function, and the input and weight signals have been expressed as vectors:

\[
x(t) = [x_1(t), x_2(t), \ldots, x_N(t)]^T
\]
\[
w(t) = [w_1(t), w_2(t), \ldots, w_N(t)]^T
\]

The equation (2-1) above is the general expression for the narrowband linear beamformer. Figure 2-2 depicts this scheme. Although previous expressions have been presented from the reception point of view, the same idea and expressions hold for the transmitter case.

![Figure 2-2. Narrowband linear beamformer](image)

Multiple adaptive algorithms have been developed and studied in the literature, and several of them have been proposed to be used in beamforming applications, from those based on temporal or spatial reference to others that exploit special characteristics of the signal such as cyclostationarity. Depending on the signal features it may be more interesting a certain type of adaptive algorithm to compute the beamforming weights. Also other aspects such as the suitability of the algorithms to be implemented in a certain software platform should be taken into account, thus considering the interest in realistic implementations and efficient or low-
complexity algorithms. Since part of the thesis contributions is real-implementation aspects of beamforming for adaptive antennas, we summarize in next paragraphs the main aspects of interest in beamforming techniques.

The most remarkable features to be chosen are [19]:

- **Type of reference:** the three conventional sorts of references are temporal, spatial and blind reference. While in the first case a pilot signal is used as reference, a direction of arrival or angular sector is required in the second case. Blind reference exploits some other characteristic of the signal, such as constant module of signal.

- **Linearity:** conventional beamformers combine the input signal as a linear function. However, some novel beamforming techniques have been presented that uses non-linear functions to optimally combine the input signals. This is the case in neural-network based adaptive algorithms or those based on the minimization of bit error rate. Higher degrees of freedom may be attained with the latter, at the expense of higher complexity.

- **Considered bandwidth:** a first version of beamforming module may consider only narrowband signals, which involves that delay spread is small and the channel response can be assumed as constant (“flat”) for the signal bandwidth. When a wideband signal should be processed, the multipath characteristics of the channel may be exploited, so the beamformer can be complemented with other structures such as Rake receivers with multiple “fingers”. Other types of wideband beamformers divide the received or transmitter signal into several sub-bands. More examples of wideband beamforming techniques exist.

- **Parameter of merit to be optimized:** when computing the beamforming weights for adaptive antennas, optimal or sub-optimal algorithms may be used to minimize or optimize a certain parameter. Conventional algorithms are based on the minimization of the mean square error (MMSE: minimum mean square error), but some other options have been studied, such as the minimization of the bit error rate (MBER).

### 2.2.2.2 Adaptive algorithms: conventional and novel algorithms

Since one of the chapters of the thesis deals with beamforming algorithms for adaptive antennas and their implementation for real-time operation, it results of interest to mention the most important adaptive algorithms for beamforming, as well as their characteristics. We will focus on the algorithms based on temporal reference, due to their special interest when the beamformer is designed to be applied in UMTS or other system with access to a pilot signal. Many communications can be found to this respect (see for example [18] and [20] for excellent summaries); we will here summarize the most interesting characteristics of the main
conventional algorithms, as well as presenting the state of the art in novel ones, without including the mathematical details.

**Conventional algorithms**

Typically, adaptive algorithms for temporal reference are based on the minimization of the mean square error (the MMSE concept), being the error computed as the difference between the beamformed signal $y(t)$ and the reference signal. The optimum solution is the well-known Wiener-Hoff filter, which offers a mathematical solution based on the statistical distribution of the received signal and the reference signal. The obtained solution is only interesting from a theoretical point of view and as an upper bound of the performance of sub-optimal solutions. In a realistic system, the Wiener solution cannot be computed, since the statistical characteristics of the signal are not known, but an estimation of them may be used instead.

With the Wiener solution as a starting point, several sub-optimal solutions have been presented in the literature, mostly based on the computation of adaptive weights iteratively and on the estimation of one or several parameters of the signal. A well known solution is the LMS algorithm (*Least Mean Squares*), which is a stochastic gradient based algorithm. The optimum solution is iteratively searched on the error surface, as a function of a certain adaptation step. This solution presents the advantage of a very low computational complexity, at the cost of slow convergence and fair performance. Several versions of the LMS algorithm have been studied, such as the normalized LMS or the LMS with adaptive step [19].

In order to improve the performance of the LMS algorithm, other sub-optimal solution is the RLS algorithm (*Recursive Least Squares*). It uses temporal averaged values of the autocorrelation matrix of the received signal. A parameter, the forgetting factor, is used to weight the recent computed matrices as more important in the averaging process. The convergence is higher than in the LMS case and, contrary to the LMS case, the results are not dependent on the statistical characteristics of the input signals. As a drawback, the computational complexity increases significantly. Several improved versions of RLS algorithm may be found, such as those considering an adaptive forgetting factor, or an accelerated version to improve convergence [21].

Similar to the idea presented in the RLS algorithm, the SMI (Simple Matrix Inversion) algorithm estimates the autocorrelation matrix and cross correlation matrix of received and reference signals as an averaged value from a temporal window. Its computational complexity is higher than the one for the RLS algorithm, and the results are very dependent on the selected window size.

Finally, another conventional algorithm for array beamforming is the *conjugate gradient* (CG) algorithm. It consists in updating the beamforming weights following the
conjugate gradient of the error, where the adaptation step depends on both the input signal and the instantaneous error. Several snapshots or realizations are used for updating each parameter. Although it offers a fast convergence, its computational complexity is high.

**Novel algorithms**

The advent of new communication systems where the use of adaptive beamforming is of interest has boosted the study and design of new beamforming methods and algorithms to be used. Of special importance are algorithms suitable for real implementation, such as those with a efficient design or simplified to reduce their complexity. Next paragraphs summarize some of the characteristics of some of the most interesting algorithms currently been proposed for beamforming in adaptive antennas.

The first studied types of efficient algorithms are those that implement *beamforming based on artificial neural networks*. Neural networks are a specific solution to pattern classification problems [22]. Conventionally, they have been used in several applications and problems, so the neural network technology is not new. Recently, and thanks to the development of powerful digital signal processor and software-radio platforms, neural networks have been proposed as an efficient method in digital signal processing for communications [23]. Regarding beamforming in antenna array, two applications have been devised for neural networks: neural networks as linear beamformers and neural networks as non-linear beamformers.

In the first case, the neural network is used as an approximation or generalization system of a function with multiple dimensions and a high degree of complexity, which otherwise would involve high computational complexity. The neural network module is used in this approximation as the weights computation module, instead of using a conventional adaptive algorithm. Examples of this type of application can be found in [24]-[27], where the performance of some linear beamformers based on neural networks is computed by means of simulations.

Conversely to the previous application, the second one considers the use of artificial neural networks as the beamforming module. Thus, the combination of input signals is performed with a non-linear system (the neural network-based system). This fact implies the possibility to improve the system performance, since problems that are not linearly separable can be solved. This can also be understood as an increase in the number of degrees of freedom, which in a linear beamformer are limited to the number of antenna elements minus 1 (N-1). The improvement in the beamformer features is attained at the expense of higher computational complexity. However, the advance in technologies that allow parallel computation, as DSP or FPGA platforms, has opened the door to the use of neural network based beamforming schemes, which inherently involve highly parallel processing. One of the pioneer works on this field is [28], where a neural network is used to mitigate the effect of fading in a channel when
multiple antennas are used at the receiver. Another non-linear beamformer based on the use of neural networks is the one presented in [29], which achieves a reduction of interference level.

Also related to neural networks, some remarkable works have been done based on the so-called Support Vector Machines (SVM) [30]. They can be considered a specific solution of neural networks, but the mathematics behind them are quite complex and the theoretical study presents a higher difficulty. A simplified alternative to SVM are the Relevant Vector Machines (RVM). In [31] another non-linear beamformer is used, which is shown to outperform the conventional linear beamformers, at the expense of higher complexity.

Another type of novel algorithms is the family of MBER-based algorithms. They are characterized by the optimization rule of minimizing the bit error rate (BER) after the beamforming processing, as opposed to the conventional MMSE-based beamforming algorithms. These algorithms have been studied for several years, but have been mainly used in equalization and detection [32]-[34]. In [35] the authors present a beamformer for an adaptive array, based on the MBER idea, for BPSK signals. The results show that the new beamformer obtains a lower BER than algorithms based on MMSE methods, especially for hostile situations (high level of interference, high number of undesired signals…). Other versions of MBER algorithms have been presented in the literature, such as [36] or [37]. Despite its good performance, most studies on this method are theoretical ones or simulation-based, but no real implementations have been presented so far, mainly due to its high computational complexity.

Finally, it is worth mentioning methods to add broadband capabilities to beamforming algorithms. The basic broadband beamforming scheme consists in using several tapped-delay lines [19], where different weights and delays are used in each line or branch. The basic concept when operating with a broadband beamforming is to obtain multiple narrow bands which are equivalent to the whole broadband. Several efficient algorithms have been proposed to optimize the computation of the required parameters. We highlight here the work in [38], which makes use of sub-bands to reduce the computational load.

### 2.2.3 Prototyping for smart antennas

A great research effort has been observed for the last few years in the area of smart antennas, with special emphasis in realistic aspects to allow for an introduction in real cellular scenarios. In spite of this, the worldwide deployment of these systems is not a reality yet, mainly due to the increase in system complexity and subsequent high cost [39]. Only smart antenna systems based on phased arrays for 2G have been commercially deployed with success in practice [40]-[45].

When facing the design and development of an adaptive antenna system for 3G, the complexity to be addressed clearly increases. Following the conventional concept in 3G, the
beamforming module is done in the baseband part and not in the RF part. It is hence performed after demodulation of the CDMA signal, for each user channel separately. Therefore, the adaptive antenna modules should be integrated into the signal processing sections of the Node-B. As a result, the implementation of adaptive antennas in 3G networks requires flexible and reconfigurable architectures in the base station or Node-B modules, and in general it would involve changing or at least updating the already installed Node-B’s, which must be especially designed to be able to work with adaptive antennas.

Figure 2-3. Smart antenna arrays base stations for WiMax, from ArrayComm [44].

Figure 2-4. 12-element adaptive antenna, for wireless local loop system Super WLL, by Kyocera [45].

Most of the existing solutions including smart antennas for 3G system have been developed for a specific base station manufacturer [46][47]. This fact has discouraged mobile communications operators from using smart antennas, due to the high associated cost and the inherent dependence on the equipment manufacturer. A “plug and play” smart antenna, with the capability of being used with any base station and independent of the specific manufacturer may be a great advantage in comparison with the conventional smart antenna concept.
2.3 Making the most of spatial diversity: multi-antenna systems at transmitter and receiver (MIMO)

Generally speaking, MIMO systems may be considered as a generalization of systems with multiple antennas at only one end (MISO or SIMO). This general concept is the one we refer to when talking about MIMO systems in this thesis. From this point of view, MIMO would be a natural evolution from smart antennas to more complex systems with multiple antennas at transmitter and receiver. However, the fact of using multi-element antennas at both ends involves the introduction of more concepts and the use of spatial diversity in a different way than just the array gain or interference reduction offered by smart antennas.

Scanning the titles of contributions to conferences and technical journals on radio and communication systems in the last few years, one could get the impression that MIMO is set to solve the bottleneck in wireless systems for good, and that most work on this type of systems has already been done. Although it seems clear that MIMO will be the enabling technology for high-speed wireless data, many questions still remain, and as we will see along this section, there are still some issues to address.

The main concepts about MIMO systems are presented in the rest of this section, as well as a complete state of the art in the main topics related with this thesis.

2.3.1 MIMO: basic concepts

The key idea in MIMO is to exploit the benefits obtained by using multiple antennas at transmitter and receiver, thus introducing an extra diversity degree. Figure 2-5 shows the concept of MIMO systems, where we have used the notation $N_T$ and $N_R$ for the number of transmit and receive antennas, respectively.

![Figure 2-5. The MIMO concept.](image)

MIMO may offer three different benefits, namely beamforming gain, spatial diversity and spatial multiplexing (Figure 2-6). By beamforming, the transmit and receive antenna patterns can be focused into a specific angular direction by appropriate choice of complex baseband antenna weights, in a similar fashion as in smart antennas but in both transmitter and
receiver. As it was stated in previous sections, the more correlated the antenna signals are, the better for beamforming. Under the line of sight (LOS) channel conditions, the gains at transmitter and receiver add up, leading to an upper limit of $N_T \times N_R$ for the beamforming gain of the MIMO system.

![Diagram](image)

**Figure 2-6. The three aspects of benefit offered by MIMO.**

Multiple replicas of the radio signal from different directions in space give rise to spatial diversity, which increases reliability of the fading radio link. The best situation in this case is a spatially white MIMO channel, i.e. completely uncorrelated antenna signals. Then, the diversity order is limited to $N_T \times N_R$. In real environments there may be some level of correlation of the antenna signals, which reduces the diversity order and is therefore an important channel characteristic.

Finally, MIMO channels can support parallel data streams by transmitting and receiving on orthogonal spatial filters, applying the idea of what is usually called spatial multiplexing. The number of multiplexed streams depends on the rank of the instantaneous channel matrix $H$, which, in turn, depends on the spatial properties of the channel and environment. The spatial multiplexing gain is limited to $\min(N_{T_0}, N_{R_0})$, which may be reached for sufficiently rich scattering environments.

Depending on the channel characteristics, and specifically on the correlation (diversity versus directivity characteristics) among antenna signals, it may be of interest to use a certain technique. The threefold trade-off between beamforming, spatial diversity and multiplexing, and its dependence on the propagation channel, is shown in Figure 2-7.
2.3.2 Narrowband and wideband characterization of the MIMO radio channel

In order to properly design and develop MIMO systems, it is of paramount importance to characterize the MIMO channel as accurate as possible and for different environments. Since a part of this thesis deals with channel characterization in different situations, an introduction on the mathematical representation and parameters of the MIMO channel are given here.

Let assume a system with $N_T$ antennas at the transmitter and $N_R$ antennas at the receiver. If a narrowband radio channel is considered, the mathematical expression for the input-output signals is given by:

$$y(t) = H(t)x(t) + n(t)$$  \hspace{1cm} (2-3)

where $x(t)$ is an $N_T$ vector representing the transmitted signal, $y(t)$ is the $N_R$ received vector, $H(t) \in \mathbb{C}^{N_R \times N_T}$ is the instantaneous narrowband radio channel matrix and $n(t)$ represents the noise vector. The time in eq. (2-3) represents the time variation or observation instant, which is also called snapshot.

When characterizing the channel, it is important to differentiate between narrowband and wideband characteristics. The system is considered narrowband if the channel transfer function is approximately flat over the working frequency bandwidth (also called flat-fading response). If the channel transfer function is not constant over the whole frequency bandwidth, it is considered a frequency selective channel, and then a frequency dependence appear in the channel transfer function. This may also be represented as a delay dependence in the channel impulse response, $H(t, \tau)$. The input-output relation is then given by:

$$y(t) = \int_{\tau} H(t, \tau) x(t - \tau) + n(t)$$  \hspace{1cm} (2-4)
When discrete delay responses or *multi-paths* (or also *taps*) are considered, the system output \( y(t) \) may be obtained as an addition of the contributions of each tap or path, as:

\[
y(t) = \sum_{i=1}^{L} H(t, \tau_i) x(t - \tau_i) + n(t)
\]  

(2-5)

where \( L \) is the number of paths that constitutes the channel response. This idea is of great interest in wideband channel modelling, as we will present later.

We may classify the features of the radio channel in narrowband and wideband characteristics. An overview of these two types of characteristics is given below, since they will be referred in some of the chapters of this thesis. More information on channel characterization may be found in [48].

**Narrowband characterization of the radio channel**

For the narrowband characterization of radio channels, we consider the time variation of the channel response, also called *fading*. Two components are usually considered: the long-term and the short-term fading

- The *long-term fading* or *slow-fading* shows slow changes in signal strength over a large travel distance or time interval (for mobile communications). It is caused by different path losses for different Tx-Rx distances (usually considered as free-space path loss), changes in the environment resulting in multi-path attenuation (shadow fading), weather conditions, etc. In general it may be statistically characterized as a log-normal distribution.

- The *short-term fading* or *fast-fading* are rapid changes in signal strength over a small travel distance or time interval. Contrary to the long-term fading, it is due to the constructive or destructive interference caused by superposition of multipath components. Thus, it is highly dependent on the speed of the mobile station (MS) and the surrounding objects. It is statistically characterized with a 2D Gaussian probability function, that is, its amplitude follows a Ricean function while the phase is uniformly distributed.

The Ricean distribution represents the fast fading statistics for a general case, where we may encounter both line of sight (LOS) and non-line of sight (NLOS) paths that contributes to the received signal. For the specific case of only NLOS contributions, the amplitude distribution of fast fading simplifies to a Rayleigh distribution. The ratio between the direct (LOS) path and the scattered (NLOS) components is called the Ricean *K-factor*.

An interesting parameter to characterize the narrowband channel response in the time domain is the *coherence time*, which is calculated from the temporal autocorrelation function (ACF) of the channel transfer function. The temporal ACF describes how fast the channel changes in time, and it is calculated on the bases of the narrowband time-variant transfer function of the channel, \( h(t) \), as:
Chapter 2

\[ r'(\Delta t) = \frac{T_{\text{spec}}}{0} h(t) h^*(t + \Delta t) dt \]  \hspace{1cm} (2-6)

From the temporal ACF, we may compute the temporal correlation coefficient as:

\[ \rho'(\Delta t) = \frac{r'(\Delta t)}{r'(0)} \]  \hspace{1cm} (2-7)

The coherence time for a certain \( x\% \) is defined as the minimum time interval \( \Delta t \) that fulfills the equation:

\[ \left| \rho'(\Delta t = T_{\text{coh},x\%}) \right| = x\% \]  \hspace{1cm} (2-8)

Typically a level of \( x\% = 1/e = 37\% \) is used.

Closely related to the coherence time and the speed of the MS is the **Doppler spectrum**, which is a narrowband characterization of the channel in the frequency domain. The Doppler spectrum is defined as the received power spectrum for a pure sinusoidal transmitted signal, and it is due to the Doppler effect that experiments each multipath component, and their subsequent combination in the frequency domain. In mobile communications, the Doppler spectrum is typically modelled as a Jakes spectrum, which involves a uniform distribution of random scatterers in 2D. Although generally accepted, this assumption is only true for rich-multipath indoor environments, which makes it not very realistic for outdoor cases, especially in rural areas. Two widely used parameters to characterize the Doppler spectrum are the mean Doppler shift \( f_d \), and the Doppler spread, \( \sigma_{f_d} \). The Doppler spread and the coherence time are inversely proportional parameters; as a rule of thumb, we may take \( \sigma_{f_d} \cdot T_{\text{coh},37\%} \approx 1 \).

We may note that so far all the presented parameters are general channel characteristics without the restriction to MIMO channels, that is, they are also valid for the conventional SISO channels.

**Wideband characterization of the radio channel**

In order to characterize the radio channel in a wideband sense, a high bandwidth is required in the measurement equipment. This is the reason why the MIMO channel was firstly characterized from a narrowband point of view, while the wideband characterization has become popular in recent years. The wideband characterization of the channel is done by analyzing the frequency selectivity, that is, how the channel changes over the frequency, for the same time instant. Similarly to the case of narrowband characterization, we may consider the wideband characteristics of the channel in the time domain or in the frequency domain.

When the time domain is used, the channel is characterized based on the **power delay profile** (PDP). For a certain time instant \( t \), it is computed from the channel impulse response, \( h(t, \tau) \), as:
Thus, the PDP shows the received power when a delta function is sent, as a function of the delay $\tau$. It is a representation of the multipath propagation, where each path or tap is characterized by a power and a delay. The PDP statistics can be studied by using two parameters: the mean delay $\tau_d(t)$ and the delay spread $\sigma_\tau(t)$. They are computed as:

$$\tau_d(t) = \frac{\int_{-\infty}^{\infty} \tau \cdot PDP(t, \tau) d\tau}{\int_{-\infty}^{\infty} PDP(t, \tau) d\tau} \quad (2-10)$$

and

$$\sigma_\tau(t) = \sqrt{\frac{\int_{-\infty}^{\infty} (\tau - \tau_d(t))^2 \cdot PDP(t, \tau) d\tau}{\int_{-\infty}^{\infty} PDP(t, \tau) d\tau}} \quad (2-11)$$

These two parameters are of great importance when characterizing wideband radio channels for communication systems, since they must be taken into account when designing certain system parameters such as the optimum symbol period, the expected delays for a rake receiver, etc. Each environment has different typical values for mean delay and delay spread. A delay spread of 5 to 10 ns may be expected in indoor residence buildings, while 100 to 200 ns (or even higher values such as 500 ns, depending on the specific scenario) are values to be found for outdoor microcell and macrocell environments.

When the frequency selectivity of the channel is characterized in the frequency domain, the study is based on the frequency autocorrelation function. Roughly speaking, this function describes how flat the channel is for different frequency bands. It is calculated by using the wideband time variant transfer function of the radio channel, $H(t,f)$, as shown in eq. (2-12):

$$r^f(\tau, \Delta f) = \int_{-\infty}^{\infty} H(t, f) \cdot H^*(t, f + \Delta f) df \quad (2-12)$$

An interesting parameter that can be computed from the frequency autocorrelation function is the coherence bandwidth, which approximately indicates the frequency bandwidth where the transfer function cannot be considered as “flat” any more. Let $\rho^f(\tau, \Delta f)$ be the frequency autocorrelation coefficient, defined from the frequency ACF as:

$$\rho^f(\tau, \Delta f) = \frac{r^f(\tau, \Delta f)}{r^f(\tau, 0)} \quad (2-13)$$

Then, the coherence bandwidth for an $x\%$ level is the minimum bandwidth $B_{coh,x\%}$ that fulfils the equation:

$$\left| \rho^f(\tau, \Delta f = B_{coh,x\%}) \right| = x\% \quad (2-14)$$
Typically a value of \( x\% = 1/e \approx 37\% \) is used in previous equation.

Similarly to the narrowband case, for the wideband case there is also an inversely proportional relation between the delay spread \( \sigma \tau (t) \) and the coherence bandwidth \( B_{coh,x\%} \). As a rule of thumb, it may be assumed that \( \sigma \tau \cdot B_{coh,37\%} \approx 1 \).

Again, it is interesting to note here that the parameters and characteristics presented up to now may be applied for general radio channels (regardless of the number of transmit or receive antennas).

As a summary, Table 2-I shows the main ideas on narrowband and wideband characterization of the channel presented in this section.

<table>
<thead>
<tr>
<th>Studied channel characteristic</th>
<th>Narrowband characterization</th>
<th>Time variation of the channel (slow and fast fading)</th>
<th>Wideband characterization</th>
<th>Frequency selectivity of the channel (flat or frequency selective channel)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Function in time domain</td>
<td>Temporal autocorrelation function, ( r'(\Delta t) )</td>
<td>Power delay profile, ( PDP(t, \tau) )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Parameters in time domain</td>
<td>Coherence time, ( T_{coh,x%} )</td>
<td>Mean delay, ( \tau_d(t) )</td>
<td>Delay spread, ( \sigma \tau )</td>
<td></td>
</tr>
<tr>
<td>Function in frequency domain</td>
<td>Doppler Spectrum, ( S(f_d) )</td>
<td>Frequency autocorrelation function, ( r'(t, \Delta f) )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Parameters in frequency domain</td>
<td>Mean Doppler shift, ( f_d^B )</td>
<td>Coherence bandwidth, ( B_{coh,x%} )</td>
<td>Doppler Spread, ( \sigma f_d )</td>
<td></td>
</tr>
</tbody>
</table>

Table 2-I. Main parameters in characterization of radio channels.

The MIMO channel: spatial characterization

The parameters and functions previously shown are used to characterize radio channels in general, and have typically been studied in SISO radio links for several years. Since the MIMO channel consists of \( N_R \times N_T \) radio links, all these parameters are also applicable to the MIMO case. However, and due to the use of multiple antennas at transmitter and receiver, there are some other parameters that should also be studied in the specific case of MIMO channels, for both the narrowband and the wideband case. They are presented below, and further referenced along this thesis.

In the case of MIMO channels, it is of relevant importance to analyze the spatial characteristics of the channel, namely the channel transfer function as a function of the angle of departure at the transmitter and the angle of arrival at the receiver, since the departing (impinging) wave will combine with a different phase for different array elements, depending on its angle of departure (arrival). This type of channel is usually called double-directional propagation channel, since it includes angular information for both the transmitter and the
receiver sides. For a narrowband case, we may use the angular resolved channel transfer function, \( H(t, \Omega_T, \Omega_R) \), while for a wideband case we may use the angular resolved channel impulse response \( h(t, \tau, \Omega_T, \Omega_R) \), where \( \Omega_T \) and \( \Omega_R \) represent the angular information (azimuth and elevation) at the transmitter and the receiver, respectively.

One interesting parameter to be studied is the power angular spectrum (PAS), which describes the relative receive or transmit power as a function of the angle, and highly depends on the channel spatial conditions (number of scatterers, position of clusters, etc). It is usually characterized for the receiver and the transmitter separately, although it is important to note that in general the PAS for the receiver and the transmitter are not independent, as it is explained later. For a certain time instant \( t \), the PAS at the transmitter can be computed from the angular resolved channel transfer function for the Tx (where the dependence with received angles \( \Omega_R \) is dropped):

\[
PAS^T_x (t, \Omega_T) = |h(t, \Omega_T)|^2
\]

(2-15)

Similarly, for the Rx:

\[
PAS^R_x (t, \Omega_R) = |h(t, \Omega_R)|^2
\]

(2-16)

In most studies, the time dependence is also dropped, so the integration over time is done. Also the angular description is usually simplified by assuming that most rays impinge in the horizontal plane, thus leading to a 2D representation with angular variation in azimuth (\( \Omega = \phi \)).

From the PAS, two parameters may be computed to study the spatial characteristics of the channel: the mean transmitted (received) direction, \( \bar{\phi}(t) \) and the azimuth spread \( \sigma_\phi(t) \) at the transmitter (receiver), which are computed as:

\[
\bar{\phi}(t) = \frac{\int_{\phi_{\text{max}}-\pi}^{\phi_{\text{max}}+\pi} \phi \cdot PAS(t, \phi) d\phi}{\int_{\phi_{\text{max}}-\pi}^{\phi_{\text{max}}+\pi} PAS(t, \phi) d\phi}
\]

(2-17)

and

\[
\sigma_\phi(t) = \left( \frac{\int_{\phi_{\text{max}}-\pi}^{\phi_{\text{max}}+\pi} (\phi - \bar{\phi}(t))^2 \cdot PAS(t, \phi) d\phi}{\int_{\phi_{\text{max}}-\pi}^{\phi_{\text{max}}+\pi} PAS(t, \phi) d\phi} \right)^{1/2}
\]

(2-18)

The mean direction of arrival (DoA) and direction of departure (DoD), and the angular spread at Tx and Rx are two parameters of great importance in spatial characterization of the channel, and are very dependent on the environment. For indoor channels the angular spread of both the Tx and Rx is considered large and quite similar. For outdoor scenarios, the angle
spread for the base station (BS) is usually much smaller than for the MS, since the BS is generally situated on a building rooftop and thus it is located further from the scatterers.

When wideband characterization is desired, the PAS not only depends on the angle of arrival (or departure) and the time (long-term behaviour), but also on the delay. This means that different multipath components are characterized by different DoA and DoD.

We may observe that in order to obtain an accurate angular resolved channel transfer function, many antenna elements may be used to measure the channel, or a post-processing to increase angular resolution in the DoA or DoD estimation is required. When the specific information on angular resolved paths is not available, another interesting parameter to be studied is the spatial correlation in the MIMO channel. For a general case, the (narrowband) spatial autocorrelation function of a channel may be computed as:

$$r^x(t, \Delta x) = \int_{-\infty}^{\infty} H(t, x) H^*(t, x + \Delta x) dx$$  \hspace{1cm} (2-19)

where the angular dependence $\Omega$ of the channel has been substituted by a spatial dependence, $x$, which in fact are directly related. This parameter may be calculated for the Tx and the Rx. Similar to what was shown for the temporal and frequency cases, the spatial correlation coefficient and the coherence distance can be computed from the spatial autocorrelation function, for both Tx and Rx.

Eq. (2-19) represents the variation of a channel over a certain direction, $x$. However, for the MIMO case it may be of interest to study the correlation between two antenna elements, when specific relative positions or spacing among elements are assumed. Then, instead of using the autocorrelation function, defined by eq. (2-19), we may use the correlation coefficient between two antenna elements to characterize the spatial correlation of transmitters (or receivers) in the MIMO channel. Let $h_{y,x}(t)$ denote the (y-th, x-th) complex element in the $N_r \times N_t$ narrowband MIMO channel transfer function $H(t)$. Then the correlation coefficient for the transmitter elements $x_i$ and $x_j$ may be computed as $\rho_{x_i,x_j}^{Tx} = \langle h_{y,x_i} \ast h_{y,x_j} \rangle$. Similarly, the correlation coefficient for the elements $y_k$, $y_l$ at the receiver is computed as $\rho_{x_i,x_j}^{Rx} = \langle h_{y,x_i} \ast h_{y,x_j} \rangle$, where the correlation coefficient for two variables $a$, $b$, $\langle a, b \rangle$ is defined as:

$$\rho_{a,b} = \langle a, b \rangle = \frac{E\{ab^*\} - E\{a\}E\{b^*\}}{\sqrt{[E\{|a|^2\} - E\{|a|^2\}][E\{|b|^2\} - E\{|b|^2\}]}}$$  \hspace{1cm} (2-20)

where $(\cdot)^*$ is the conjugate operation and $E\{\cdot\}$ denotes the expectation operation.

Depending on the type of information used from $H$ to compute the correlation coefficient, we may consider three different cases to compute the spatial correlation in Tx and
Rx for the MIMO channel: complex correlation coefficient, envelope correlation coefficient and power correlation coefficient. The three cases are defined as follows:

- **Complex spatial correlation coefficient:**
  \[
  \rho_{xy,xy}^{\text{cplx}} = \langle h_{x,x}, h_{y,y} \rangle \tag{2-21}
  \]

- **Envelope spatial correlation coefficient:**
  \[
  \rho_{xy,xy}^{\text{env}} = \langle |h_{x,x}|, |h_{y,y}| \rangle \tag{2-22}
  \]

- **Power spatial correlation coefficient:**
  \[
  \rho_{xy,xy}^{\text{pow}} = \langle |h_{x,x}|^2, |h_{y,y}|^2 \rangle \tag{2-23}
  \]

The complex correlation coefficient carries the full information of the channel, both the amplitude and phase (angular) information. Thus, it is generally preferred in the context of modelling. However, the envelope and power correlation coefficients have a clearer engineering interpretation, so sometimes they are preferred for characterization of the channel, especially when channel measurements are available [49]. In Chapter 5, a thorough analysis of spatial correlation for several measured scenarios is done, and further comments on the correlation coefficient are given.

It is interesting to notice that the spatial correlation at the transmitter is computed for any receiver end. Conversely, the spatial correlation at the receiver is computed for any transmitter end. Thus, the spatial correlation is characterized independently for the two links with this method, as depicted in Figure 2-8.

![Figure 2-8. Representation of Tx and Rx spatial correlation](image)

The spatial correlation at the transmitters or at the receivers are of great interest, since they give an idea of the spatial diversity available at each link end. Moreover, spatial correlation of antenna signals has been early identified as the show-stopper of MIMO systems [50]. Some
works [51] have assumed that Tx and Rx propagation environments are completely independent, and thus the spatial correlations at Tx and Rx fully characterize the channel conditions. However, this assumption is only true if the clusters of scatterers are local to the Tx and Rx, that is, the main propagation scenario for the Tx are clusters close to the Tx and far from the Rx, and conversely for the Rx. As a result, for this specific case the multipath components arriving at the receiver have “forgotten” the specific propagation conditions at the transmitter and the MIMO correlation properties are characterized by separate correlation matrices at the receiver and the transmitter.

The previous approach neglects the correlation terms across the link, which is usually called cross correlation or joint correlation, and are the terms that make the MIMO system something more than just the addition of MISO and SIMO. These terms do matter, at least in some indoor scenarios [52]. A surprising result, emphasizing that separate Rx and Tx correlation matrices are not able to completely describe MIMO channels was recently presented in [53], where it was shown that the so-called diagonal correlations may boost the ergodic capacity beyond the previously accepted upper limit of totally uncorrelated i.i.d. random channels. Thus, in order to properly study a MIMO system we should characterize not only the Tx and Rx correlation properties, but also the overall degrees of diversity in the system. The study of the rank of the channel matrix $H$ (eigenvalues decomposition) and the correlation matrix of the whole channel response $R$ may be of interest.

The eigenvalue decomposition and interpretation in MIMO systems

The substantial increase in data rate and channel capacity predicted for MIMO systems is based on the premise of achieving multiple orthogonal subchannels between the two ends of the path. A condition to attain it is to have a sufficiently rich scattering environment. The concept of orthogonality emphasizes the idea of independence among these subchannels, thus allowing the communication of multiple data streams simultaneously and improving the system performance. As a result, it is of paramount importance to characterize the spatial diversity and the available multiple subchannels in the system.

A mathematical way to express the number and importance of each orthogonal subchannel in a link between two terminals is the singular value decomposition (SVD) of the channel matrix $H$ or the eigenvalue decomposition (EVD) of the instantaneous correlation matrix $R$ when $R$ is a square matrix defined as:

$$ R = HH^H \in \mathbb{C}^{N_x \times N_x}, \text{ if } N_{Rx} \leq N_{Tx} $$

$$ R = H^H H \in \mathbb{C}^{N_y \times N_y}, \text{ if } N_{Rx} > N_{Tx} $$

(2-24)

Let $H \in \mathbb{C}^{N_x \times N_y}$ be the MIMO channel matrix. Its SVD is defined as:

$$ H = U \Sigma V^H $$

(2-25)
Similarly, the EVD of the correlation matrix $R$ is defined as:

$$
HH^H = U \Gamma U^H, \quad \text{if } N_{Rx} \leq N_{Tx}
$$

$$
H^H H = V \Gamma' V^H, \quad \text{if } N_{Rx} > N_{Tx}
$$

(2-26)

For both the SVD and the EVD:

- $U = [u_1, u_2, ..., u_{N_x}] \in \mathbb{C}^{N_x \times N_x}$ is the unitary matrix of left singular vectors, $u_i$.
- $V = [v_1, v_2, ..., v_{N_T}] \in \mathbb{C}^{N_T \times N_T}$ is the unitary matrix of right singular vectors, $v_i$.
- $\Sigma = \text{diag}(\sigma_1, \sigma_2, ..., \sigma_{N}) \in \mathbb{R}^{N_x \times N_T}$ is the diagonal matrix of the singular values $\sigma_i$, which are defined so that $\sigma_1 > \sigma_2 > ... > \sigma_N \geq 0$.
- $\Gamma = \text{diag}(\lambda_1, \lambda_2, ..., \lambda_{N}) \in \mathbb{R}^{N_T \times N_T}$ is the square diagonal matrix of the eigenvalues $\lambda_i$, which are defined so that $\lambda_1 > \lambda_2 > ... > \lambda_N \geq 0$.

The correlation matrix is then defined so that the diagonal matrix $\Gamma$ is an $N \times N$ matrix where the dimension $N$ is $N = \min(N_T, N_R)$, which is the upper limit of spatial multiplexing gain that an $N_R \times N_T$ MIMO system may offer. The actual non-zero eigenvalues of the diagonal matrix, $K$, defines the rank of the matrix $R$, which is the number of parallel subchannels available in the system:

$$
K = \text{rank}(R) \leq \min(N_T, N_R)
$$

(2-27)

The singular values and the eigenvalues are related by the equation:

$$
\sigma_i^2 = \lambda_i
$$

(2-28)

From an engineering point of view, the singular vectors $u$ and $v$ may be interpreted as weights that beamform the signal, so that $K$ orthogonal subchannels are obtained, while the eigenvalues represent the gain for each subchannel. A schematic representation of this interpretation is depicted in Figure 2-9.

![Figure 2-9. Illustration of parallel subchannels for an $N_R \times N_T$ MIMO system, where $K$ is the rank of the MIMO channel.](image-url)
2.3.3 MIMO capacity

The spectral efficiency and the capacity of MIMO channels have widely been studied for different transceiver schemes and scenarios, as well as for multiple antenna configurations. From the point of view of information theory [54], the channel capacity is defined as the maximum mutual information offered by the channel. This is the theoretical maximum amount of information that can be transmitted over a bandwidth-limited channel for which error-free transmission is possible in the context of Gaussian channel [55]. For MIMO channels, Telatar [7] derived the expression of channel capacity:

\[
C = \max_{Q} \log_2 \det \left( \mathbf{I}_{N_t} + \frac{\rho}{N_r} \mathbf{H} \mathbf{Q} \mathbf{H}^H \right)
\]

(2-29)

where \( \mathbf{I} \) is the identity matrix, \( \rho \) is the signal to noise ratio at the receiver, \( \mathbf{H} \) is the MIMO channel matrix, and \( \mathbf{Q} \) is the covariance matrix of the transmitted signals, which must satisfy a transmitted power constrain so that \( \text{Tr}(\mathbf{Q}) \leq N_r \).

Depending on the power allocation method, different covariance matrix \( \mathbf{Q} \) will be obtained. Usually, two opposite cases are considered: no channel knowledge (channel state information, CSI) at the transmitter and perfect CSI at the transmitter. Since they will be used along the thesis, a summary of the method and the capacity expressions are presented below.

**Uniform power allocation scheme**

When channel is unknown at the transmitter, it is reasonable to allocate power as a spatially uniform distribution. Thus, the covariance matrix is the identity matrix \( \mathbf{Q} = \mathbf{I}_{N_t} \), and the capacity equation (2-29) simplifies to the well-known expression [7]:

\[
C = \log_2 \det \left( \mathbf{I}_{N_t} + \frac{\rho}{N_r} \mathbf{H} \mathbf{H}^H \right)
\]

(2-30)

An example of MIMO scheme that uses uniform power allocation scheme is the use of space time block codes, such as the well-known Alamouti scheme.

**Water-filling power allocation**

In the ideal case of perfect instantaneous knowledge of channel at the transmitter, the optimum power allocation scheme that maximizes capacity is the so-called water-filling scheme or SVD decomposition (eigenbeamforming) of the channel [56]. There is a simple algorithm to find the solution, and the resulting capacity is given by:

\[
C_{WF} = \sum_{i=1}^{K} \log_2 \left( \mu \lambda_i \right)
\]

(2-31)

where the parameter \( \mu \) is chosen to satisfy
Theoretical background and state of the art

\[ \rho = \sum_{i=1}^{K} \left( \mu - \frac{1}{\lambda_i} \right)^+ \]  

(2-32)

where \( \rho \) is the signal to noise ratio SNR, \((\cdot)^+\) denotes taking only those terms which are positive, and \(\lambda_i\) are the non-zero eigenvalues of the correlation channel matrix \(R\) defined in (2-24). The idea behind this solution is to transmit more power through the better orthogonal subchannels, which are the ones with higher associated eigenvalue. The Tx and Rx optimum beams are obtained from the singular vectors as shown in Figure 2-9, in order to create the orthogonal subchannels. Figure 2-10 depicts the optimum power allocation with the water-filling scheme: the orthogonal subchannels with high \(\lambda_i\) (low \(\lambda_i^{-1}\)) are first “filled” with power, while the worse ones are last associated with power and the power allocated for them is lower. The worst subchannels may be non-active, due to the total transmitted power constrain.

![Threshold: \(\sum P_i = P_{Tx,Total}\)]

Figure 2-10. Water-filling scheme for CSI at Tx (optimum Tx power allocation).

In a real channel, the actual data rate that is achieved is limited not only by the channel, but also by the transmit and receive signal processing schemes (coding, modulation…) and the antennas configuration, among others. Thus, when the theoretical capacity is calculated, it may be understood as the upper-limit for the system.

**Ergodic capacity, outage capacity**

In a real scenario, instead of deterministic known channels we should considered channels with a certain random distribution. Usually, a variable fading is assumed, which can be modelled as a random variable with a certain distribution. In this situation, two notions of capacity may appear: the ergodic capacity and the outage capacity [57].

The ergodic or Shannon capacity is defined when the transmitted information or codeword is assumed to span an infinite number of independently fading blocks. Roughly speaking, it can be understood as an “average” capacity over the channel, or the capacity when the channel is deterministic.

However, in real applications delay and finite transmitted codewords must be considered. In this situation, and strictly speaking, the Shannon capacity is zero, since no matter how small the transmitted data rate is, there is always a nonzero probability that the given
channel realization will not support this rate. For real systems, then, it is interesting to define the $q\%$ outage capacity $C_{\text{out},q}$ as the information rate that is guaranteed for $(100-q)\%$ of the channel realizations, i.e.:

$$\text{Prob}(C < C_{\text{out},q}) = q\%$$

(2-33)

The outage capacity is a parameter of interest for deployers and operators, since it allows to guarantee a certain system performance with a certain probability %. Usually values of 1\% to 10\% are used in communication systems.

**Capacity and channel normalization**

The calculation of capacity from eq. (2-29) assumes a normalized $H$ channel matrix, that is, no path loss or gain in average is considered. However, in a real scenario there will obviously be path losses which will vary from one location to another or from one type of scenario to another (e.g. LOS or NLOS, indoor or outdoor, etc). As a result, a question arises regarding the issue of channel normalization, especially when measured radio channel matrices are used: how should the normalization of $H$ be computed?

Many researchers have consider an instantaneous normalization of the MIMO channel $H$ (see for example [58]-[60]) when computing channel capacity, method that was first introduced in [61]. Thus, the effect of path loss and slow and shadow fading is removed from the channel matrix. With this assumption, the received power to transmit power ratio is calculated at each snapshot or time instant, averaged over all the Tx-Rx links. This is equivalent to use the Frobenius norm. The mathematical expression for the normalized $H$ matrix is:

$$H(n)_{\text{norm}} = \frac{H(n)}{\|H(n)\|_{\text{Frob}}} = \sqrt{\frac{N_{\text{Tx}} \cdot N_{\text{Rx}}}{\sum_{i=1}^{N_{\text{Tx}}} \sum_{j=1}^{N_{\text{Rx}}} h_{ij}^* h_{ij}}}$$

(2-34)

This method is equivalent to considering a fixed SNR at the receiver (averaged over all the receivers), which is the same as assuming a perfect transmitted power control for each user in the system, as well as unlimited power at the transmitter. It is clear that these assumptions can only be considered an ideal case, but from a system point of view and for algorithm testing purposes it is a widely accepted and used method to compute the system capacity in an easy to compare (although not totally realistic) way.

A modified method to normalize the $H$ matrix of measured MIMO channels was used in [62], where the normalization factor is computed as the Frobenius norm averaged over a sliding window of about 1m (which corresponded to around $7\lambda$ for the presented measurement). This way a more realistic power control is assumed, since the control takes into account a certain time slot, instead of an instantaneous received snapshot.
The previous methods to normalize the channel matrix in a MIMO system are devised from a system point of view, where power control is a natural assumption. However, from a propagation point of view it may also be very interesting to account for the slow and shadow fading, especially when several environments are to be compared. Moreover, also system designers may be interested in these aspects, since different path loss and slow fading may involve different requirements for transmitted power and antenna gains. In this respect, some authors [63] have proposed the normalization of the MIMO channel based on the average received SNR. The transmitted power is fixed to a value so that a certain average SNR is obtained. The idea with this normalization is to obtain a fair comparison between the i.i.d. SISO channel and the realistic (measured or simulated) MIMO channel, by making the median capacity both channels identical. This way, the actual improvement obtained with multiple antennas when compared with conventional SISO systems can be evaluated, and the effect of path loss, slow and shadow fading is also included.

Closely related to the normalization issue is the discussion ongoing among the research community whether a rich multipath scenario (usually NLOS) or a high SNR (usually LOS) one is preferred from the MIMO point of view. As shown in eq. (2-29) both factors contribute to the achievable capacity of the system. In [64] the authors compared LOS and NLOS scenarios with and without channel normalization. A high SNR may imply a low degree of scattering [65], which implies that a trade-off between both factors should be considered. On the other hand, it was shown ([66], [67]) that proper normalization of the measured H matrices reveals the expected variation of capacity with path loss.

In Chapters 4 amd 5 a discussion about channel normalization is given, based on measured radio channel responses. As a preliminary conclusion about normalization issue we may say that different normalization of the measured instantiations of the MIMO channel is appropriate for different MIMO schemes, since both constant transmitted power and constant receive SNR MIMO systems may be considered.

2.3.4 MIMO signal processing: transceiver schemes and algorithms.

As stated previously, one of the aspects the MIMO system performance depends on is the signal processing scheme. If one studies the works on MIMO signal processing and transceiver that have been proposed in the literature so far, one may conclude that most of the work has already been done, given the significant amount of studies, designs and related works that may be found. Nevertheless, there is still a lot of work going on regarding this topic, which is an evidence for some open issues still to be addressed. Although the basis for MIMO transceivers and schemes has already been laid, there are still some aspects to be solved. Attending to their importance for real systems, we may highlight two of them: partial CSI at transmitter (to consider realistic feedback) and optimization for reduced complexity (to assess
real-time implementation). Since the study and evaluation of MIMO signal processing is not one of the main purposes of this thesis, only a brief overview on state of the art of this aspect is given. However, this summary should not be taken as an exhaustive bibliographic investigation.

Classification of MIMO techniques

MIMO techniques can be classified according to multiple aspects. A widely accepted division considers two main groups of MIMO schemes: the ones based on spatial diversity and the ones based on spatial multiplexing.

The spatial diversity techniques aim at taking advantage of available diversity thanks to the use of multiple antennas, in order to mitigate channel fading and to reduce error rate (see Figure 2-11 for a simplified scheme). Multiple copies of the data are sent, with a certain codification or multiplied by beamforming weights. When there is no knowledge of the channel at the transmitter, it seems reasonable to allocate power in a spatially uniform way, as seen in previous section. Well-known spatial diversity MIMO schemes are the Alamouti scheme [68] and the delay diversity scheme [69]. Using the same concept as in the Alamouti codification scheme, which was designed for only two Tx antennas, other codes for higher number of antennas have been proposed. The orthogonal space-time block codes (OSTBC) [70] are a specific family of codes of great interest due to their orthogonality characteristics. As a drawback, the rate that can be obtained with these codes is always lower than 1 (except for the Alamouti scheme, which is a specific case of OSTBC for $N_{Tx} = 2$). Nevertheless, the simple decoder scheme of these codes makes them of great interest for realistic implementation. In order to increase the code rate, quasi-orthogonal space-time block codes (QOSTBC) have been proposed [71], where full-rate designs are obtained to the detriment of orthogonality and slightly increase in the decoder complexity. Another type of codes are the space-time Trellis codes (STTC), firstly presented in [72], which are based on Trellis-coded modulation (TCM) [73]. STTC techniques combine modulation and coding to achieve not only diversity (as STBC) but also coding gain in MIMO systems. The side effect of this additional provision is a serious increase in complexity. Sphere decoders have been proposed as a solution to reduce complexity in STTC schemes.

The space-time coding schemes named previously are just some of the best-known schemes for MIMO systems. Obviously, there are other options, which are not presented here since they are out of the scope of this chapter. An example of open research directions in this
field is the design of STC when the channel matrix cannot be determined at the receiver is one of them, where differential techniques have been proposed. Also of current interest are the codes designed for MIMO-OFDM broadband systems, such as [74] and [75]. More information on space-time coding for MIMO may be consulted in [76].

Although not explicitly, conventional beamforming applied to MIMO systems may be considered a special type of spatial diversity scheme, since the same bit stream is sent through all the Tx antennas, after being multiplied by a certain weight vector. Conventional beamforming may be also used with MIMO systems [77], where the beamforming must be done at both the transmitter and the receiver end (Figure 2-12). Therefore, channel information is required not only at the receiver but also at the transmitter, as opposed to STC that does not require CSI at Tx. Also the eigenvalue decomposition may be used to compute the optimum weights, where the eigenvector associated to the maximum eigenvalue is used (in order to use the best subchannel). Beamforming is of interest when the channel is directive (spatially correlated) and the long term channel statistics are stable. As advantages, beamforming for MIMO shows a low complexity compared with other MIMO schemes, and it shows good performance in scenarios with low SNR and some level of interferences. However, its good properties are not so clear when the channel diversity is high (rich multipath), as assumed for most of other MIMO schemes.

Figure 2-12. Example of beamforming for MIMO. Buildings are depicted in red, relevant paths in yellow and the optimum patterns in blue for 4-element antennas in Tx and Rx.

Unlike spatial diversity, the main idea behind spatial multiplexing (SM) is to break the sequence of information bits into sets of sub-streams to be simultaneously sent by the transmit antennas. Thus, the aim here is to increase data rate, obtaining a capacity close to the upper bound. Moreover, the complexity requirements for SM techniques, while heavy, remains constant and independent of the modulation in use, which is not the case of STTCs. In general, in order to perform SM, the number of receive antennas must be equal to or greater than the number of transmit antennas. The main idea behind SM encoding techniques relies on the use of powerful decoding techniques on the receiver side. The transmitter side is usually quite simple, and consists in dividing the information sequence into several sub-streams through a demultiplexer. The sub-stream where the signal processing is conducted is conventionally called
“layer”, and these technique is referred to as “layered space-time” scheme (Figure 2-13). It was firstly presented by Foschini in [8], where it was named BLAST (Bell Labs Space-Time scheme). The demultiplexing operation can be applied to bits or symbols, depending on the position of the demultiplexing module and the encoder-modulator module. Regarding this, three options are considered for the transmitter scheme: vertical, horizontal or diagonal encoding schemes, which results in V-BLAST [78], H-BLAST and D-BLAST schemes [79], respectively. The vertical scheme requires more complex receivers, but it obtains a better diversity gain. In the horizontal encoding case, the data bits are demultiplexed into $N_{Tx}$ sub-streams that are independently encoded, interleaved and modulated. In vertical encoding SM, the data stream is at first coded, interleaved and modulated and the resulting symbols are then demultiplexed into $N_{Tx}$ sub-streams. Diagonal SM encoding is similar to horizontal encoding, but an interleaver is included at the end of the Tx chains to rotate the transmitted frames. Regarding the receiver scheme, several solutions may be used [80], such as maximum likelihood receivers, linear receivers, MMSE receivers and successive cancellation receivers.

When total CSI at the transmitter is available, it has been shown [56] that the optimum solution is linear prefilter of pre-equalization realized with the Singular Value Decomposition (SVD), or waterfilling scheme (see previous section and Figure 2-10 for an intuitive explanation). It is usually classified as a spatial multiplexing technique, since multiple orthogonal sub-channels are created and several streams of data are sent simultaneously (as depicted in Figure 2-9). A disadvantage of this method is that it requires the transfer of at least the instantaneous matrix $V$ back to the transmitter. Moreover, the SVD involves complex calculation tasks, which requires powerful signal processors. Some sub-optimal versions of the SVD method have been presented in the literature [81] in order to reduce complexity. Also, other sub-optimal methods have been proposed [82] that make use of CSI at the transmitter.

![Figure 2-13. Spatial Multiplexing architecture.](image)

Figure 2-14 summarizes the conventional MIMO techniques depending on the type of spatial method and the channel knowledge at the transmitter.
Up to now, we have considered either the “worst case” of no CSI at Tx or perfect CSI at Tx. However, in real systems the knowledge of the channel at Tx should be considered as partial, due to errors in retransmission, quantization of feedback information and delay. Currently, much attention is being given to the design of MIMO schemes with partial CSI at the transmitter. The main idea is to use the feedback information to realize a certain pre-coding of the signal in the transmitter, as in [83]. Other techniques use the partial information to adapt the MIMO technique, combining beamforming and space-time techniques depending on the spatial correlation of the channel [84]-[86]. In [87], the pre-processing module is designed to allow adaptability for OFDM systems, resulting in the so-called 2D coding-beamformer, which offers good performance at the cost of high complexity. Another precoding scheme is proposed in [88], which uses partial channel information to compensate for the spatial correlation in the channel, and which presents the advantage of low complexity and few required feedback information. Many other theoretical studies with partial CSI have been realized; a good summary may be consulted in [89].

Other signal processing schemes of interest are the antenna selection techniques applied to MIMO systems [90]. While providing good performance, they allow a reduction in system complexity. The selection may consider only one antenna or a set of antennas, and several parameters may be used to make the selection (best achievable capacity, maximum received power…). Other aspects in signal processing for MIMO that are currently generating special interest are techniques for interference cancellation (a complete work on this topic may be found in [91]) and multi-user schemes for MIMO [92] such as opportunistic beamforming [93]. A study of these techniques is out of the scope of this thesis.

### 2.3.5 MIMO channel measurements

Although theoretical studies may be of interest for channel characterization and system design, measurements in real scenarios are required in order to properly characterize the MIMO channel and to validate the proposed channel models. The characterization of wireless channels...
started some decades ago, but it still attracts lots of interest. Regarding MIMO systems, a substantial number of measurement campaigns have been realized in recent years. Different purposes may be devised in a measurement campaign, which also set the type of measurement system or testbed. The initial measurements for MIMO systems reported in the literature were aimed to characterize the radio channel (PDP, K factor for LOS situations, multipath characterization...), including multi-antenna and double-directional features (spatial correlation, PAS,...) for both narrowband and wideband cases. Some measurements were used to validate proposed MIMO channel models. Another use of MIMO measurements is the evaluation of proposed algorithms and transceiver schemes, which may be considered in either off-line or real-time operation of the testbed. Finally, in recent years some system prototypes including MIMO features have been developed or are under development. The evaluation of these prototypes provides the more complete but also the more complex system characterization, and is of great interest from the point of view of manufacturers and system designers.

Multiple features must be considered in the set-up of a measurement campaign, such as type and number of antennas, polarization, frequency of operation, bandwidth or environment. Table 2-II summarizes the most significant characteristics of the experimental set-ups in different measurement campaigns for MIMO systems. Although not exhaustive, this table shows some of the most remarkable measurements performed to characterize MIMO channels in recent years. A good survey on space-time measurement campaigns can be found in [94].

<table>
<thead>
<tr>
<th>Ref.</th>
<th>$N_{Tx} \times N_{Rx}$</th>
<th>Carrier freq. [GHz]</th>
<th>Polarization</th>
<th>Bandwidth [MHz]</th>
<th>Environment</th>
</tr>
</thead>
<tbody>
<tr>
<td>[95]</td>
<td>8×8</td>
<td>5.2</td>
<td>single</td>
<td>120</td>
<td>LOS/NLOS indoor micro-cells</td>
</tr>
<tr>
<td>[96]</td>
<td>16×8</td>
<td>5.2</td>
<td>single</td>
<td>120</td>
<td>LOS/NLOS picocell outdoor</td>
</tr>
<tr>
<td>[97]</td>
<td>3×21</td>
<td>5.8</td>
<td>single</td>
<td>400</td>
<td>NLOS indoor</td>
</tr>
<tr>
<td>[98]</td>
<td>8×8</td>
<td>5.2</td>
<td>single</td>
<td>120</td>
<td>LOS/NLOS indoor</td>
</tr>
<tr>
<td>[99]</td>
<td>4×12</td>
<td>2</td>
<td>single</td>
<td>5</td>
<td>Outdoor urban</td>
</tr>
<tr>
<td>[100]</td>
<td>4×12</td>
<td>2</td>
<td>single</td>
<td>5</td>
<td>Outdoor urban</td>
</tr>
<tr>
<td>[101]</td>
<td>4×4</td>
<td>0.9</td>
<td>single</td>
<td>35</td>
<td>Subway</td>
</tr>
<tr>
<td>[101]</td>
<td>17×9</td>
<td>5.2</td>
<td>single</td>
<td>250</td>
<td>Indoor</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>62.5</td>
<td>Outdoor</td>
</tr>
<tr>
<td>[101]</td>
<td>4×8</td>
<td>2</td>
<td>dual</td>
<td>120</td>
<td>Urban outdoor</td>
</tr>
<tr>
<td>[102]</td>
<td>4×4</td>
<td>2.05</td>
<td>dual</td>
<td>5</td>
<td>Outdoor to indoor micro-cells, indoor pico-cells</td>
</tr>
</tbody>
</table>
Table 2-II. MIMO measurement campaigns and their main characteristics.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>N_{Tx} × N_{Rx}</th>
<th>Carrier freq. [GHz]</th>
<th>Polarization</th>
<th>Bandwidth [MHz]</th>
<th>Environment</th>
</tr>
</thead>
<tbody>
<tr>
<td>[103]</td>
<td>4 × 4</td>
<td>1.9</td>
<td>dual</td>
<td>0.03</td>
<td>Outdoor and Outdoor to indoor suburban</td>
</tr>
<tr>
<td>[104]</td>
<td>12 × 15</td>
<td>1.95</td>
<td>dual</td>
<td>0.03</td>
<td>Indoor</td>
</tr>
<tr>
<td>[105]</td>
<td>10 × 10</td>
<td>2.45</td>
<td>dual</td>
<td>0.025</td>
<td>LOS/NLOS Indoor</td>
</tr>
<tr>
<td>[62]</td>
<td>16 × 64</td>
<td>2.154</td>
<td>dual</td>
<td>120</td>
<td>Indoor picocell, outdoor microcell</td>
</tr>
<tr>
<td>[106]</td>
<td>4 × 4</td>
<td>-</td>
<td>dual</td>
<td>3.5</td>
<td>Urban/suburban, hilly areas</td>
</tr>
<tr>
<td>[107]</td>
<td>2 × 2</td>
<td>2.48</td>
<td>dual</td>
<td>4</td>
<td>Suburban outdoor</td>
</tr>
<tr>
<td>[108]</td>
<td>16 × 16</td>
<td>2.11</td>
<td>dual</td>
<td>0.032</td>
<td>Urban outdoor</td>
</tr>
</tbody>
</table>

The measured bandwidth defines whether the obtained measurements are narrowband or wideband. While first reported measurements were narrowband due to system constrains, several wideband measurements have recently been presented thanks to the development of powerful wideband channel sounders. When wideband measurement equipments are available, the time resolution is better and more multi-path components are time-resolvable. This fact, jointly with the use of multiple antennas and high-resolution techniques for angle estimation, has led to characterization of the MIMO channel multipath behaviour [109], which is of great interest for proper channel modelling. Recently, it has been shown that in addition to discrete, identifiable multipath components, there exist a large number of propagation components that cannot be modelled discretely. They are known as dense multipath components, which are attributed to diffuse scattering, higher order reflections and diffraction. Although some work has been presented in this field [110], more measurements and analysis are required.

We also note that there are several types of measured scenarios, from picocell (ranges up to 100m) to macrocells (ranges of kms). It is interesting to note that most measurements have been realised either for indoor or outdoor environments, but combined outdoor-to-indoor ones are scarce. In chapter 5 we present the results of channel measurements that were carried out in this type of scenario, as well as we motivate its interest in wireless communications and compare our work with previous outdoor-to-indoor measurements reported in the literature.

Polarization is also a key aspect in MIMO channel characterization. Some of the works consider antennas with dual polarization, which allow the analysis of propagation with multiple antennas for both vertical and horizontal polarization. Typical scattering leads to a co-polarized received signal that is 4 to 10 dB higher than the cross-polarized signal level [111]. It has also been found [112] that vertically polarized waves persist longer in certain environments than
horizontally polarized ones. More recently, polarization has been evaluated separately for each
discrete multipath component [113] and as a function of azimuth and elevation [114]. However,
more measurements are required to fully clarify polarization characteristics of MIMO channels
[115].

2.3.6 MIMO channel models

A huge number of channel models for MIMO systems have been developed so far, depending on the application, the complexity we can afford, etc. The parameters that should be included in a MIMO model depends on the purpose of the model: for MIMO system deployment and network planning we need site-specific models, while for MIMO algorithm development and system design and testing a site-independent model is preferred.

The most widely accepted classification of MIMO channel models considers two sets of
models: deterministic models and stochastic models. The first ones are based on an accurate
description of a specific propagation environment. The stochastic models, contrary to the
deterministic ones, do not rely on site-specific description, but they aim at reproducing observed
phenomena by statistical means. Between the purely deterministic models and the purely
stochastic non-physical models there are “mixed” geometrically-based or physical models with
statistical parameters, which combine the best of the two worlds by starting with geometrical
input about the environment and then superimpose statistical information.

Below, the most representative channel models are briefly discussed. Since in
subsequent work the 3GPP channel model is used, we will include a description of this model.
A detailed description of MIMO channel models is out of the scope of this Thesis, and
interested authors should refer to the references presented here and some other good surveys on
channel models such as [116],[117] or [118].

Deterministic channel models

Deterministic channel models rely on detailed representations of propagation conditions
and environments. Therefore, they are very site-specific and the results directly illustrate the
specific physical characteristics. Depending on the purpose of the channel model, this accuracy
may be seen as an advantage (if we require knowing the exact performance in a specific
scenario) or a disadvantage (if we need to design the system parameters for a reference or
typical scenario). Another drawback of these channel models is their complexity and the huge
processing time they require. The two main deterministic approaches towards the
characterization of MIMO channels are the recorded impulse responses and the ray-tracing

Techniques.

The recorded impulse responses are obtained from measurement campaigns. Despite the
numerous measurement campaigns performed so far (see previous section), it is very rare that
the recorded impulse responses are used as such to simulate radio channels, since this method
would suffer from the significant memory resources it requires and the fact that the channel responses are site-specific. Rather, they are used to analyse or extract channel parameters that are used in stochastic models.

Based on geometrical optics, the ray-tracing technique predicts the multipath propagation in a given environment, from its geometrical description (building databases, floorplans…) and its electromagnetic properties. Its main advantage is that it enables to evaluate not only the received signal as the addition of received multipaths, but also the DODs and DOAs are easily analysed. However, realistic simulations involve the consideration of many rays and multiple bounces, which makes this type of channel model to be the most computational demanding. Moreover, they are also tied to the environment that has been considered in the simulation, that is, they are site-specific. Despite this, ray-tracing is a widely used method to model channels, and several examples can be found in the literature, such as [119], [63]. Similarly to recorded impulse responses, the obtained data from ray-tracing modelling may be used to develop stochastic models, as in [120].

**Stochastic Channel Models**

The stochastic models aim at simulating channel characteristics for typical scenarios, instead of for specific environments. Therefore, they reflect the propagation environment in an indirect way. These models often serve to assess the performance of algorithms and transceiver schemes.

We may classify the stochastic models into non-physical ones and physical (or geometrical) ones. The latter ones are based on geometrical characteristics (clusters positions, scatterers, TX-RX distance…) but this information is generated statistically, based on some statistical distribution. Conversely, the non-physical models use other types of information or channel characteristics, such as spatial correlation.

Among the non-physical channel models, we must mention the next ones:

- **i.i.d. MIMO Channel Model.** This is the simplest channel MIMO channel model, since it just assumes that the elements of the MIMO channel matrix are zero-mean i.i.d. complex Gaussian random variables. It is widely used, especially for algorithm testing and design, due to its simplicity and tractability (low computational and memory requirements). However, it neglects all correlation among Tx and Rx signals, which is not realistic for most cases.

- **Correlation-based Kronecker EU-IST METRA model.** It is based on the power correlation or covariance matrix of the MIMO radio channel. The expression of the full covariance matrix is simplified by assuming complete independence between Tx and Rx propagation environments, so the full covariance matrix is expressed as a Kronecker product of transmit and received correlation matrices (hence the term *Kronecker* to refer to it). It was proposed for indoor NLOS scenarios, for narrowband channels [102], and also for wideband channels [121] by introducing
a tap-delay line to model the multipath effect (but keeping the idea of independent Tx and Rx correlations). Although the Kronecker model is quite popular, it does not deliver the expected capacity in certain circumstances, especially for the key-hole effect: the fact of assuming independence between Tx and Rx correlation is equivalent to assuming that multipath components arriving at Rx have no information about the Tx physical conditions. Therefore, this approach neglects the correlation terms across the link (“cross-correlation” or “joint correlation”), which must be taken into account, at least in certain scenarios, as shown in [52].

- **Correlation-based Kronecker EU-IST SATURN model.** It is a model that also relies on the Kronecker product and the independent correlation matrices at Tx and Rx, and was proposed also for indoor NLOS channels. It is based on the statistical characteristics of the measured data, both narrowband [59] and broadband [122]. As in the IST METRA project, its main drawback is its main assumption itself: independent correlation matrices at Tx and Rx can only model separable joint angular power spectra. This model underestimates the channel capacity, especially for indoor environments and large antenna arrays. However, and similar to the IST METRA model, it is quite popular because it allow for independent array and signal processing optimization at Tx and Rx.

- **Weichselberger Model** [123]. This model aims to mitigate the simplification of previous Kronecker models of neglecting the spatial structure of the MIMO channel. The spatial eigenbasis of the receiver (and transmitter) correlation matrix are modelled, but a power coupling between both of them is allowed, thus showing the Tx-Rx correlation dependence. Regarding capacity and diversity results, the Weichselberger model performs better than the purely Kronecker models. Nevertheless, it cannot reproduce the arbitrary multipath structure, and it shows deficiencies when including beamforming aspects to be simulated.

    As discussed above, all non-physical channel models show deficiencies, since they are not based on realistic physical structures and scatterers distribution. To address this, multiple geometrical-based stochastic models have been proposed. The most representative ones are mentioned below:

- **Wideband Directional Channel Model, WDCM** [124], [125]. This model is a geometrically-based one, where the parameters are given for microcells. The scatterers are assumed to be distributed in an elliptical area and single bounces are considered (Figure 2-15, a)). It is quite interesting due to its simplicity and low computational requirements.

- **Geometrical MIMO Channel Model based on SISO parameters** [126]. As the WDCM, it is based on the geometry of the scatterers, but it is more complete (and also more complex) since it includes multiple layers of scatterers (multiple delay ellipses). Double bounces are also included, as well as a circle of local scatterers around the MS. This model was proposed for macrocellular broadband fixed wireless systems. To better represent such environments, two local rings are introduced in this model: a disc of exclusion, representing a scattering-free area.
around the BS, and a smaller circular ring surrounding the CPE including a subset of the scatterers in the first ellipse (Figure 2-15, b)). An extension of this model to include multiple polarization was presented in [127].

![Geometrically-based MIMO channel models with elliptical distribution of scatterers.](image)

Figure 2-15. Geometrically-based MIMO channel models with elliptical distribution of scatterers.

- **Double Directional Channel Model (DDCM)** [128]: it is a parametric stochastic model, where the parameters are computed through spatial scatterers distribution. The multipath characteristics are simulated as tapped delay lines with each multipath characterized with several parameters (amplitude, delay, AoA, AoD…), so angular information is included in both the transmitter and the receiver side. It is a very complete one and has widely been adopted for wideband channel simulations.

- **Extension of Saleh-Valenzuela Model for MIMO**: The Saleh-Valenzuela Model [129] is a widely accepted channel model for wideband SISO communications, and it is based on a clustering phenomenon usually observed in measured radio channels. In order to extend it to MIMO channels, in [130][131] angular information of the clusters was also included. The model is especially suitable for simulation due to its modest computational requirements compared to ray-tracing techniques. As a drawback, it implies the stationarity of the channel statistics and thus it is not suitable for modelling of large-scale movements of Tx and Rx, since they would significantly change the clustering characteristics in the simulation (delays, DoD, DoA…).

- **One-ring [50] and two-ring [132] Channel Models**: in the first model, the base station (BS) is assumed to be elevated and therefore not obstructed by local scattering while the mobile station (MS) is surrounded by scatterers (Figure 2-16, a)). Some simplifications are assumed, such as NLOS between the BS and MS, only one bounce for each ray and the same received power from each ray at the receive array. On the other hand, the two-ring model assumes that both the BS and MS are surrounded by scatterers. This can be the case for indoor wireless communications. An illustration of the two-ring model is shown in Figure 2-16, b). In this model, each ray is reflected twice, thus giving more degrees of freedom. Both of them are quite simple channel models, but still they fail to simulate some channel behaviour such as LOS situations.
a) One-ring model.  

b) Two-ring model. 

Figure 2-16. Illustration of the one-ring and the two-ring models.

- **Distributed scattering MIMO model** [133]: the channel is modelled as a group of scatterers distributed at the transmitter and another one at the receiver, not necessarily in a ring shape (Figure 2-17). Different parameters are selected to define each group of scatterers. It is a channel model for outdoor and narrowband systems.

![Figure 2-17. Scatterers distribution in the Distributed Scattering MIMO model](image)

- **General MIMO channel Model based on Geometrical description** [134]: several aspects are included in this model (far clusters, double scattering, waveguiding) so the author claims that different behaviours in measured channels that other channel models are not able to simulate (such as keyhole effect) are addressed in this channel model.

- **COST 259 channel model and its extension in COST 273**: A directional channel model developed by the European research initiative COST 259 was reported in [135] and can be used to model different MIMO propagation channels. It is a geometrical-based model and includes double scattering, near and far scatterers. A new model, including novel features such as simulation of placement of clusters and interacting objects (IOs), has been developed by COST 273 as the successor of the COST 259 channel model, and a Matlab implementation is available at [136]. However, we note that any long term evolution is omitted from this Matlab implementation of the COST 273 MIMO model. The calculated impulse responses for all antenna combinations are valid for short movements only. For calculating the MIMO impulse response the plane-wave assumption is used. Also note that the current version does not support dual polarization.
· **EM Scattering model** [137]: it is a physical MIMO model based on Electro-Magnetic (EM) considerations. Both the properties of the channel and the antennas were taken into account, by including a term to model the antenna polarization and pattern. One disadvantage of this model is its complexity and the difficulty to fix some parameters. Also, if the scatterers are too close to the BS or MS, the near field effect should be considered.

· **Virtual channel model** [138]: it can be considered more a method for describing the effect of the channel on specific systems. It describes the channel transfer matrix in a beamspace whose resolution depends on the antenna configuration. It is thus more a system analysis tool than a propagation model, but may be of interest to design transceivers and algorithms.

Many other channel models have been presented in the literature, some of them derived from the ones mentioned above, and some others in order to cope with special behaviour or circumstances in the environment. We would like to mention two channel models that are part of standardization effort: the 802.11n channel model and the 3GPP-3GPP2 channel model.

· **The 802.11n channel model** [139] is the result of the research effort to propose a channel model for indoor WLAN systems. It is based on the cluster modelling approach, where tap-independent and cluster-dependent angular and power properties are characterized. The concept of antenna correlations at the transmitter and the receiver (as in the Kronecker model) is also used here, although the cluster characterization allows for more realistic results. Polarization issues are not included in this model. A Matlab version of this model is available at [140].

· **The 3GPP-3GPP2 channel model** [10]: it is a serious attempt to unify the propagation modelling approach for space-time models started a few year ago with the European Scientific Action COST 259, which received even more attention when used as the basis for the channel model adopted by the standardization body 3GPP for 3G communications. A Matlab implementation of the model was developed in the frame of the WINNER project and it is available at [141]. The scope in this model was to develop specifications for both system-level evaluation and link-level evaluation. The last one should be used only for calibration purposes. The model is a 2D geometrically-based one, and includes multiple options such as consideration of LOS, single or double polarization, possibility to include the antenna radiation pattern, etc. However, it does not take into account the 3D components in a real channel; a new channel model based on the 3GPP-3GPP2 one including 3D components was proposed in [142]. However, the main drawback of this model is the difficulty to know for a certain scenario some parameters that were proposed to model the proportion of 2D or 3D components (g parameter in the model), so it is difficult to compare it with measurements. The 3GPP-3GPP2 model will be further presented with more detail in chapter 6, since it was used there to evaluate different antenna arrays for MIMO.
After the previous description of the existing channel models, we may think that the research work of modelling the MIMO radio channel is almost finished. On the contrary, there are still some open issues to be solved, which explains the current intensity in work on this field. One of the current lines of interest is the inclusion of polarization in the MIMO channel model. Although it has been considered in some of the channel models presented up to now (see for example [102], [131], [137], [10], [142]), more theoretical work and channel measurements to validate it are required to give a better insight into this topic. Also of great interest is to account for 3D components instead of only 2D ones as conventionally done. 3D modelling of the channel may be of great importance for indoor scenarios, where small distances and richness of scatterers may give more importance to elevation components. Nevertheless, the works on this area are scarce, being [142] and [143] two of the few relevant examples. Finally, the current interest that is generating the antenna perspective in MIMO systems is pushing the introduction of the antenna effect in the MIMO channel models, but there are still few of them that include the antenna parameters. The ones that include some antenna parameters (see for example [137], [139], [10]) usually simplify the antenna element by allowing simulation of only linear arrays, ideal radiation patterns or not considering coupling among elements. A more complex study on the antenna effect and its introduction in widely used MIMO channel models is an open issue.

2.3.7 Antenna aspects in MIMO systems

Typically, MIMO systems have been studied from the information theory point of view (“system people”). It’s only recently that the MIMO research community is beginning to pay more and more attention to the antenna effect and its optimum design for these type of systems, from an antenna and propagation point of view (“electromagnetic and antenna people”). It is known that the performance of MIMO systems do not only depend on the transceiver scheme and related signal processing behind the antennas, but also on the antenna characteristics and the multipath structure of the channel, hence the interest in studying the antenna configurations and designs. Since some of the chapters for these thesis deals with antenna aspects in MIMO, in this section we give an overview of recent advances on antenna aspects for MIMO systems.

Several topics are included in this research area. We will focus on the ones that we considered as key aspects when studying antennas: the antenna and its electromagnetic model, the impact of antenna and array configuration in MIMO systems, design of antenna arrays for MIMO and finally methods and parameters to evaluate antennas for MIMO.

The antenna from an electromagnetic point of view

In the signal processing and system research area, it is quite common to find that the antenna is treated as an ideal sensor, especially when the work is focused on design of algorithms. Also when channel model and measurements are addressed, the antenna effects are
usually assumed as negligible, or a post-processing is realised to reduce their effect. However, in a real system the antenna plays an important role: it is the element that allows the transition from the (guided) generated signal to the radio channel or transmission medium. Thus, assuming that the antenna elements are ideal omnidirectional ones, with no coupling, no losses and ideal matching (Figure 2-18, a)) will lead to non-realistic results.

From an electromagnetic point of view, the radiant element is characterized by its radiation and circuit features, such as the amount of signal that is it able to transmit (receive) when it is excited with a certain currents distribution (impinging electromagnetic wave, in Rx), and the spatial directions where the transmitted energy is mostly confined (or where more energy is captured, in Rx). When multiple elements are used, there are also effects such as coupling among elements, which may vary the element characteristics. Figure 2-18 compares the “antenna as an ideal sensor” representation with a realistic antenna representation for MIMO.

Regarding radiation characteristics, one of the most important parameters to characterize the antenna is the radiation pattern, $F$. It is defined as the radiation characteristic of the antenna (amount of radiated fields, in amplitude and phase) as a function of spatial coordinates (usually spherical coordinates, azimuth $\phi$ and elevation $\theta$, are used), typically normalized to the maximum value. It is important to note that it is a 3D parameter, and not a 2D parameter as considered in some simplified studies (for example, in some cases only the horizontal plane is considered, as in the channel model from 3GPP [10]). Also note that for each spatial direction the full information about radiation is given by two components, being each one a complex number (thus, with a certain amplitude and phase):

$$
F(\theta, \phi) = F_\phi(\theta, \phi)\hat{u}_\phi + F_\theta(\theta, \phi)\hat{u}_\theta = F_{CP}(\theta, \phi)\hat{u}_{CP} + F_{XP}(\theta, \phi)\hat{u}_{XP} \quad (2-35)
$$

where $\hat{u}$ is an orthonormal vector, and the expression on the right is the radiation characteristic expressed as copolar and crosspolar components. For a properly designed antenna, the crosspolar component should be much lower than the copolar one.
The antenna can also be characterized from a circuital point of view. In this case, the equivalent circuits for the antenna in Tx and Rx are the ones shown in Figure 2-19. The antenna is represented as an impedance $Z_i$, whose real part is given by two components:

$$Z_i = R_i + jX_i = (R_{\text{rad}} + R_{\text{loss}}) + jX_i$$  \hspace{.5cm} (2-36)

Each term in $R_i$ has a physical meaning: part of the energy delivered to the antenna $P_{\text{ant}}$ is not radiated due to losses in the antenna (represented by $R_{\text{loss}}$) and the rest of the energy is radiated (represented by $R_{\text{rad}}$). The relationship is usually called the radiation efficiency:

$$\eta_{\text{rad}} = \frac{P_{\text{rad}}}{P_{\text{ant}}} = \frac{R_{\text{rad}}}{R_{\text{loss}} + R_{\text{rad}}}$$  \hspace{.5cm} (2-37)

Some of the power at the input of the antenna is not delivered to the antenna, due to impedance mismatch, which is usually measured as the reflection coefficient $\rho$ or return losses $(20\log_{10}|\rho|)$

$$\rho = \frac{Z_{\text{in}} - Z_0}{Z_{\text{in}} + Z_0}$$  \hspace{.5cm} (2-38)

When losses due to impedance mismatch are also considered, the efficiency may be called antenna efficiency and is defined as:

$$\eta = \frac{P_{\text{rad}}}{P_{\text{in}}} = \frac{G_0}{D_0}$$  \hspace{.5cm} (2-39)

which also relates gain $G_0$ (where impedance mismatch has been considered) and directivity $D_0$.

When multiple antenna elements are closely placed, another effect appears: the coupling among elements. This effect has been studied for some years in array applications, such as radar and antenna arrays for base stations in mobile communications. It is known that due to mutual coupling among elements of an array, the radiation pattern and input impedance for each element varies from the ones that would be obtained for an isolated element. Therefore, it is of great interest to study the impact of coupling in MIMO systems and try to take advantage of it, as we will see in next subsection.
Impact of antennas on MIMO systems

The effect of antenna characteristics and array configuration in MIMO systems is currently a hot research topic. Multiple issues can be considered when studying the impact of antennas, such as the element radiation pattern, the array configuration, the element polarization and mutual coupling among elements.

The element radiation pattern changes the observed MIMO channel. Thus, $H$ does not only depend on the physical structure of the channel (number and position of scatterers, etc) but also on the antennas in the system. In some cases, the antenna elements are designed in order to obtain a radiation pattern as omnidirectional as possible, so all generated multipaths in the channel can be exploited. Contrary to this strategy, the antenna elements can be designed to achieve a high angle or pattern diversity. This results when antennas have distinct radiation patterns, so the correlation among transmitted (or received) signals is small even if the channel does not offer a high uncorrelated scenario. When the antenna phase centers are closely spaced so that angle diversity is the dominant factor, or when the channel does not offer much multipath richness, the essentially omnidirectional pattern created by most small elements results in relatively large values of correlation, leading to low capacity. However, when element patterns are appropriately designed to minimize correlation among transmitted or received signal, then capacity gains are possible. One suggested approach for realizing such a situation involves the use of multi-mode antennas where the patterns for different modes exhibit high orthogonality (low correlation) [144],[145]. Another option has been proposed in [146], the so-called switched parasitic antenna. Finding other antenna topologies that offer this orthogonality in a compact form factor remains an area of active research. Another important aspect of element radiation pattern involves the manner in which the antenna excites the multipath environment. As an example, a recent study [147] has compared the capacity performance of dipole antennas with that of higher-gain spiral antennas whose radiation patterns tend to be more directive toward 45° and 135° in elevation. The results show that the lower gain dipoles offer superior capacity (by about 10%), since these antennas put more energy into the horizontal plane where most of the multipath components are concentrated.

The array topology and configuration also influences the observed channel matrix $H$. The question here is which array topology is best in terms of maximizing capacity (perhaps in an average sense over a variety of propagation channels) or minimizing symbol error rates. This is difficult to answer definitively, since the optimal array shape depends on the site-specific propagation characteristics, although some general observations are possible. Many studies have been done in this area, including simulations and measurements. Some works have focused on studying linear array configurations, considering orientation [148], element spacing [149], or type of elements [150]. Other works compare different array configurations in MIMO scenarios [62],[151]-[153]. Finally, we observe that for single antenna systems, the capacity bound is independent of the antenna (other than the antenna gain). The strong dependence of MIMO
capacity on array configuration is therefore troubling since this number is not a true upper bound on the physical channel throughput. In response to this, recent work has formulated the Intrinsic Capacity for a specific channel and spatial antenna apertures independent of the array configuration [154]. This creates new research avenues in identifying antenna elements and arrays that provide optimal or near-optimal performance.

The element polarization is also a very interesting characteristic to be studied in MIMO systems. Recent work [155] has suggested that in a rich multipath environment, sensing the three Cartesian vector components of the electric and magnetic fields can provide six uncorrelated signals at the receiver. This has boosted the interest in studying multi-polarized MIMO channels, as well as antenna arrays which include multiple polarizations. Moreover, the use of multi-polarized antennas offers the advantage of reducing the array size without necessarily meaning an increase in signal correlation, at least not as much as for single-polarized antennas. Some channel measurements have been presented in the literature where dual-polarized antennas are used (see section 2.3.5 and Table 2-II for a summary). From a practical standpoint, constructing a multi-polarized antenna that can exploit all the polarization diversity is problematic. In [156] a 3-element array is proposed for a mobile-type user terminal, consisting of three monopoles. However, using \( \lambda/4 \) monopoles, halfwavelength dipoles and full-wavelength loops leads to strong mutual coupling and nonideal pattern characteristics that can reduce the number of independent channels. One interesting geometry is a cube consisting of dipole antennas to obtain a high-degree of polarization diversity in a compact form [157]. The design and fabrication of other practical antennas that can exploit polarization remains an area where further research is warranted.

Finally, antenna mutual coupling is a key issue of concern for MIMO systems. In one hand, it can be considered a shortcoming in closely spaced antenna arrays, since it produces impedance mismatching and thus it reduces the antenna efficiency, as well as a distortion in the radiation pattern. On the other hand, this same pattern distortion can be used to create pattern diversity that can lead to reduce signal correlation [158]. More comprehensive studies have examined the effect of coupling and antenna termination on the capacity [159], [160] of MIMO. A complete network analysis is given in [161] for coupling in MIMO systems and its effect in capacity. Two main conclusions have surfaced as a result of these studies. First, because of the induced angle diversity combined with improved power collection capability of coupled antennas, the capacity of two coupled dipoles can be higher than that of uncoupled antennas (through proper termination), particularly for small dipole spacing where coupling is high [161]. Second, for a fixed-length array, the strong coupling between elements packed into the same physical space will ultimately lead to an upper bound on capacity performance, even if the number of antennas is increased. Despite these theoretical studies, few works with antenna implementation and coupling and capacity measurements can be found, being one of the few examples [162]. A method to reduce the negative effect of mutual coupling on impedance
mismatching in MIMO and the design of antennas for optimum coupling among elements (aiming at maximum capacity) is still an open issue.

**Design of antennas for realistic terminals in MIMO**

As seen in previous paragraph, many works have been done on studying the antenna effect in capacity and performance for MIMO systems. However, in order to deploy MIMO technology in real radio communication systems, a further step is necessary: design of antenna arrays with multiple elements for realistic user equipments. While the use of multiple antennas in the base station or access point (AP) is usually feasible, user terminals have size and weight restrictions that make the use of conventional antenna elements such as dipoles or patch antennas problematic. Thus, novel array topologies and antenna elements for multi-antenna systems are of great interest. An interesting work is the one presented in [163], where monopole elements are mounted on a card-type ground plane and the array features regarding MIMO communications are studied. This configuration is proposed for laptop-type user terminal. Another proposed antenna array for laptop terminal is the one used in [108], where dual polarized patches are situated in the laptop upper face. Compact configurations [164] are also very interesting and are currently an active research activity is being carried out. It has been proposed the use of Planar Inverted-F antennas (PIFA), due to their small electrical size. An example is shown in [165], where a 2-element PIFA is used as antenna array in Tx and Rx and its performance is compared with dipoles. More designs for different terminal equipments are an open research area. Finally, also other options such as multimode antennas [145] or reconfigurable antenna arrays with microelectromagnetic systems have recently been proposed.

**Evaluation of antennas for MIMO systems**

The analysis and evaluation of antennas for communication systems has been studied for many years. Currently, multiple standards and recommendations establish several aspects about antenna measurement and evaluation, such as the IEEE standard 149-1979 [166]. There are well-known parameters to characterize antennas in general, such as the radiation pattern, the input impedance, the antenna efficiency and so on.

When trying to extend the evaluation of antennas to MIMO systems, it is not so straightforward. While the parameters to characterize antennas in general are well defined and worldwide accepted, the way to evaluate the performance of an antenna array for MIMO is still an open issue, since multiple new concepts have to be considered, such as pattern diversity, correlation among elements or polarization diversity. Several parameters have been proposed to characterize antennas for MIMO systems. In a recent study [167] some parameters are analysed to characterize antennas in fading environments. When single antennas are analysed, the **mean effective gain** (MEG) is an interesting parameter. It can be calculated from the radiation field function of the antenna, and it is a function of the orientation of the antenna, its polarization and
the azimuth, elevation and polarization distributions of the incoming waves in the environment. When multiple elements are used, the MEG may be extended to the **mean effective link gain (MELG)** [168], which accounts for all the combination of Tx-Rx links. Regarding radiation characteristics, when only one element is considered we may measure the **isolated element radiation pattern**. However, for multiple elements the measured parameter would be the **embedded** or **active element pattern** [167]. This is the radiation pattern of a single element when all the other elements are present, but are not excited and instead terminated with loads representing the source impedance on their ports. The embedded element pattern is used to describe blindness in classical arrays, whereas in MIMO antennas it plays an even more significant role: it gives information about the pattern diversity in the array (related with the correlation among signals).

Of great interest in MIMO systems is the spatial diversity offered by the jointly effect of the channel and the antennas. However, characterizing this parameter is not a straightforward task: depending on the channel structure the optimum antenna features may vary. In [167] the **apparent**, **effective** and **actual diversity gains** for an antenna array for MIMO applications are defined, depending on whether we use a reference one of the branches of the diversity antenna, an ideal single antenna or the existing practical antenna under test.

Finally, the achievable **capacity** for a MIMO scenario when using the array under test is the ultimate parameter to be characterized. However, it is not clear how to measure it. The more realistic but also more complex way to obtain the achievable capacity with a certain antenna array would be by conducting channel measurements. Nevertheless, this method requires the use of a channel sounder, as well as been time-consuming. Moreover, it is not clear which scenarios should be used as reference ones, since it depends on application. Besides, the measurement conditions are hardly repeatable, so it is hard to compare different antenna arrays. The use of a reverberation channel has been proposed as an optional way to evaluate antenna arrays for MIMO systems [167], with the advantage of allowing repeatable measurement conditions. Recently, a method based on the measured radiation patterns [168] has been proposed to evaluate diversity, which has been called the plane wave based method (PWBM).

A thorough study and comparison of some of these methods will be given in chapter 6, based on the comparison of characteristics for two antenna arrays in MIMO. Moreover, other parameters of interest to characterize the antenna arrays for MIMO are proposed and computed.

### 2.3.8 Prototyping for MIMO systems

The increasing interest in MIMO techniques has led to a prolific work in MIMO channel sounders and testbeds. Some of them aim at measuring the channel characteristics as accurately as possible, while the objective of others is the evaluation of the overall system by
developing the whole multi-antenna system, including the signal processing algorithms, in a real-time operation. Multiple prototypes classifications may be considered, attending to different characteristics. In next paragraphs we will describe the main features to be considered when developing a MIMO testbed. Moreover, an overview on the most remarkable existing prototypes is given.

**General considerations about MIMO prototypes**

There are many aspects to be considered when characterizing and classifying MIMO platforms; the main ones are summarized below:

- **Type of environment which the prototype is intended for:** indoor or outdoor, stationary or dynamic.
- **Wireless standard** (if any is considered), which entails a certain carrier frequency, bandwidth, modulation, etc.
- **Operational basis:** real-time or offline, burst-like or continuous processing.
- **Level of implementation:** measurement device, demonstrator, prototype, commercial product.
- **Hardware (software radio platform):** DSP, FPGAs, ASIC…
- **Number of antennas** in transmission and reception.
- **Implemented MIMO algorithms,** and whether there is feedback information available at the transmitter or not.
- **Single or multiple users considered.**
- **Unidirectional or bidirectional data traffic** (simplex, semiduplex or full duplex).

Depending on the objective and the application the MIMO prototype is intended for, the may present different features. In next section a summary of the main MIMO prototypes presented in the literature, as well as some of the characteristics, is given.

**Review of MIMO testbeds and prototypes**

The number of testbeds and MIMO prototypes that have been or are being developed is immense, and a detailed explanation of all of them is out of the scope of this document. Therefore, only an overview of the main ones is given here. For interested readers, the first part of [171] summarizes an interesting study about the topic. One of the main conclusions that were drawn from it was that the multi-user scenario and the feedback channel possibility are seldom implemented in current testbeds and prototypes. In chapter 4 of this thesis a novel MIMO testbed will be presented, and the main differences and advantages compared to other existing prototypes will be highlighted there.

Initially, several MIMO prototypes were built for no specific standard or regulation. One of the first MIMO indoor experiments found in the literature is the real-time laboratory
prototype reported in [78]. The working frequency was 3.65 GHz, and not specific modulation scheme was used (from BPSK to 64-QAM were implemented). Although the presented system was narrow-band, it represented a significant achievement at that moment, since it demonstrated that the MIMO concept could actually increase system performance.

After this first step in MIMO prototyping, several other platforms were developed to test and evaluate MIMO concepts. An obvious research topic is to measure and analyse the MIMO channel for different scenarios. In order to gain insight into this topic, several MIMO channel sounders have been developed and different measurement campaign has been reported (see section 2.3.5 for a summary of the most relevant ones).

Another significant aspect to be evaluated is the algorithms performance in real MIMO systems, leading to the study of specific standards. One of the most prolific topics has been prototypes showing different characteristics and performance of MIMO schemes for the mobile Third Generation, 3G. A prototype that was integrated with a base station for UMTS systems is reported in [172], capable of outdoor transmission and supporting a $4 \times 4$ antenna system in UMTS standards, which allows for up to 1 Mbps downlink capacity.

Recently, many research activities in this topic have focused on wireless LAN standards and applications. A straightforward implementation consists in an offline mode of operation, which allows to implement the MIMO algorithms and most of the baseband signal processing in a PC-based environment. This operation mode simplifies the algorithms testing, since no real-time constrains have to be fulfilled. In [173] a $2 \times 2$ MIMO prototype based on this idea is reported. Although not specifically designed for WLAN standard, a 20 MHz OFDM signal is transmitted. The signal is pre-calculated, loaded in the transmitter memory and afterwards sent to the transmitter module, in an offline basis. These data are repeatedly transmitted through the MIMO channel. The receiver collects the raw data, which is post-processed after being sent to a PC connected to the receiver module. This method of operation simplifies the system implementation and allows an easy way of testing algorithms. Nevertheless, this testbed does not include real-time performance evaluation or feedback analysis. Several other testbeds have been presented [174]-[177], following the same idea of pre- and post-processing in an offline basis, with different levels of complexity in the communication between PC and testbed, as well as different features (carrier frequency, bandwidth, etc.). The MIMO testbed presented in this thesis (Chapter 4) may be considered into this type of prototypes.

Another interesting testbed for MIMO system is the one developed in the Virginia Polytechnic Institute (MCMS testbed) [178]. Allowing up to $4 \times 16$ antennas, it has the advantage of a huge tuning range from 250 MHz to 6 GHz, as well as a bandwidth of up to 40 MHz. An example of “over-the-air” signal collection and processing for a real 802.11b wireless LAN with this prototype was presented.
A further step in MIMO prototypes is the one being developed at Georgia Tech [179]. The system has been designed for the 802.11a and 802.16 standards (WLAN and WiMax standards), with up to 4 antennas at transmitter and receiver side. Although the prototype does not fully work in real time, some of the digital signal processing modules (FFT, synchronization, frequency offset estimation and correction, channel estimation) are implemented in FPGAs and process the data in a real time basis. The remaining signal processing is performed off-line in a PC (detection, MIMO algorithm, etc.). Only preliminary results for this prototype have been reported. A similar idea has been implemented with the so-called “STARS” prototype [180], where a very flexible platform is used to develop a MIMO testbed that can be configured for multiuser scenarios.

As an improvement of offline prototypes, some real-time implementations for general purpose have been reported. Following the same approach as the one presented in [78], a 2x2 real-time MIMO system called VT-STAR, based on DSP-processing, is reported in [181]. A flexible architecture and easy implementation thanks to the use of SDR development platforms are the main features of the system. However, the supported data rate is very low due to the low signal processing capability of the system.

Few MIMO prototypes with feedback capabilities have been reported. An example can be found in [182], where a real-time implementation of a space-time scheme with wired feedback is presented. Although the system bandwidth is very small (only 9.6 KHz) due to hardware constrains, the results for adaptive modulation and MIMO algorithms are very interesting as a first study in prototypes with feedback possibility.

Some real-time prototypes are currently being developed, most of them for WLAN application. The first iteration of a MIMO Multi-user prototype is presented in [183]. Up to 8 antennas are considered for the final design, although the current system consists of 2 antennas at transmitter and receiver. The design and first steps of a 4 x 4 MIMO prototype focused on an extension of the 802.11a standard is presented in [184]. Some of the signal processing modules (such as Viterbi decoder) are implemented in ASIC, which results in a more complex development and less flexible implementation, but allows a higher performance, whereas most of the signal processing modules are intended to be implemented in FPGAs. Other interesting examples of MIMO prototypes can be found in the recently published special issue [185].

Very few commercial prototypes are so far available, to the best knowledge of the author. Furthermore, the existing ones may in fact be considered as general purpose platforms for MIMO testing, measuring and developing of algorithms. A very interesting example is the HaLo prototyping platform, from Signalion [186], especially suitable for algorithms design and testing in multi-antenna wireless systems. It consists in a flexible and scalable hardware platform that includes some of the basic modules to be used in a real-time implementation (such as acquisition, frequency and time synchronization), an easy-to-use interface with
Matlab/Simulink flow, and several reference modules for the simulation of the physical layer of 802.11a/n standards, among others. A very interesting example of a 4 x 4 MIMO-OFDM implementation using the HaLo prototype was presented in [187].

Table 2-III summarizes the features of the prototypes presented in the above paragraphs.

<table>
<thead>
<tr>
<th>Ref</th>
<th>Year</th>
<th>$N_{Tx} \times N_{Rx}$</th>
<th>Environment</th>
<th>Wireless standard</th>
<th>Op. basis</th>
<th>Single or multi-users</th>
<th>Simplex or duplex</th>
<th>Carrier freq (GHz)</th>
<th>BW (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[78]</td>
<td>1998-99</td>
<td>8x12</td>
<td>Indoor</td>
<td>-</td>
<td>real-time</td>
<td>single</td>
<td>simplex</td>
<td>1.9</td>
<td>0.03</td>
</tr>
<tr>
<td>[172]</td>
<td>2001-03</td>
<td>4x4</td>
<td>Outdoor</td>
<td>UMTS</td>
<td>real-time</td>
<td>single</td>
<td>simplex</td>
<td>2</td>
<td>3.84</td>
</tr>
<tr>
<td>[173]</td>
<td>2001</td>
<td>2x2</td>
<td>Outdoor &amp; indoor</td>
<td>-</td>
<td>offline</td>
<td>Single</td>
<td>simplex</td>
<td>3.65</td>
<td>20</td>
</tr>
<tr>
<td>[174]</td>
<td>2003</td>
<td>8x8</td>
<td>Indoor</td>
<td>-</td>
<td>offline</td>
<td>single</td>
<td>simplex</td>
<td>2.4</td>
<td>16</td>
</tr>
<tr>
<td>[175]</td>
<td>2004</td>
<td>4x4</td>
<td>Indoor</td>
<td>-</td>
<td>offline</td>
<td>single</td>
<td>simplex</td>
<td>2.45</td>
<td>1.9</td>
</tr>
<tr>
<td>[176]</td>
<td>2005</td>
<td>4x4</td>
<td>Indoor</td>
<td>-</td>
<td>offline</td>
<td>single</td>
<td>simplex</td>
<td>1.67-2.45</td>
<td>15</td>
</tr>
<tr>
<td>[177]</td>
<td>2004</td>
<td>3x3</td>
<td>Indoor</td>
<td>802.11a</td>
<td>offline</td>
<td>single</td>
<td>simplex</td>
<td>5</td>
<td>20</td>
</tr>
<tr>
<td>[178]</td>
<td>2005</td>
<td>4x16</td>
<td>Indoor</td>
<td>-</td>
<td>offline</td>
<td>single</td>
<td>simplex</td>
<td>0.25-6</td>
<td>40</td>
</tr>
<tr>
<td>[179]</td>
<td>2004</td>
<td>4x4</td>
<td>Indoor</td>
<td>802.11, 802.16</td>
<td>real-time, offline</td>
<td>single</td>
<td>simplex</td>
<td>-</td>
<td>27.3</td>
</tr>
<tr>
<td>[180]</td>
<td>2004</td>
<td>4x4</td>
<td>Indoor</td>
<td>-</td>
<td>real-time, offline</td>
<td>multiple</td>
<td>semi-duplex</td>
<td>2.4</td>
<td>20</td>
</tr>
<tr>
<td>[181]</td>
<td>2001</td>
<td>2x2</td>
<td>Indoor</td>
<td>-</td>
<td>real-time</td>
<td>single</td>
<td>simplex</td>
<td>2.050</td>
<td>0.75</td>
</tr>
<tr>
<td>[182]</td>
<td>2005</td>
<td>2x2</td>
<td>Indoor</td>
<td>-</td>
<td>real-time</td>
<td>single</td>
<td>simplex</td>
<td>1.8</td>
<td>0.01</td>
</tr>
<tr>
<td>[183]</td>
<td>2004</td>
<td>8x8</td>
<td>Indoor</td>
<td>-</td>
<td>real-time</td>
<td>multiple</td>
<td>duplex</td>
<td>10.525</td>
<td>25</td>
</tr>
<tr>
<td>[184]</td>
<td>2004</td>
<td>4x4</td>
<td>Indoor</td>
<td>802.11a</td>
<td>real-time</td>
<td>multiple</td>
<td>simplex</td>
<td>2.4, 5.2</td>
<td>20</td>
</tr>
<tr>
<td>[187]</td>
<td>2004</td>
<td>4x4</td>
<td>Indoor</td>
<td>802.11a</td>
<td>real-time, offline</td>
<td>single</td>
<td>semi-duplex</td>
<td>2.4-2.25, 5-5.825</td>
<td>20</td>
</tr>
</tbody>
</table>

Table 2-III. Characteristics of some relevant existing prototypes.

2.3.9 Summary on state of the art in MIMO: what is new, what is lacking?

In this section, we summarize the main open issues and current active research topics, with special emphasis in the ones that are addressed along this thesis.

Regarding channel characterization, a paramount issue that is under investigation now is the characterization of multipath components and diffuse scattering. The double-directional characterization of discrete multipath components in the radio channel, as well as the dense multipath components and their polarization behaviour are of great importance and still an area to work in. It is closely related to MIMO radio channel modelling. Also in the same research topic, identification of scatterers and clusters is an active research area.
Another question to be answered is the preference for rich multipath or received power. As previously shown, most works on MIMO systems assume fixed SNR at the receiver, which involves that the receiver should be “hidden” from the transmitter in order to get a richer multipath scenario. However, the received power will be lower for those cases, so the specifications for maximum available power at the transmitter should be higher.

Polarization is an attractive property of electromagnetic fields to be exploited in MIMO systems, and its study in many of the MIMO aspects (modelling, signal processing including it, multipolarized antennas) is currently of paramount importance. More measurement campaigns to better characterize the MIMO channel regarding polarization are desirable, as well as new channel models including polarization features in novel aspects as disperse multipath components. Some initial work has been done regarding the effect of polarization in existing MIMO algorithms, but novel algorithms that take advantage of polarization are future lines of investigation. Also in the antenna module the polarization is a key issue to be studied. So far much research work has been done by using conventional double-polarized antennas such as dipoles or patch antennas; the way to optimally include polarization diversity in a compact design of the array is an open research area.

Regarding the signal processing modules, a key aspect under current investigation is the type of information to be used at the transmitter. While the best performance would be obtained with instantaneous channel information, its difficulty (if not impossibility) to be obtained in real systems leads to studying the optimum way to exploit channel information in average, or other second order statistics of the channel.

As previously stated, more research is needed in antenna aspects for MIMO systems, such as the design of compact antennas that include pattern diversity, or the optimum way to obtain polarization diversity to achieve maximum capacity. Moreover, the way to evaluate antennas for MIMO is still not clear, and more work in this line is also of great interest.
# Adaptive Antenna for Mobile Systems

Design, implementation and evaluation of signal processing modules

## 3

### 3.1 Introduction and motivation ................................................................. 70

### 3.2 Beamforming and 3G, theoretical study ............................................ 73

3.2.1 Introduction ......................................................................................... 73

3.2.2 Analysis and theoretical study of beamforming methods .................... 73

3.2.2.1 Conventional adaptive algorithms: LMS (and NLMS) and RLS ...... 74

3.2.3 Beamforming for 3G systems: advantages and drawbacks .................. 83

### 3.3 An adaptive antenna for 3G systems: general description and operation .. 84

3.3.1 Introduction ......................................................................................... 84

3.3.2 System operation: the “plug and play” concept .................................. 84

3.3.3 General description and main features .............................................. 87

3.3.4 Simulation results and expected performance ..................................... 93

### 3.4 Signal processing modules: design and implementation for real-time operation ................................................................. 96

3.4.1 General architecture of signal processing modules .............................. 96

3.4.1.1 SDR concept .................................................................................... 96

3.4.1.2 Modular diagram ............................................................................ 97

3.4.2 Hardware platform ............................................................................. 98

3.4.2.1 Selection of SDR platform ............................................................... 98

3.4.2.2 Overall hardware architecture ....................................................... 105

3.4.3 Signal processing modules for adaptive antenna ............................... 107

3.4.3.1 General synchronization sub-module ............................................ 107

3.4.3.2 User data synchronization sub-module .......................................... 109

3.4.3.3 Demodulation in uplink and downlink .......................................... 111

3.4.3.4 Adaptive beamformer ................................................................. 112

3.4.3.5 Digital downconversion and ADCs ............................................. 118
3.4.3.6  Digital upconversion and DACs....................................................... 119
3.4.4  Code optimization and load distribution ............................................. 120
  3.4.4.1  Maximum number of instruction per DSP ......................................... 120
  3.4.4.2  Evaluating computational load of each module .............................. 121
  3.4.4.3  Code optimization and computational load of adaptive antenna modules 121
  3.4.4.4  Load distribution in DSPs ................................................................. 123
3.4.5  Analysis of impairments due to non-ideal effects and proposed solutions for their mitigation................................................................. 124

3.5  Novel methodology to evaluate adaptive antennas............................... 130
  3.5.1  Introduction ...................................................................................... 130
  3.5.2  Evaluation under controlled scenario: 2-steps measurements in an anechoic chamber .................................................................................. 130
  3.5.3  Evaluation under realistic scenario ..................................................... 133

3.6  Set-up stage and modules to operate in a real 3G network ...................... 135
  3.6.1  Need for set-up in the plug-and-play adaptive antenna ....................... 136
  3.6.2  Call establishment in UMTS ............................................................... 136
  3.6.3  Proposed set-up for adaptive antenna in a UMTS network .................. 137
  3.6.4  Real-time DSP implementation ........................................................... 139
    3.6.4.1  Implemented modules ................................................................. 139
    3.6.4.2  Real-time operation .................................................................... 141
  3.6.5  Test results of set-up module ............................................................ 141

3.7  General conclusions, contributions and further research on adaptive antennas for mobile system .......................................................... 143
ADAPTIVE ANTENNA FOR MOBILE SYSTEMS: design, implementation and evaluation of signal processing modules

The first step for taking advantage of spatial diversity by using multiple antennas is the use of adaptive antennas, or more generally speaking, smart antennas based on beamforming techniques (MISO or SIMO systems). Despite their undeniable advantages (at least theoretically speaking), they have not been fully deployed in mobile systems, mainly due to their cost and complexity. However, their interest as the preface for studying MIMO systems is clear, and the analysis of a real prototype presents many novel aspects to be studied.

This chapter is devoted to the design, implementation and evaluation of an adaptive antenna for mobile systems, specifically for UMTS. Several novelties are presented, as a plug-and-play design to make the antenna independent of the network, some signal processing issues and the method to evaluate and measure the antenna performance. Moreover, the shortcomings of the implemented system (as its application mainly for rural or hotspots scenarios) are shown.
3.1 Introduction and motivation

Trying to improve mobile radio systems, and after having exhaustively used frequency, time and coding resources, smart antennas were presented as the solution to take advantage of a new resource: spatial diversity. Although used since some time ago mainly in radar and military applications, smart antennas applied to non-military wireless communications were proposed relatively recently.

The smart antenna concept is applied to several kinds of antenna arrays. We may remember here that with the term “smart antennas” we refer to systems with antenna array at one link end and that utilize beamforming techniques, as indicated in Chapter 2. Phased arrays, switched multibeam antennas and adaptive array antennas are usually included under this smart antenna concept with the only condition of including the possibility of controlling the radiation pattern at a certain level. Great theoretical advantages have been reported for the smart antenna implementation in base stations for mobile cellular communications, as we will see later, but this kind of antenna has not been extensively applied to those systems yet. The reasons are several: their complexity, the requirements to be enforced to the base stations, their not so good results in dense urban environments, etc. It seems clear that more studies based on real implementations are needed in order to overcome the current shortcomings that prevent smart antennas from being extensively used in mobile systems.

The main features of phased arrays, switched multibeam antennas and adaptive arrays are explained in chapter 2. When comparing the possibilities and performance offered by the three aforementioned types of smart antennas, we can state that adaptive arrays show significant advantages over the other 2 smart antenna types [188]: while phased arrays and multibeam antennas can only improve signal to noise rate (SNR) thanks to a higher gain in the user direction, smart antennas can also increase the signal to interference plus noise rate (SINR) since they can cancel (or at least significantly reduce) the interference level by reducing the antenna gain in the interfering direction. For CDMA systems, where the system is usually limited by interference level, an increase of sector capacity can be obtained for those cells with base stations equipped with smart antennas. The capacity increase is higher in cells with high interference levels, usually produced by users with high bit rates. An example of these scenarios is hot spots, where the use of smart antennas is of great interest, as shown in some theoretical studies (see next section). As the main shortcoming of smart antennas, it has been shown that, when used with no other processing system, they are only suitable for rural and other scenarios where multipath components do not carry significant signal power, as non-dense urban scenarios. However, their performance is reduced when they are used in scenarios with significant multipath components, as typical urban ones [189], [190]. This is due to the negative effect of multipath rays, which are considered as interference for the adaptive system if they are not properly processed. A way to reduce the negative effect of multipath components and even
to use them to enhance the system performance is using a multi-tap receiver, as the well-known rake receiver firstly proposed in [191], and widely used in broadband CDMA systems [192]. The rake receiver concept can be extended to the multi-antenna case [193], giving way to a broadband beamforming system. This adaptive antenna system can be used when spread delay is large enough to estimate the delays between the main path and other secondary paths that may arrive to the adaptive antenna, as in urban macrocell scenarios. Nevertheless, when multipath delays are quite small compared with chip rate and thus inter-symbol interference (ISI) is significant, rake receivers and adaptive antennas are not suitable and MIMO techniques are preferred [190], [194]. As a result, the work about adaptive antennas presented in this chapter is applicable to rural or similar environments (low multipath richness). On the contrary, next chapters will present the work focused on MIMO systems, with main application in indoor and microcell scenarios.

The main signal processing block in an adaptive antenna is the beamforming module. Many beamforming adaptive algorithms have been presented in the literature and are still an active research field. Conventional beamforming techniques are based on linear processing of input signals, where the weighting values are computed by using different types of references (spatial reference, time reference, etc…). Novel algorithms based on non-linear combination of input signals have also been presented recently, as neural network-based beamformers and MBER algorithms (see chapter 2, section 2.1.2.2 for a summary). Although very interesting for their promising improvement in performance, the latter ones involve a much higher computational complexity than the former ones, which prevents them from being used for real implementations, at least those real systems that are carried out on software platforms specially developed for linear processing as fixed-point DSPs. We have performed a study of multiple linear and non-linear beamforming algorithms for adaptive antennas, with the ultimate objective of selecting the most suitable method to be used in our adaptive antenna prototype, where specific features as the type of Software-Radio platform to be used, the required bandwidth and the available computational capacity for a reasonable hardware were taken into account.

A significant research effort has taken place in the last years to introduce smart antenna systems in cellular scenarios (see chapter 2, section 2.1.3, for a summary on state-of-the-art prototypes and commercial smart systems). However, the deployment of these antenna systems has not become a reality yet due to their cost and complexity. In practice, most of the commercially deployed smart antennas are switched-beam antennas for 2G systems. One of the reasons for this is the complexity of adaptive antennas in 3G. In contrast to 2G systems, where beamforming can be done in RF, beamforming in 3G must be applied after demodulating the CDMA signal, so that adaptive antenna functions need to be integrated into the digital baseband-processing sections of the base station. As a result, some existing smart antenna solutions for 3G have been developed for a unique base station equipment manufacturer, as [46]
and [47]. This fact makes the deployment of smart antenna systems unfeasible for mobile communications operators, due to the high associated cost and manufacturer dependency.

To overcome these problems, the implementation of adaptive antennas in 3G base stations should aim at a reconfigurable and flexible architecture. These features can be obtained using software radio platforms [195]-[197]. Moreover, a “plug and play” smart antenna solution, appropriate for any base station from any manufacturer, is of great interest, in order to reduce complexity in the integration of the smart antenna with the base station. We present here a smart antenna prototype based on these two premises: software radio platforms and plug and play solution.

The work presented here is part of the practical implementation of an adaptive plug and play smart antenna for 3G mobile communication systems based on W-CDMA like UMTS. The system has been called ADAM, which stands for “ADaptive Antenna for Multioperator scenarios”, as it can be connected to any base station site even shared by several operators. The work done by the author of the thesis and presented in this chapter is focused on the signal processing modules, since they present the most important novelties beyond the state of the art regarding the prototype implementation. However, a general description of the whole system is done in order to give a general picture of the system. The plug-and-play concept applied to an adaptive antenna for UMTS was previously presented in [198] and it gave rise to a patent [199]. The implementation, measurements and results of ADAM prototype were presented in several papers and the signal processing design and implementation gave rise to another patent [200] (see chapter 7 on contributions of this thesis).
3.2 Beamforming and 3G, theoretical study

3.2.1 Introduction

Adaptive antennas are basically based on a beamforming module that combines the signals received from the multiple radiant elements of an antenna array to optimize a certain parameter. Many algorithms have been presented in the literature, but beamforming is still an active research area. Some algorithms use a temporal reference, others work with a spatial reference and finally there are algorithms that take advantage of other features of the signal, as constant modulus or cyclostationary characteristics. In addition, conventional beamforming algorithms are designed to minimize the mean square error of beamformed signal compared with the reference signal. However, other novel methods have also been presented, such as algorithms that minimize the bit error rate.

The beamforming algorithm is one of the key modules in the adaptive antenna. Therefore, a preliminary study of state of the art beamforming methods was done. Some of the possibilities were studied, including conventional and novel adaptive algorithms for beamforming, with special emphasis put in its application to a real-time adaptive antenna prototype. Due to the wide and prolific area of research in adaptive algorithms, presenting an exhaustive study here would be too long and is not the objective of this thesis. Therefore, only the most interesting cases are presented below, with special emphasis in the two algorithms (NLMS and RLS) that were actually implemented in a DSP platform.

3.2.2 Analysis and theoretical study of beamforming methods

As previously mentioned, the research field of adaptive algorithms applied to beamforming present a wide variety of solutions. In chapter 2, section 2.1.2.2, an overview of beamforming methods, adaptive algorithms and their classification is given. Some of them have been selected as good representatives of beamforming methods, and their features have been analysed in order to be used in the adaptive antenna prototype. The analysis is focused on the algorithms NLMS (normalized least mean square) and RLS (recursive least square) as significant conventional algorithms. Also beamforming based on neural networks and the MBER optimization condition has been studied by means of simulation. NLMS and RLS are linear algorithms based on simple operations, and thus they are more suitable for implementation in fixed-point platform as the one used here. On the other hand, beamforming algorithms based on neural networks or MBER are of great interest theoretically, but are not very suitable for real implementation. Thus, only results from simulations for NLMS and RLS are presented here. The simulations are used to select the most interesting method to be implemented in the adaptive antenna prototype. The theoretical study and simulations that we performed for neural networks and MBER beamforming for cellular systems can be consulted in [201], [202].
3.2.2.1 Conventional adaptive algorithms: LMS (and NLMS) and RLS

Two well-known adaptive algorithms used in many applications (filtering, minimization problems, etc…) are the LMS and RLS algorithms [203]. Their features have been studied for many years, and multiple versions have been presented in the literature. LMS algorithm is a steepest-descent method, where the adaptive weights are iteratively updated by following the instantaneous gradient of the quadratic error surface in the negative direction. The LMS weights \( w_{LMS} \) are firstly initialized in then iteratively computed by following eq. (3-1):

\[
\begin{align*}
    w_{LMS}(n+1) &= w_{LMS}(n) + \frac{\mu}{\|x(n)\|^2} \cdot \nabla e(n) \\
    &= w_{LMS}(n) + \mu \bar{x}(n)e^*(n)
\end{align*}
\]

where (\( \cdot \)\(^*\)) indicates conjugate. \( \bar{x}(n) \) is the output signal vector of the antenna array (input signal for the adaptive beamforming) at instant \( n \), \( d(n) \) is the reference signal and \( e(n) \) is the error signal after beamforming. The parameter \( \mu \) is called the step size and it indicates the amount that the weights are moved in the opposite direction of the gradient. It has to be carefully selected: a too small \( \mu \) would reduce the speed of convergence of the algorithm, while a too large \( \mu \) would cause that the algorithm never converges.

As main drawbacks of this algorithm, it has a low convergence and has an undesired high dependence on the correlation of input signals. The great advantage of LMS is its very low complexity, with the order \( O(L) \) being \( L \) the number of elements in the antenna array.

In order to avoid dependence of convergence on input signals, a very interesting version is the NLMS algorithm: the step size is normalized by the power of input signals (norm of \( x(n) \) vector), so the weights are updated as in eq. (3-2):

\[
\begin{align*}
    w_{NLMS}(n+1) &= w_{NLMS}(n) + \frac{\mu}{\|x(n)\|^2} \cdot \bar{x}(n)e^*(n)
\end{align*}
\]

It can be demonstrated that for NLMS algorithm to converge it is needed that the step size \( \mu \) is in these bounds: \( 0 < \mu < 2 \).

The RLS algorithm is based on the use of an estimation of the autocorrelation matrix of the input signal \( \bar{x}(n) \), \( \mathbf{R}^{-1} \), which is estimated as temporal weighted averages. It is used as step size to update the weights. The recursive equation is:

\[
\begin{align*}
    w_{RLS}(n) &= w_{RLS}(n-1) - \frac{\mathbf{R}^{-1}(n)\bar{x}(n)e(w_{RLS}(n-1))}{\delta_0 + \bar{x}^H(n)\mathbf{R}^{-1}(n-1)\bar{x}(n)}
\end{align*}
\]

In order to reduce computational load, the Matrix Inversion theorem is used. This way, the inversed matrix \( \mathbf{R}^{-1} \) can be computed recursively as:

\[
\begin{align*}
    \mathbf{R}^{-1}(n) &= \frac{1}{\delta_0} \left[ \mathbf{R}^{-1}(n-1) - \frac{\mathbf{R}^{-1}(n-1)\bar{x}(n)\bar{x}^H(n)\mathbf{R}^{-1}(n-1)}{\delta_0 + \bar{x}^H(n)\mathbf{R}^{-1}(n-1)\bar{x}(n)} \right]
\end{align*}
\]

where an important parameter is the forgetting factor, \( \delta_0 \). It is directly related with the number of samples that are used to estimate the autocorrelation matrix, \( \mathbf{R} \). Therefore, it has a significant
Adaptive antenna for mobile systems

influence in the algorithm behaviour: for high values of $\delta_0$ (always smaller than 1) the algorithm will include a larger number of samples in the estimation of $R$ and thus it will converge faster. However, it will not be able to follow fast changes in the channel. Conversely, for smaller $\delta_0$ the algorithm converges slower but does not include a long “memory” of the past.

The matrix $R^{-1}$ is initialized as: $R^{-1}(0) = 1/\varepsilon_0$ where $\varepsilon_0$ is chosen as a small number (close to zero).

As the main characteristics of RLS, it is a faster algorithm and presents no dependence of convergence with statistics of input signal. However, as the main drawback of the algorithm, its complexity is much higher than LMS algorithm: order $O(L^2)$, where $L$ is the number of elements of the array.

Both NLMS and RLS algorithm were studied and their performance compared as algorithms to be used in the beamforming module of an adaptive antenna for UMTS. A complete UMTS system at link level (uplink) was simulated, including QPSK modulation, scrambling, spreading, multiple users in the same cell (same frequency, different user code), selection of either NLMS or RLS algorithm to beamform the signal, different options to estimate the channel following the standard recommendations for channel modelling from 3GPP [204] (urban, suburban, rural, ideal no multipath...), etc. Different parameters are obtained as output of the system, such as array factor, SINR and BER for each user. The implemented simulation tool is not thoroughly explained here for the sake of summarization, but Figure 3-1 depicts the its main blocks. The iterative expressions that were used to compute the optimal beamforming weights are summarized in Figure 3-2.

---

**Figure 3-1.** Schematic representation of simulated UMTS system to study NLSM and RLS algorithms.
The realized simulations help to cover two main purposes: first of all, we have studied the best values to be used as step size $\mu$ and forgetting factor $\delta_0$ for our application (beamforming with adaptive antenna in UMTS). Secondly, these simulations are valuable to know the performance of the system that can be expected and its comparison with a conventional sectored antenna.

A good selection of the step size $\mu$ in NLMS algorithm is of great importance, also considering that for the algorithm to converge the condition $0 < \mu < 2$ must be fulfilled. In a first step in the study, several simulations were carried out with different values of $\mu$ factor, to show its effect in the algorithm performance when it is applied in a UMTS scenario with a smart antenna. Table 3-I summarizes the values for parameters in the simulation environment.

<table>
<thead>
<tr>
<th>System parameters</th>
<th>TX parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Adaptive algorithm</td>
<td>Num DPDCCH 1</td>
</tr>
<tr>
<td>Num. UMTS frames</td>
<td>SF_{data} 64</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameters of channel</th>
</tr>
</thead>
<tbody>
<tr>
<td>SNR</td>
</tr>
<tr>
<td>Scenario</td>
</tr>
<tr>
<td>30 dB</td>
</tr>
<tr>
<td>Ideal (no multipath)</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameters of adaptive algorithm</th>
<th>Parameters of antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial weights</td>
<td></td>
</tr>
<tr>
<td>step size, $\mu$</td>
<td>Num. antennas 4</td>
</tr>
<tr>
<td>variable</td>
<td>Element spacing $\lambda$</td>
</tr>
</tbody>
</table>

Table 3-I. Parameters used in simulation to study $\mu$ in NLMS.
As an example, some intuitive results are shown for a situation where the desired user is located at -20º of the broadside direction of the array, and there are four interfering (extracell) users impinging with DoA = [-50º, 0º, 15º, 25º] in the azimuth plane. Two values have been used as step size: a small value (μ = 0.3) and a large value (μ = 1.5). Figure 3-3 depicts the radiation patterns that would be obtained with a 4-element adaptive array using NLMS in a UMTS environment, after processing 5 UMTS frames, with two different values of step size μ. For comparison reasons, also the radiation pattern for a typical sectored antenna is presented. We observe that the final result is almost identical for the two values of μ, although very small differences in levels for nulls can be noted. Despite this, similar results in final radiation pattern do not mean that the algorithm behaves equally for the two step size values, as we present afterwards. Moreover, some differences in performance are noted when SINR and BER are studied.

![Simulated radiation pattern of adaptive antenna with NLMS algorithm](image)

**Figure 3-3. Example of radiation pattern obtained from simulation in an UMTS scenario, with an adaptive antenna using NLMS and two values of step size μ.**

Figure 3-4 shows the evolution of one of the adaptive weights after the five UMTS frames, assuming that in each frame there are 150 pilot bits to be used to update the beamforming weights. It can be seen that after 5 UMTS frames the algorithm has achieved the final suboptimal value of weight, for both μ values. However, for μ = 0.3 the convergence is much slower, and the algorithm needs around 600 iteration (4 frames) to reach the final weight value, while for μ = 1.5 only approx. 150 iterations (one frame) are needed. On the other hand, for the larger μ value the residual error is much higher and the solution has larger fluctuations, which is not desired because it may cause flooding errors in the algorithm.
Chapter 3

Figure 3-4. Evolution of one of the weights, for two different values of $\mu$.

Figure 3-5 shows the evolution of the mean square error for the two $\mu$ values, while Figure 3-6 shows the SINR. Again, it is clear that a slightly better performance in terms of SINR is obtained with the lower $\mu$ value, at the cost of a slower convergence.

Figure 3-5. Evolution of mean square error with NLMS algorithm, for two values of step size $\mu$. 

$\mu = 0.3$  
$\mu = 1.5$
Several simulations with different $\mu$ values and situations were done, with up to 3 desired users and up to 10 interfering users. Their results are not included here for the sake of summarizing. As a result, a fair trade-off value of $\mu = 0.5$ can be chosen for the adaptive antenna to be implemented, since it allows for achieving the final weights values with approximately 3 UMTS frames and offers a fair residual mean quadratic error. Nevertheless, thanks to the reconfigurability of the Software Radio Platform on which the signal modules are implemented, this parameter can be changed according to the scenario. For example, with low mobility of users a smaller $\mu$ will achieve better results, while for a high-mobility one (rural scenario with a highway to be covered) a higher $\mu$ value will be better.

Similarly, the RLS algorithm was studied. As a representative example, the performance of the algorithm in the UMTS scenario is compared for two values of forgetting factor, $\delta_0$. By using the same parameters as the ones used for NLMS (see Table 3-1), two different values of forgetting factor were analysed: $\delta_0 = 0.9$ and $\delta_0 = 0.9999$. The obtained mean square error for the two values is presented in Figure 3-7, where only the results for the first 150 iterations (reference bits) are shown. It is clear that RLS convergence is much faster than NLMS, since only approx. 10 iterations are needed to achieve the final sub-optimal weights. The two forgetting factors give similar results in terms of array factor and mean quadratic error, although the higher value obtains a slightly lower residual error. However, it was observed that for $\delta_0 = 0.9999$ the adaptive system was not able to properly follow changes in scenarios with users moving even at low speed, while the algorithm works properly with $\delta_0 = 0.9$ for this scenario. Therefore, it may be of interest to select a forgetting factor of $\delta_0 = 0.9$, or even lower for high mobility scenarios.
Figure 3-7. Mean square error for RLS, with two different values of forgetting factor, $\delta_0$.

When comparing the radiation pattern obtained with both algorithms, they present quite a similar result for the situation of the example, although a detailed view shows that deeper nulls are given by the RLS algorithm, thus giving a slightly better result for the RLS algorithm. Figure 3-8 shows the radiation pattern for both NLMS ($\mu = 0.5$) and RLS ($\delta_0=0.999$) after 5 DPCCH frames.

Figure 3-8. Comparison of NLMS and RLS performance in a UMTS scenario. Radiation pattern.

Figure 3-9 shows the evolution of SINR for the same situation, for both NLMS and RLS. As expected, RLS converges much faster than NLMS in this scenario, and also achieves a slightly higher final SINR value.
In order to compare NLMS and RLS performance in a more general way, multiple situations in terms of number of users and interferences and different DoAs were simulated. 28 cases are analysed, as shown in Table 3-II. For each case, 100 MonteCarlo simulations were done, so a more realistic information can be achieved. The step size and forgetting factor were selected as 0.5 and 0.9 respectively. The partial interference cancellation approach was used (see section 3.3.2), so intracell users are considered as desired ones while extracell users are considered as common interferences. A static scenario is assumed for users. Several situations were simulated in terms of number of desired users and number of interfering users. Figure 3-10 presents the BER that was obtained for a data channel (without considering any channel coding or correction) with spreading factor SF = 64. For comparison reasons, also the BER obtained for a conventional sectored antenna was shown. The main conclusion is that RLS performs slightly better than NLMS in this scenario, although the difference is not significant. Nevertheless, NLMS may be more interesting for a real time implementation, due to its lower computational complexity. Moreover, RLS performs better in a static scenario, but it may be a worse option for fast changing scenarios, due to the high “memory” of previous information when the used δ₀ is very close to 1. When comparing with the sectored antenna, the adaptive antenna clearly outperforms it, especially for cases with low number of users. When the number of users in the system is very high (especially the number of intracell users), the BER obtained with the adaptive antenna gets closer to the BER obtained with the sectored antenna, thus offering similar results for the worst case.
### Table 3-II. Considered cases for study comparing NLMS and RLS.

<table>
<thead>
<tr>
<th>Situation</th>
<th>N. desired users</th>
<th>N. interf. users</th>
<th>N. total users</th>
<th>Situation</th>
<th>N. desired users</th>
<th>N. interf. users</th>
<th>N. total users</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>15</td>
<td>5</td>
<td>2</td>
<td>7</td>
</tr>
<tr>
<td>2</td>
<td>2</td>
<td>1</td>
<td>3</td>
<td>16</td>
<td>6</td>
<td>1</td>
<td>7</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>3</td>
<td>4</td>
<td>17</td>
<td>1</td>
<td>7</td>
<td>8</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>2</td>
<td>4</td>
<td>18</td>
<td>2</td>
<td>6</td>
<td>8</td>
</tr>
<tr>
<td>5</td>
<td>3</td>
<td>1</td>
<td>6</td>
<td>19</td>
<td>3</td>
<td>5</td>
<td>8</td>
</tr>
<tr>
<td>6</td>
<td>1</td>
<td>5</td>
<td>6</td>
<td>20</td>
<td>4</td>
<td>4</td>
<td>8</td>
</tr>
<tr>
<td>7</td>
<td>2</td>
<td>4</td>
<td>6</td>
<td>21</td>
<td>5</td>
<td>3</td>
<td>9</td>
</tr>
<tr>
<td>8</td>
<td>3</td>
<td>3</td>
<td>6</td>
<td>22</td>
<td>6</td>
<td>2</td>
<td>9</td>
</tr>
<tr>
<td>9</td>
<td>4</td>
<td>2</td>
<td>6</td>
<td>23</td>
<td>7</td>
<td>1</td>
<td>10</td>
</tr>
<tr>
<td>10</td>
<td>5</td>
<td>1</td>
<td>6</td>
<td>24</td>
<td>5</td>
<td>4</td>
<td>10</td>
</tr>
<tr>
<td>11</td>
<td>1</td>
<td>6</td>
<td>7</td>
<td>25</td>
<td>8</td>
<td>1</td>
<td>10</td>
</tr>
<tr>
<td>12</td>
<td>2</td>
<td>5</td>
<td>7</td>
<td>26</td>
<td>3</td>
<td>7</td>
<td>10</td>
</tr>
<tr>
<td>13</td>
<td>3</td>
<td>4</td>
<td>7</td>
<td>27</td>
<td>7</td>
<td>3</td>
<td>10</td>
</tr>
<tr>
<td>14</td>
<td>4</td>
<td>3</td>
<td>7</td>
<td>28</td>
<td>9</td>
<td>1</td>
<td>10</td>
</tr>
</tbody>
</table>

**Figure 3-10.** BER of data channel for UMTS with adaptive antenna, cases of Table 3-II.
Finally, it is of interest to study the complexity of each algorithm. We have calculated the number of operations that are needed to carry out one loop for each iteration for both NLMS and RLS. Table 3-III shows the results, in number of additions and products, first in a general way and afterwards for a number of antennas equal to 4 (actual number of antenna elements in the prototype). As expected, RLS requires much more operations and thus a higher computational capacity. For the specific case of using 4 antennas, RLS requires more than 20 times more operations than NLMS.

<table>
<thead>
<tr>
<th></th>
<th>NLMS</th>
<th>RLS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Num real additions</td>
<td>4·L</td>
<td>20·L²</td>
</tr>
<tr>
<td>L = 4</td>
<td>16</td>
<td>320</td>
</tr>
<tr>
<td>Num real products</td>
<td>4·L+2</td>
<td>L·(21·L+8)</td>
</tr>
<tr>
<td>L = 4</td>
<td>18</td>
<td>368</td>
</tr>
</tbody>
</table>

Table 3-III. Required number of operations for one loop (iteration) of NLMS and RLS

### 3.2.3 Beamforming for 3G systems: advantages and drawbacks

As mentioned in previous chapters, one of the systems where smart antennas should be of applicability are 3G systems, since they are based on the use of different orthogonal codes to distinguish and demodulate-modulate different users working in the same frequency and time, that is, CDMA (Code Division Multiple Access). Since these CDMA-based systems are usually limited by the level of interferences, it seems clear that adaptive antennas will provide a non negligible improvement.

Should we consider beamforming for UMTS, and adaptive algorithm based on temporal reference seems the natural one to be used. This is due to the structure of dedicated channels in this system: DPCCH (dedicated user channel that is always active when the user has initiated a communication) contains pilot bits that are specified by the standard and thus they are known and can be used as a temporal reference.

One of the main drawbacks in UMTS is that the 3G implemented version is UTRA-FDD, that is, uplink and downlink do not use the same frequencies, and this means that the channel in one direction cannot directly be assumed as equal to the channel in the other direction. As a result, some kind of compensation may be needed to compute optimal beamforming weights for downlink from uplink.
3.3 An adaptive antenna for 3G systems: general description and operation

3.3.1 Introduction

From previous results, it seems clear that the mobile communication systems that suffer from high interference levels will greatly benefit from adaptive antennas. After the theoretical study, next step seems clear: a adaptive antenna prototype have been proposed, designed and implemented. The adaptive antenna proposed here is based on the plug-and-play concept first proposed in [198], [199], which is explained below. A general overview of the overall adaptive antenna is given, to focus afterwards on the signal processing part.

3.3.2 System operation: the “plug and play” concept

As previously mentioned, despite their clear advantages smart antennas have not been widely adopted in 3G communication systems. One of the reasons for this is the high complexity that the smart antenna involves. Contrary to conventional sectored antennas, which are solely based on radiofrequency signals and do not require any specific base station to support their operation, smart antennas based on adaptive algorithms and full reconfiguration of radiation pattern require that the optimal combination of signals from different antenna elements is performed at baseband level.

In the conventional concept of adaptive antenna for UMTS, the beamforming stage is done after demodulation, at a bit level, so the base station must cooperate in the beamforming process. Figure 3-11 represents a general structure of beamforming for uplink for a W-CDMA-based system as UMTS, as conventionally understood. We observe that there are several output signals after beamforming, specifically there is a stream of beamformed data per user. Then, the Node-B should be able to cope with this type of information.

In order to simplify the signal processing structure and to make the adaptive antenna independent of the Node B to be used behind it, a novel architecture is proposed. It aims at allowing a plug and play operation, that is, a beamforming and combination of signals that is performed independently of the base station. As a result, the adaptive antenna will be suitable to be used with conventional Node B’s, not requiring specific “smart” base stations, thus reducing deployment costs on already deployed networks and avoiding manufacturer dependence for mobile operators.

The key point in the new architecture is that some signal processing will be done by the adaptive antenna, but just the one required to compute the beamforming weights and to optimally combine the user signals. Therefore, not all the physical channels have to be processed (modulated or demodulated). Moreover, some of the operations can be performed at chip rate, avoiding the computational cost of signal despreading.
Figure 3-12 shows the proposed plug-and-play architecture, for uplink. As a plug and play functionality is demanded, the beamformed signals are upconverted and sent through a direct connection between the smart antenna output and the base station input by means of the standardized Uu interface [203], similarly to the way a sectored antenna would send signal to the Node B. In order to reduce the computational requirements of the system, in the uplink only those interferences common to the intracellular users and all the extracellular interferences are cancelled: we have called it a partial interference cancellation approach, as opposed to the total interference cancellation approach. The relationship between the extracellular and intracellular interferences is called the extracellular interference factor, $F$, and has a value between 0.4 to 1.4 depending on the environment and the service [206]. This implies that more than 50% of the interferences are cancelled on average, as the common intracellular interferences (interference due to other systems, as impulsive noise, etc) should also be taken into account.

In the proposed plug-and-play approach for uplink (Figure 3-12) for W-CDMA, beamforming is performed at chip level and with the same beamforming weights for all the users. As previously stated, this is equivalent to cancelling only the common interferences (partial interference cancellation). This way the computational complexity is highly reduced, since demodulating and modulating all the dedicated data channels of users is not needed. The cost to be paid is some reduction in performance of the adaptive antenna, its value depending on the $F$ factor in the specific environment we are working in. We note also that with this approach all the uplink physical channels that are transmitted by each user are beamformed, even the control uplink channels (PRACH, PCPCH). This implies that if a user is received at a certain azimuth angle with respect to the antenna array, and there are strong interferences at this angle of arrival, the adaptive antenna will not be able to “see” the new user and no dedicated channels will be allocated for him. However, this situation makes sense: if a user is been received in the same angle as strong interferences, his quality of service will be very low and it is better that he gets served by other base station.
Chapter 3

Figure 3-12. General architecture of adaptive antenna (uplink), with the plug-and-play concept.

Figure 3-13 shows the general architecture for the proposed plug-and-play concept in downlink. The optimal cancellation of all interferences (intra and extracell users) is kept, so all the user channels have to be demodulated to be independently beamformed for each user. Afterwards, they are modulated again, and combined to be sent to the corresponding antenna. Therefore, a total interference cancellation is applied for downlink.

Figure 3-13. General architecture of adaptive antenna (downlink), with the plug-and-play concept.

Contrary to what is done for uplink, in downlink only the dedicated user channels are beamformed. In order to do so, broadcast channels are sent only through one of the antennas, which is equivalent to transmit them with a conventional sectored antenna. This way the broadcast and control operations, which are carried out by higher levels in Node B, are not affected by the adaptive antenna. Moreover, the channels used for synchronization in the UMTS system, such as CPICH, P-CCPCH and S-CCPCH, are transmitted with a non-directive antenna, allowing it to be received by all users, as needed for a correct operation of the system.
Adaptive antenna for mobile systems

According to the physical layer of UMTS [203], time reference and user synchronization may be obtained in the uplink from DCH (in particular DPCCH). However, downlink allows several ways to obtain time reference and user synchronization: CPICH, P-CCPCH, S-CCPCH, and even pilot symbols or diversity pilots [206]. Our implementation gets user synchronization from DPCCH in the uplink, and from CPICH in the downlink. Table 3-IV and Table 3-V summarize which physical channels are processed in up and down streams to get system information and which channels are beamformed or not by this adaptive antenna prototype. It is important to note that this prototype was designed and implemented for Release 99 of 3GPP standards, FDD-UTRA, and that high speed channels of newer versions (HSDPA, HSUPA) are not considered.

<table>
<thead>
<tr>
<th>Channel</th>
<th>Function in smart antenna</th>
<th>Beamforming</th>
</tr>
</thead>
<tbody>
<tr>
<td>DPCCH</td>
<td>User synchronization and uplink channel characterization</td>
<td>Yes</td>
</tr>
<tr>
<td>DPDCH</td>
<td>-----</td>
<td>Yes</td>
</tr>
<tr>
<td>PRACH</td>
<td>-----</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 3-IV. Beamforming of uplink physical channels.

<table>
<thead>
<tr>
<th>Channel</th>
<th>Function in smart antenna</th>
<th>Beamforming</th>
</tr>
</thead>
<tbody>
<tr>
<td>SCH</td>
<td>Cell slot synchronization</td>
<td>No</td>
</tr>
<tr>
<td>CPICH</td>
<td>Downlink frame synchronization</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>User synchronization (scrambling code identification)</td>
<td>No</td>
</tr>
<tr>
<td>P-CCPCH</td>
<td>-----</td>
<td>No</td>
</tr>
<tr>
<td>S-CCPCH</td>
<td>-----</td>
<td>No</td>
</tr>
<tr>
<td>AICH</td>
<td>-----</td>
<td>No</td>
</tr>
<tr>
<td>CSICH</td>
<td>-----</td>
<td>No</td>
</tr>
<tr>
<td>PICH</td>
<td>-----</td>
<td>No</td>
</tr>
<tr>
<td>DPCH</td>
<td>-----</td>
<td>Yes</td>
</tr>
<tr>
<td>PDSCH (DPCH)</td>
<td>-----</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 3-V. Beamforming of downlink physical channels.

3.3.3 General description and main features

With the purpose of giving a general overview of the overall prototype, we present here the general operation of the system and its modular architecture. This chapter, however, is devoted to the signal processing part (mainly beamforming, modulation and implementation effects on beamforming performance), which was the part where the work of the author was focused. Therefore, the radiofrequency and antenna modules are just briefly commented. Interested authors can refer to [207], [208] for more details on these aspects.

Figure 3-14 shows a “high level” architecture of the developed adaptive antenna, where the general modules are depicted. The system consists of an antenna array module and a radiofrequency (RF) to intermediate frequency (IF) module to connect the air interface with the digital processing subsystem. This subsystem is where the “intelligence” of the smart antenna lies, and it is where beamforming of signals is performed. Finally, the smart antenna has an RF-
IF module that allows communication between the smart antenna system and the base station (Node B in UMTS nomenclature), so a standardized interface (radiofrequency interface, or Uu interface) is used. This way a transparent operation from the point of view of the Node B is possible.

Figure 3-14. General diagram of the adaptive antenna.

In the downlink, the RF signal from Node B is down-converted to IF, digitized and then sent to the signal processing modules, where several DSPs process the signal. It is demodulated, beamformed (with a set of different weights for each user) and modulated again. The resulting signal is added up so the signals of beamformed users are put together to be sent to the antennas. Finally, the beamformed signals are analog converted, up-converted to RF and transmitted through the antenna array.

In the uplink, an equivalent process is performed but using a common beamforming vector for all the users. The analog signal received by the antenna elements is downconverted by the RF-to-IF chains and digitalized. Afterwards, it is processed in the digital signal processing module, where several DSPs work in parallel to beamform the signal of all intracell users (with a single vector of weights for all of them). Once the baseband signal has been computed, it must be analog converted and then up-converted again in the IF-to-RF chains to the original RF carrier in order to interface adequately with any standard Node B, as it can be seen in Figure 3-14.

As explained above, this architecture performs a total interference cancellation in the downlink but only a partial interference cancellation in the uplink.

The three main modules that formed the smart antenna prototype are briefly explained below. A detailed explanation of the signal processing modules is given in next sections. A detailed explanation of the antenna modules and the RF modules is out of the scope of this thesis.
Antenna module

The adaptive array prototype uses four commercial sectored antennas for the UMTS band [209], each with a non-directive radiation pattern in azimuth and a directive radiation pattern in elevation. They were characterized by means of measurements in anechoic chamber. The measured -3dB beamwidth for H-plane is 60°, and the measured -3dB beamwidth for E-plane is 7°. Each antenna has two ±45° polarization ports, and its gain is 17.5 dBi. The individual antennas are put together in a uniform linear array structure, as shown in Figure 3-15. With this configuration, inter-element separation is 15 cm (width of each sectored antenna), which is equivalent to 0.975λ and 1.070λ at the uplink (1950 MHz) and downlink (2140 MHz) central frequencies of UMTS FDD-UTRA, respectively.

Figure 3-15. Antenna module of the adaptive array prototype.

Radiofrequency module

The radiofrequency modules are responsible of up-converting the signal from IF to RF and down-converting from RF to IF, from both the antenna elements and the base station. A superheterodyne design was selected for the radiofrequency modules, with two frequency conversions: from RF to a high IF (f_{IF,high}) and from f_{IF,high} to the final IF where the digital conversion will be performed (f_{IF,low}). Several filters and amplifiers are used in the design. A simplified diagram of RF-to-IF module and IF-to-RF module are depicted in Figure 3-16 and Figure 3-17, respectively, where some of the less important components have not been represented for the sake of simplicity.
According to the software-radio concept, it is of great interest to realize digital-to-analog and analog-to-digital conversion as close to the antenna as possible. Thus, a very interesting option is to digitalize the signal at IF, instead of baseband. Moreover, this scheme reduces the phase noise, thanks to the use of digital oscillators. However, with this approach there is a trade-off that must be taken into account when selecting the intermediate frequency. On one side, a high IF simplifies the analog scheme; specifically the constraints for the filter after DACs are less tight. On the other side, a low IF makes easier the A/D and D/A conversion, and allows to find ADCs and DACs with higher vertical resolution or number of bits. After studying the available filters and DACs-ADCs in the market, an IF of $f_{IF,low} = 44$ MHz was selected as lower IF in the system. Regarding the upper IF, $f_{IF,high}$ it was chosen by considering the availability of filters, amplifiers and mixers in the market. Two different intermediate frequencies were selected for transmitter and receiver: 190MHz for the first one and 380MHz for the second one.

The implemented modules are customized boards for this prototype. Four types of boards are implemented: uplink receiver, uplink transmitter, downlink receiver and downlink transmitter. Figure 3-18 shows the position of each of the RF-IF modules, which communicates the digital processing modules with the radiant elements or the base station. The downlink transmitter and uplink receiver consist of 4 channels each, one for each antenna. Since having the same phase reference for the four antenna elements of the array is a must in a smart antenna system, the same oscillator must be used for the four channels, which is achieved by using signal dividers.
Figure 3-18. Radiofrequency modules in the overall structure of the prototype.

The RF module has to comply with the specifications for RF channels of UMTS FDD-UTRA standard. Specially demanding are the low interferences to adjacent channels specified by the standard. Different filters are selected to address this issue. Moreover, power amplifiers are used in the downlink transmitter stage, to guarantee the required transmitted power level. Low noise amplifiers are used in the receiver modules, so low noise level is introduced.

As an example of implemented boards, Figure 3-19 shows a photograph of some of the RF and connection boards implemented for the smart antenna prototype.

Figure 3-19. RF-IF modules.
Signal processing and digital module

The main operation of the adaptive antenna is carried out in the signal processing module. It consists of a platform with DACs, ADCs and processors that work in parallel to optimally compute the beamforming weights and to combine the signals. Different types of platforms for signal processing can be selected, from FPGAs to DSPs or microprocessors. The selected platform for our prototype is thoroughly explained in section 3.4.2.

According to the Software Radio Concept, analog to digital converters (ADCs) and digital to analog converters (DACs) are located just before the analog RF-IF chains, hence working with IF signals instead of the typical baseband signal. This allows most of the system modules to be implemented in software, which is a great advantage with respect to pure hardware implementations because the system can be easily reconfigured and updated with more advanced versions. As the smart antenna should be transparent for the base station, it should not implement the base station physical procedures, such as power control and handover, which are performed by the base station (Node B) itself. Moreover, polarization diversity is performed by the base station and the ADAM antenna is connected to both base station ports and processes each polarization independently.

The performance improvement that may be achieved with an adaptive antenna depends on the following aspects: antenna array geometry, adaptive algorithm that controls the beamforming process and propagation and interference environment. These issues were previously studied by simulation. They are not the objective of this thesis, but some of the main results will be presented in next section to show the great possibilities offered by the proposed adaptive antenna.

The main characteristics of the overall adaptive antenna prototype are summarized in Table 3-VI. More details on the signal processing module are given in next sections.

<table>
<thead>
<tr>
<th>Main characteristics of adaptive antenna prototype</th>
<th>General features</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating frequency (UMTS freq. band)</td>
<td>1920-1980 (uplink)</td>
</tr>
<tr>
<td></td>
<td>2110-2170 (downlink)</td>
</tr>
<tr>
<td>Carrier separation</td>
<td>4.6 to 5.6 MHz</td>
</tr>
<tr>
<td>Number of antenna elements (and RF chains)</td>
<td>4</td>
</tr>
<tr>
<td>Element spacing</td>
<td>0.15 m (0.975λ, uplink, 1.097λ, downlink)</td>
</tr>
<tr>
<td>3-dB beamwidth, H-plane, for each antenna element</td>
<td>60°</td>
</tr>
<tr>
<td>3-dB beamwidth, E-plane, for each antenna element</td>
<td>7°</td>
</tr>
<tr>
<td>Downtilt E-plane (elevation)</td>
<td>2° ± 0.5°</td>
</tr>
<tr>
<td>Element polarization</td>
<td>± 45° linear</td>
</tr>
<tr>
<td>Element gain</td>
<td>17.5 dBi</td>
</tr>
<tr>
<td>RF-IF modules</td>
<td></td>
</tr>
<tr>
<td>-----------------------------------</td>
<td>-------------------</td>
</tr>
<tr>
<td>Minimum freq. step for carrier selection</td>
<td>200 kHz</td>
</tr>
<tr>
<td>Sensibility</td>
<td>-112 dBm</td>
</tr>
<tr>
<td>Carrier bandwidth</td>
<td>4.4-5 MHz</td>
</tr>
<tr>
<td>Maximum output power for each carrier</td>
<td>1 W</td>
</tr>
<tr>
<td>Upper intermediate frequency, receiver</td>
<td>380 MHz</td>
</tr>
<tr>
<td>Upper intermediate frequency, transmitter</td>
<td>190 MHz</td>
</tr>
<tr>
<td>Lower intermediate frequency</td>
<td>44 MHz</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Signal processing modules</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Type of signal processing platform</td>
<td>DSPs+FPGAs</td>
</tr>
<tr>
<td>Used arithmetic at DSP</td>
<td>Fixed point</td>
</tr>
<tr>
<td>Number of bits for A/D, D/A</td>
<td>12 bits</td>
</tr>
</tbody>
</table>

Table 3-VI. Main features of smart antenna ADAM.

3.3.4 Simulation results and expected performance

Previous to the implementation of the adaptive antenna, a viability study was done to study feasibility and expected performance of the system. A detailed study on performance for smart antennas in W-DCMA systems can be found in [210]. Some simulation results based on the specific features of the adaptive antenna prototype presented here were shown in [211], [212]. Although they are not the objective of this thesis, the most interesting results are included here, in order to motivate the interest of implementation of this prototype.

As previously said, the operation of ADAM must be completely transparent to Node B. This approach has an impact on the performance achieved with the adaptive beamformer. To analyse this impact, a single-cell scenario with a variable number of mobile users is studied, including the effect of external interference on system performance and comparing the total and partial interference cancellation methods.

Figure 3-20 shows the array factor achieved for uplink with both cancellation schemes when an external interference source and three intracell users are present. NLMS is used as adaptive algorithm. The azimuth angle \( \phi \) is represented so the broadside direction is referred as 90°. The signal of desired user impinges on the array at \( \phi = 70° \), while two other intracell users impinge from \( \phi = 105° \) and \( \phi = 115° \). An external interference is received at \( \phi = 40° \). The radiation pattern for the sectored array is also plotted for comparison reasons. This radiation pattern is estimated as \( \cos(\phi)q \) where the \( q \) parameter is computed to give a radiation pattern with a 3dB beamwidth of 60° (to simulate measured radiation patterns for the used antenna elements).

As observed in Figure 3-20, both array factors cancel the external interference contribution. However, if the total cancellation scheme is used, contributions from other mobile users will also be cancelled, whereas with partial cancellation a simultaneous pointing in the directions of mobile users appears as a result of the linear combination of \( \mathbf{w} \). As the number of
users uniformly distributed within the cell is increased, the final uplink radiation pattern tends to provide a sectored coverage.

![Normalized AF, total cancellation](image1.png)

![Normalized AF, partial cancellation](image2.png)

**Figure 3-20.** Normalized array factor obtained with total interference cancellation and partial interference cancellation methods (●: mobile users, x: external interference).

A simulation environment with a uniform distribution of mobile pedestrian users has been studied. Mobile speed is 3 km/h, and multipath fading is given by the two-path profile proposed by 3GPP specifications. The angle of arrival of impinging rays for a user is characterized as a Laplacian azimuth spectrum along with a Gaussian distribution for each user, with an angular spread of 10°. The number of rays impinging on the array per user is found as a Poisson random variable with a mean value of 25 [213]. Each user transmits only one data channel, with a spreading factor of 64. In the simulations, a perfect power control algorithm is assumed for mobile users, and external interference power $P_{int}$ is set to $F$ dB over the transmitted power per user.

Figure 3-21 shows the average uplink SINR increase of the two methods with respect to a typical sectored antenna in two scenarios: $F = -150$ dB (only intracell mobile users are present in the cell), and one external interference with $F = 20$ dB. The results are obtained from link simulation where the prototype characteristics are used as parameters for the simulated smart antenna. In the first scenario, the SINR improvement converges to 6 dB (the gain due to pointing the main beam to the user) when the total cancellation scheme is used. With partial cancellation, this adaptive antenna will provide very similar performance to the individual sectored antenna. However, in the second scenario, the partial cancellation scheme outperforms the sectored antenna in more than 5 dB.
Adaptive antenna for mobile systems

In downlink, only the total interference cancellation scheme has been considered. Figure 3-22 shows the average SINR increase experimented by the mobile user when the proposed downlink beamforming algorithm is used. In the downlink, the adaptive antenna always obtains a better performance than the sectored antenna, thus giving an increase in SINR of 3 to 9 dB.

As a conclusion, we can say that this adaptive antenna will take profit of hot spots, improving the capacity in the vicinity of high occupied cells. In these situations, mainly higher power external interferences from multimedia services are cancelled by ADAM prototype. In these situations, the antenna would help the cells in the vicinity of a hot spot to expand their coverage and to compensate the “cell breathing” of high occupied cells.

System-level simulations were also carried out to estimate the expected performance of an adaptive antenna as the one implemented here, in a cellular system [210], [211]. The total throughput per cell in the downlink was computed. The results are not included here, for the sake of summarizing. The main conclusion was that by using the ADAM prototype, the throughput is increased by a factor of 2 in each sector, in relation to the situation with sector antennas. This capacity increase comes from the lower number of users put to outage when the adaptive antenna is used.
3.4 Signal processing modules: design and implementation for real-time operation

3.4.1 General architecture of signal processing modules

3.4.1.1 SDR concept

As previously mentioned, the software-radio concept was considered when designing the smart antenna of this work. This means that the digital modules are placed as close to the antenna as possible, so that the maximum reconfiguration is achieved. Three main advantages are obtained when using an SDR architecture:

- Improvements and new features are easier to introduce in the system.
- It is easier to update the system with new versions, once it has been implemented and after a first release.
- An SDR architecture allows to “download” different software-based implementations of the system to be used on the same hardware, thus offering the option to work with different communication standards if needed.

Due to its features of reconfigurability and flexibility, SDR architecture is optimum for developing a prototype as the one presented here.

Several configurations can be considered for the receiver in the SDR concept [196], from the ideal one with direct analog to digital conversion after the antenna (see Figure 3-23-a)) to the less ambitious case where the analog to digital conversion is carried out after analog conversion from intermediate frequency (see Figure 3-23-b)), where the second one is the average case between the ideal SDR architecture and the conventional heterodyne architecture (depicted in Figure 3-23-c)). Obviously, the same concept applies to the transmitter.
The high radio frequency where UMTS and other mobile systems operate (close to 2 GHz) makes unfeasible the ideal SDR architecture, since there are no analog-to-digital converters with high enough sampling frequency and enough vertical resolution (number of bits) for this conversion. Therefore, the solution with analog conversion from RF to IF, ADC operating at IF and digital downconversion from IF to baseband was chosen. The conversely diagram was chosen for the transmitter modules. Next section specifies the modular architecture for the SDR modules, where all the required signal processing is carried out. Following sections detail the design and implementation of each signal processing module.

3.4.1.2 Modular diagram

After analyzing the general SDR architecture to be used, we present here the modular diagram that was designed for signal processing in the adaptive antenna, for uplink (Figure 3-24) and downlink (Figure 3-25). The RF-IF and antenna modules are included just to simplify the general understanding of the system, and are only schematically shown.

As previously explained, for uplink a partial interference cancellation is done: only the interferences that are common to intracell users are cancelled, in order to reduce complexity. All the uplink channels are beamformed. After analog to digital conversion, the signal is downconverted to baseband. A synchronization module is used to synchronize the local user code with the received signal, for each user. After that, the dedicated control channel for each user is demodulated, and the control bits are used as reference bits to be fed to the weight computation module. The weight vector is dynamically computed by using an adaptive algorithm. Finally, the received signals are combined according to the optimal beamforming weights, and the resulting beamformed signal is upconverted and analog converted.

As opposed to uplink, in downlink only dedicated channels are beamformed. This is necessary, so that the channels that have to be broadcasted to the cell are not beamformed, otherwise the adaptive antenna would prevent some users from receiving the expected broadcast channels. The operation is similar to the uplink signal processing modules, but for downlink a total interference cancellation approach is performed. After synchronization, the dedicated channels for all users are demodulated (both data and control channels), then beamformed and finally modulated again. The resulting signals are combined for all users, and finally upconverted, analog converted, and sent to the corresponding antenna element. The broadcast
channels are sent through only the first antenna element, being equivalent to being transmitted by a conventional sectored antenna.

Figure 3-24. General diagram of signal processing modules, uplink.

Figure 3-25. General diagram of signal processing modules, downlink.

The design and implementation of each of the signal processing modules are explained in next sections.

3.4.2 Hardware platform

3.4.2.1 Selection of SDR platform

Several types of hardware for signal processing can be found in the market. The most utilized ones are usually classified into three groups: ASICs (application-specific integrated circuit), FPGAs (Field Programmable Gate Array) and DSPs (Digital Signal Processor). While
the first ones are designed and customized for a specific application and once implemented the design cannot be changed, the other two are reconfigurable through programming. Moreover, ASICs are economically feasible for large production volumes, while FPGAs and DSPs are more interesting for prototyping. Finally, if the software-radio concept is to be applied, ASICs have not sense since they do not allow the required reconfigurability to be available. As a natural conclusion, we narrowed the selection of SDR platforms to FPGAs and DSPs.

FPGAs contain programmable logic blocks and programmable interconnects that allow the same FPGA to be used in many different applications. In general, they are more efficient than DSPs in terms of number of simple instructions per second that they can perform. As a drawback, their programming is more complex than programming DSPs, since timing and synchronization among signals in the FPGA gates are quite complex to control, and it is required to know the specific programming language that is used to program them. On the contrary, DSPs are the more versatile and reconfigurable SDR platform, and they can usually be programmed with a conventional programming language, but they perform a lower number of instructions per second. As a result, in order to optimally select the hardware to be used, FPGA-based HW is preferred for the simple but intensive operations (filtering, downconversion, upconversion), while DSPs are selected as the core “intelligence” of the adaptive antenna (computation of optimal weights, modulation-demodulation, etc).

In next paragraphs we briefly explain the main parameters that were considered when selecting the ADCs and DACs, the FPGAs and the DSPs in the prototype.

**ADCs and DACS**

Several manufacturers offer multiple models of ADCs and DACs for communication applications. The main parameters to be taken into account when selecting ADCs and DACs for our application are:

- **Vertical resolution**: it is the number of bits that digitally represent the analog signal, after sampling and quantifying the signal. The lower the vertical resolution is, the higher the quantification noise is. For UMTS applications, the recommended number of bits to use is at least 12 [188].

- **Maximum supported sampling frequency**: for the ADC, it is the maximum sampling rate that can be used, while for the DAC it represents the maximum data rate that the DAC will be able to receive as input. It limits the maximum operating bandwidth of the system. For our application, since the minimum baseband bandwidth to be processed is 5 MHz (bandwidth of UMTS user channels after scrambling and spreading), a minimum sampling frequency of $f_{s,\text{min}}=10$ MHz is needed, which would force an intermediate frequency of $f_{\text{IF}}=5$ MHz (or a multiple if subsampling is allowed by the ADC operating bandwidth). For a more flexible design, a higher sampling frequency is needed.
· **Input (for ADC) or output (for DAC) signal range**: it indicates the limits for the level (or power) of signal to be received (ADC) or transmitted (DAC).

· **Maximum operating bandwidth**: it shows the limit in bandwidth of the analog signal that can be received (ADC) or transmitted (DAC).

· **Available reconfigurability**: it is of great interest that the ADCs and DACs parameters (such as sampling frequency, level of signal…) can be highly reconfigured.

Some of the aforementioned parameters are closely related. One clear relationship is sampling frequency and vertical resolution: the higher the vertical resolution we need, the lower the available sampling frequency in commercial ADCs. Figure 3-26 shows the situation of ADC technology, where the values were taken from commercial ADCs of main manufacturers [214], [215]. It is clear that vertical resolution is logarithmic inverted proportional to sampling rate. A similar curve is followed by DACs. We may note that these values were obtained at the moment of designing the adaptive antenna prototype (year 2003). Due to fast evolution of this technology, improved values may be currently found in the market.

![Figure 3-26. Technology situation for ADCs in terms of vertical resolution and sampling rate (taken at the moment of ADAM design)](image)

ADCs and DACs can be found as individual devices or as part of more complex boards that include other functionalities, as filtering, amplification and so on. When built together with FPGAs, they are usually referred to as front-ends. For the system implemented and presented as part of this thesis, front-ends for up and down conversion were selected, due to their good features and high flexibility. They are presented below.

**Front-ends and FPGAs**

In the selected SDR architecture, the digital signal must be downconverted to baseband by means of an IQ demodulator after the ADC. Similarly, before converting the digital signal to analog, an upconverter from baseband to IF is needed. One possibility could be to implement it
directly in general DSPs. However, this task is quite intensive and requires a very high computational capacity in our case, due to the high UMTS sampling rate. This would make the implementation unfeasible for conventional general-purpose DSPs.

Another interesting solution is to use FPGA technology to accomplish the most demanding tasks in terms of computational requirements. IQ demodulators or broadband downconverters, as well as IQ modulators or broadband upconverters are easily found in the market, based on FPGA platforms, together with ADC or DACs if needed; they are usually called front-ends. These modules can process the signal independently of the general DSPs, which can be used then to do the subsequent processing. Moreover, the free area in the FPGA is sometimes available for the user, so he can take advantage to include customized signal processing modules. This option was considered as the most interesting one for the adaptive antenna, so front-ends were acquired for the prototype.

When choosing the FPGA to be included in a design, the major vendors that offer general-purpose FPGAs are Xilinx and Altera. Both of them offer a wide range of FPGAs with different features. The number of logic cells in the FPGA (which limits the number of operations to be carried out), number of input/output pins, internal memory and clock rate are some of the main parameters to be considered when selecting the FPGA to be used. Some FPGA solutions include also an embedded microprocessor for control or high level computation.

Selected upconverter and downconverter modules

Taking into account the aspects mentioned above, general purpose front-ends were selected for both upconversion/DAC and downconversion/DAC. Apart from the general features, it is important to note that in the selection we considered two extra characteristics of special relevance for the adaptive antenna prototype: a high flexibility by reprogramming and the option of synchronizing multiple channels. The main features of the selected modules are presented below, and summarized in Table 3-VII.

The selected receiver boards are the model 6235 receiver front-ends, from Pentek [216]. Each receiver board consists of two ADCs from Analog Devices, and a Virtex II FPGA, where two IQ demodulators are implemented. Therefore, two identical receiver channels per receiver board are available. The vertical resolution for the ADCs is 12 bits, and its maximum sampling frequency is 105 Msps. The maximum allowed bandwidth for input signal is 80 MHz, and the maximum input power is 5 dBm. As a very interesting feature of the system, it presents a great number of parameters to be selected: sampling frequency, decimation of sampled signal, IF to be digitally generated, by-passing of IQ downconverter if preferred, etc. That allows a versatile implementation of the IQ downconversion module. Also of interest is the possibility of synchronizing several of these modules to synchronously receive input signals from 3 or more
channels, by using a shared clock. Moreover, the FPGA can be used for implementing extra modules if needed. Figure 3-27 shows a top and a perspective photograph of the 6235 board.

Figure 3-27. The 6235 2-channel digital receiver plus A/Ds

The selected transmitter boards are the model 6229 transmitter front-ends, also from Pentek [217]. Each 6229 board consist of two CMOS digital quadrature upconverters with DACs, from Analog Devices, plus two lowpass filters and two amplifiers. Therefore, each 6229 provides two identical and independent broadband channels. Each DAC accepts 12-bit digital signal as input, and its maximum sampling frequency is 200 MHz. The maximum output frequency for the generated analog signal is 80 MHz. Again, an interesting feature here is the reconfigurability of several parameters, as the output IF, gain in amplifiers, oversampling factor, type of synchronization source, etc. Also here there is the possibility of chaining together several 6229 boards to get multiple synchronized output signals. Figure 3-28 shows a photograph of one 6229 board.

Figure 3-28. The 6229 2-channel digital upconverter plus DACs
Selection of DSPs and data buses

All the signal processing stages that are not performed in the front-ends, such as synchronization and beamforming modules, are implemented in general-purpose digital processors (DSPs). As stated before, DSPs were selected due to their flexibility and multipurpose possibilities, as well as being easy to program. Some characteristics have been considered to select the DSPs, which are summarized below:

· **Arithmetic type**: either fixed-point or floating point can be selected. While floating point processors are optimized to accomplish these type of operations, they are able to perform a lower number of operations per second. Conversely, fixed-point DSPs are capable of realizing a higher number of simple instructions but at the cost of using less accuracy and simpler operations. Since the highest computational load in the adaptive antenna is the modem and the beamformer (mostly linear operations), fixed-point arithmetic is preferred for the prototype.

· **Clock rate**: the higher it is, the greater the number of instructions per second that can be executed, and the higher the computational capacity that can be obtained.

· **Computational capacity**: it is the maximum number of instructions per second that the processor can perform, and it is directly related to the clock rate, but also depends on the capability of the DSP to perform multiple operations in parallel.

Other features are usually considered when choosing a DSP, such as power consumption, but are less restrictive for our application.

In order to increase the computational capacity, a structure of various DSPs in parallel can be used. However, it must be considered that when many data are transferred the bottleneck is usually found in the buses that communicate multiple DSPs or other input/output devices.

Finally, in order to know the requirements in terms of required computational capacity for the prototype, a rough estimation was initially done considering all the modules to be implemented, presented in [198].

Taking into account all the previous information, a digital processing structure consisting of several 4-DSP boards (referred to as QUADs) were selected, all by Pentek. Each board (or QUAD) is named as Model 4292 Quad C6203 DSP VME Board [218], and it is formed by four C6203 DSPs by Texas Instruments, a shared global memory, four dedicated memories, several buses and interfaces for interconnection and input/output and one microcontroller. Each DSP operates at a clock rate of 300-MHz and uses fixed-point arithmetic. Every QUAD is capable of delivering a combined peak processing power of 9600 MIPS. According to the preliminary computation of computational requirements, and considering some extra overhead, up to six QUADs are needed for the adaptive antenna prototype.
In order to increase the data transfer rate between QUADs, a high-speed data bus called RaceWay Interlink has been used. This device is a high-speed backplane fabric capable of delivering 32-bit word transfers between VME boards, such as the QUADs presented previously. It provides multiple, simultaneous high-speed communication paths between DSPs, which make the bus a valuable asset to real-time applications. The bus is capable of communicating up to eight VME boards at a data transfer rate of 267 MB/s, which means an aggregate transfer rate up to 1,068 GB/s. In our case, where 6 QUADs were used, up to 801 MB/s could be transferred among QUADs if needed. However, thanks to the distribution of tasks among the DSPs (see below) the requirements of data transfer between QUADs are much smaller.

Figure 3-29 shows a top picture and a perspective picture of one of the QUAD boards used for the prototype. In Table 3-VII the main features of the DSP boards are shown.

**Figure 3-29. The 4292 4-DSP QUAD**

Table 3-VII summarizes the main features of the selected hardware for the prototype.
### Main features of selected hardware

<table>
<thead>
<tr>
<th>Front-ends</th>
<th>Feature</th>
<th>Requirement</th>
<th>Current value</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC vertical resolution</td>
<td>≥ 12 bits</td>
<td>12 bits</td>
<td></td>
</tr>
<tr>
<td>Maximum sampling frequency at ADC</td>
<td>≥ 10 Msps</td>
<td>105 Msps</td>
<td></td>
</tr>
<tr>
<td>Maximum input bandwidth (receiver)</td>
<td>≥ 5 MHz</td>
<td>80 MHz</td>
<td></td>
</tr>
<tr>
<td>Maximum accepted input power (receiver)</td>
<td>-</td>
<td>5 dBm</td>
<td></td>
</tr>
<tr>
<td>DAC accepted input bits</td>
<td>≥ 12 bits</td>
<td>12 bits</td>
<td></td>
</tr>
<tr>
<td>Maximum generated output frequency</td>
<td>≥ 5 MHz</td>
<td>80 MHz</td>
<td></td>
</tr>
<tr>
<td>Maximum input data rate at DAC</td>
<td>≥ 10 Msps</td>
<td>200 Msps</td>
<td></td>
</tr>
<tr>
<td>Maximum delivered output power (transmitter)</td>
<td>-</td>
<td>4 dBm</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>DSPs</th>
<th>Feature</th>
<th>Requirement</th>
<th>Current value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type of arithmetic</td>
<td>N/A</td>
<td>Fixed-point</td>
<td></td>
</tr>
<tr>
<td>Clock rate</td>
<td>High</td>
<td>300 MHz</td>
<td></td>
</tr>
<tr>
<td>Number of instructions per second (per DSP)</td>
<td>High</td>
<td>2400 MIPS</td>
<td></td>
</tr>
<tr>
<td>Number of DSPs per QUAD</td>
<td>N/A</td>
<td>4</td>
<td></td>
</tr>
</tbody>
</table>

Table 3-VII. Main features of selected hardware for the adaptive antenna prototype

#### 3.4.2.2 Overall hardware architecture

Once all the hardware equipment was selected, it was mounted on a rack and all the boards are interconnected. Figure 3-30 shows a general architecture of the hardware implementation, where the blocks for the two polarizations are identical. The digital processing module, formed by several processors, is common to both polarizations. The software-radio modules are marked with filled colour. Analog RF-to-IF and IF-to-RF chains are shown just for a better understanding of the architecture, but they are not explained here since they are out of the scope of this thesis.

The DSPs boards (QUADs) are connected to a standard VME backplane. In order to speed up communications, the RaceWay Interlink is also connected to the VME backplane. The interface between the front-ends and the DSP board are VIM-2 (Velocity Interface Mezzanine) buses. Finally, the analog modules are connected to the front-ends by means of standard SMA connectors.

For monitoring tasks, a personal computer can be connected to the digital processing module to control the process and allow viewing of key variables and parameters. The
communication between the software radio modules and the control PC is done through a dedicated PCI connection and a TCP/IP protocol.

Figure 3-31(a) shows the development system for software radio modules, whereas Figure 3-31(b) shows the test equipment. Finally, Figure 3-32(a) shows the full adaptive antenna system (RF, IF, software modules) and Figure 3-32(b) presents the antennas used in the prototype.
3.4.3 Signal processing modules for adaptive antenna

We include here a thorough explanation of the signal processing modules that constitute the digital processing stage. This module has been divided into several sub-modules: general synchronization, data user synchronization, MODEM (modulation-demodulation), weight computation and beamformer, which are implemented on DSP platform. In addition, the digital upconverter digital downconverters are detailed here, which are implemented on FPGA platform.

3.4.3.1 General synchronization sub-module.

Once the call from user has been initiated and one or several dedicated channels have been given to the user, the adaptive antenna must be able to read the control bits from DPCCH (dedicated physical control channel) so they are used in the adaptive algorithm and beamforming procedure. Moreover, in order to beamform the signal in downlink, demodulation and modulation of user signal is needed. As specified in 3GPP standards, each physical channel in W-CDMA is spread by combining two types of codes with complementary properties: OVSF (Orthogonal variable spreading factor) codes and scrambling codes (Gold codes). Thus, basic information needed in a W-CDMA modulation-demodulation process is the used codes and, like in any spread-spectrum technique, the timing reference [219].

We can distinguish here two main procedures that are needed: first of all, while the call is being initiated the user equipment receives through a control channel the information about
the dedicated channels that have just been associated to the user; this is the *call establishment procedure*, defined in 3GPP standards. The adaptive antenna must be able to follow this procedure and read the information about the just established physical dedicated channels that are required to perform afterwards synchronization and demodulation-modulation. Since this procedure is quite complex and require processing not only at physical layer but also at higher layers, a specific module have been developed for this task, and its operation is explained in a separate section (section 3.6).

The second procedure to be able to synchronize and demodulate the user information is the one needed to find the downlink time reference that shows the time instant where demodulation of user must be accomplished. In downlink, all the physical channels (common signalling channels and dedicated user channels) use the same time reference, that is, if the time reference of one channel is known, the timing of the other channels is automatically known. The procedure to find the common timing reference for all downlink channels is called *cell search procedure*. Typically, cell search procedure is completed after three steps: *slot synchronization*, *frame synchronization* and *code-group identification*, and finally scrambling code identification. Common signalling channels needed in this stage are SCH (Synchronization Channel) and the P-CPICH (Primary Common Pilot Channel).

First and second steps use SCH codes. During the first step, the cell slot synchronization is acquired; it can be done by correlating the signal received from the base station with the primary SCH codes by employing the coarse synchronization algorithm, as it will be explained in section 3.4.3.2. After the cell slot timing is achieved, the frame synchronization procedure is initiated. In this second step, the secondary SCH codes must be used. Once the combination of secondary SCH codes used by the base station is identified, it is possible to acquire the general frame synchronization for downlink and the primary code group of cell simultaneously.

Finally, the exact primary scrambling code used by the cell is determined in the third step. This search is limited to the set of eight different scrambling codes determined by the primary code group. The reference channel employed in this step is the P-CPICH, which is transmitted continuously over the entire cell. The P-CPICH is an unmodulated code channel, which is scrambled with the cell-specific primary scrambling code of the cell. The P-CPICH is unique for each cell. After the primary synchronization code has been identified, the cell search procedure is finished and it is possible to apply the general synchronization algorithm in downlink with the P-CPICH channel.

Unlike downlink, each user has a specific synchronization reference in the uplink. If MODEM knows the parameters of active users for uplink (obtained in the set-up stage, as explained in section 3.6), the synchronization scheme is very simple. For each user, the timing reference is extracted from the DPCCH, applying the coarse and fine synchronization algorithms directly as explained in next section.
3.4.3.2 User data synchronization sub-module

The timing information of the transmitted frame is essential in order to properly demodulate the despread signal. Even if there is a single chip duration error, the received spread spectrum signal can not be properly demodulated.

Once used codes in physical channels have been obtained (see section 3.6), the appropriate timing reference is extracted. Several synchronization methods for spread-spectrum communications can be found in the literature [220]-[222]. We have selected a two-step approach based on coarse and fine synchronization and decimation [223], for giving very good results in CDMA systems, having low computational complexity and allowing parallel computation of signal. Firstly, coarse synchronization or initial code acquisition is done, which accomplishes the synchronization of the received signal and the corresponding code, with an uncertainty of half a chip period ($\pm T_c/2$). Secondly, fine synchronization or code tracking is performed, to track and keep synchronization between the received signal and the code with accuracy always lower than half a chip period.

To perform the synchronization, the scrambling code properties are used. These codes have an autocorrelation function that reaches its maximum when the code and the received signal are aligned.

**Coarse synchronization**

As stated before, the objective of the coarse code synchronization is to achieve an initial code acquisition between received signal and the corresponding scrambling code. This is equivalent to matching the phase of the spreading signal with the code.

There are different general acquisition techniques [223]-[225]. In the *serial search* all the possible phases are tested one by one sequentially. The complexity for this method is low but the associated acquisition time is high. In the *parallel search*, all the possible phases are tested simultaneously. The complexity is higher but acquisition time is much lower than in serial search. An intermediate approach between the serial and parallel search strategies has been implemented in order to achieve the coarse synchronization with a moderate computational load, considering the complexity versus time-to-acquisition trade-off. A study of the computational load required by different implementation approaches is shown in Table 3-VIII.

<table>
<thead>
<tr>
<th>Branches</th>
<th>Clock Cycles/bit</th>
<th>Acquisition Time (number of frames)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Serial Search</td>
<td>1</td>
<td>3000</td>
</tr>
<tr>
<td>2</td>
<td>5700</td>
<td>0.5</td>
</tr>
<tr>
<td>3</td>
<td>8700</td>
<td>0.33</td>
</tr>
<tr>
<td>4</td>
<td>10800</td>
<td>0.25</td>
</tr>
<tr>
<td>10</td>
<td>27300</td>
<td>0.1</td>
</tr>
</tbody>
</table>

Table 3-VIII. Number of clock cycles and acquisition time for coarse synchronization algorithm.
Considering the capacity of the used DSP’s, the three-branches serial-parallel approach has been implemented. The block diagram of the coarse synchronization stage is shown in Figure 3-33.

![Figure 3-33. Block diagram of coarse synchronization.](image)

In the figure, several blocks can be distinguished: correlators, threshold generators, signal control modules and a scrambling code generator. The received match-filtered signal is correlated with different cycle-delayed code versions. The maximum correlation value from the branches is compared with the first threshold, \( \gamma_1 \), which is obtained taking into account the second maximum correlation value. In order to avoid situations in which the background noise may cause a wrong correlation which exceeds the first threshold, it is necessary to set another threshold to minimize this effect. This second threshold, \( \gamma_2 \), is calculated from the average of all the correlations except the maximum value. If the input signal surpasses both thresholds, then it is coarse-synchronized and fine synchronization is triggered.

**Fine synchronization**

The purpose of code tracking is to track synchronization with a good accuracy. Code tracking starts its operation only after coarse synchronization has been achieved. After coarse synchronization, a small phase error is still present. In order to correct this error, the loop structure shown in Figure 3-34 is used, which is a modified version of the early-late tracking module [224].
The first block is a decimator that selects the correct sample at the right time, depending on the correlation value. In the second step, the decimated signal is delayed or advanced half a chip period, creating the late, early and on-time branches. These three signals are correlated with the locally-generated scrambling code, and the maximum absolute value of the correlations is selected. According to this selection, the timing information is updated.

3.4.3.3 Demodulation in uplink and downlink

Once the timing information and scrambling and channelization codes are determined, UMTS physical channels can be demodulated.

In uplink, DPCCH is demodulated for each user and each antenna, in order to extract the pilot bits that will be used as the input signal in the adaptive algorithm to compute optimal beamforming weights. To complete this task, two operations must be carried out: the complex-valued signal is descrambled by a complex-valued scrambling code, $S_{dpcch,n}$, which identifies a user, and the signal is despread using the channelization code, $c_{n}$, which identifies the DPCCH channel. This process is shown in Figure 3-35.

In downlink, the DPCH (DPCCH+DPDCH) is demodulated. Firstly, the signal from Node B is descrambled by a complex-value scrambling code, $S_{dl,n}$, which identifies the cell and afterwards, the signal is despread through the correlation with a real-valued channelization code, $c_{ch,SF,n}$, which identifies the user in downlink. DPCCH and DPDCH bits, which are time-multiplexed, are obtained after this operation. Once the DPCH bits have been beamformed for
every user, the conversely operation is performed: spreading with $c_{ch, SF, n}$ and scrambled with $S_{dl,n}$. The block diagrams of the MODEM for downlink are shown in Figure 3-36(a) and Figure 3-36(b).

![Figure 3-35. Uplink demodulator diagram.](image)

![Figure 3-36. Downlink demodulator and modulator diagrams.](image)

### 3.4.3.4 Adaptive beamformer

Once synchronization has been achieved and demodulated DPCCH bits are available, the following stage is the adaptive beamforming. The aim of this module is to calculate the set of array weights that make the array output signal satisfy an optimization criterion. Apart from this computation, the beamforming module adequately combines the received signal vector in order to produce the spatially filtered W-CDMA signal as array output (for uplink) or input (for downlink).

In uplink, the received signal is used to compute the optimal beamforming weights. The received signal is weighted and combined, and the resulting signal is sent to the Node B. In downlink, the spatial information from uplink is used to beamform the signal to be transmitted through the multiantenna array. The signal from Node B is multiplied by the beamforming
downlink weights and sent to the corresponding antenna, where all the users are combined. Next paragraphs give a thorough explanation on the used algorithms and operation.

**Beamforming in uplink**

In uplink, the pilot bits from DPCCH that have been demodulated by the demodulation module are used to compute the optimal beamforming weights. The weights are adaptively computed, so for each new demodulated pilot bit the system updates the weight vector. The actual beamforming operation is performed by the combiner, which multiplies the input signals by the current beamforming weights at each time instant, and adds the result so a single stream of data is obtained as output to be sent to the Node B.

It is important to note that the combining (beamforming) operation is performed at chip rate for uplink, as opposed to bit rate: the output digital baseband signals, which are obtained from the antennas, are directly combined, without going through demodulation (descrambling and despreading). This way the system complexity is highly simplified, since there is no need for demodulation and modulation of each data channel of all the users. Moreover, a single stream of data is obtained as output (as opposed to the “one stream per user” approach in conventional beamformers), which complies with the plug-and-play approach proposed here. The cost is that a single beamforming vector must be used for all the intracell users, instead of one weight vector for each user. Thus, **partial interference cancellation** is performed, instead of the optimal total interference cancellation (see section 3.3.2). In addition, this method involves beamforming of all the uplink channels, both dedicated and common channels transmitted by the user.

Regarding the algorithm for weight computation, an adaptive algorithm was used since the allow real-time update of weights, and unlike the optimal Wiener-Hopf solution (which is unfeasible due to high computational cost and need of knowing statistical values) they are adequate for realistic implementation. Several conventional and novel adaptive algorithms for beamforming were studied and analysed (see section 3.2), but the well known NLMS algorithm was chosen for the implementation on DSP, due to its simplicity and low computational requirements. Several studies regarding its implementation on fixed-point arithmetic and selection of algorithm parameters were done; they are presented in subsequent sections.

In order to perform beamforming at chip rate, a single weight vector (or **total weight vector**) must be used. However, a different reference signal is known for each user, thus having multiple reference signals. Two possible methods have been proposed and studied: the method of **aggregated weights** and the method of **aggregated references**. In the first one, an optimal weight vector is computed for each user by using the adaptive algorithm with the corresponding reference signal, and the total weight vector is calculated as a linear combination of all the user weight vectors. This solution is depicted in Figure 3-37. The second one calculates first an
average total reference signal by combining all the references and all signals of the served users, and then computes directly the optimal total weight vector. The diagram of this solution is shown in Figure 3-38.

Figure 3-37. Diagram for computation of total weights with method *aggregated weights*.

Figure 3-38. Diagram for computation of total weights with method *aggregated reference*.

The two methods present advantages and drawbacks. The method based on aggregated reference presents a lower computational complexity, since the adaptive algorithm is performed
only once for all the users, as opposed to the method based on aggregated weights that needs to run the adaptive algorithm for each of the users separately. On the other side, the first method obtains information (weights) for each user in a separate way, so a weight vector can be associated to each user to obtain information about it (angle of arrival, etc…). These weights can be used in the downlink. In the case of the second method only a set of weights is obtained, with aggregated information about all the users together, so the only way to use this information for downlink would be to follow the same approach of partial cancellation and beamforming all the users together. Next section presents simulation results showing performance of the two methods, showing that they perform very similar in terms of SINR and final array factor. For the final implemented version, the method based on aggregated weights was selected, since a total interference cancellation was to be implemented in downlink and a set of weights is needed for each user.

Figure 3.39 shows the beamforming structure used in the uplink. Demodulated DPCCH bits from the MODEM and the received signal vector $\mathbf{x}(t)$ are the inputs of the beamforming module. These bits are used to acquire the slot synchronization, which is necessary to obtain the correct pilot bits that have to be used as the reference signal in the weight computation process. After that, the single-user weights are computed, and the common beamforming weight vector is calculated. The output of the beamforming process is obtained by multiplying the received signal vector by the common weight vector and combining the resulting signal vector, without the need of demodulating each user’s data channels. It must be noticed that the signal vector impinging in the array antenna is composed of contributions from several users, so that both the slot synchronization and the single-user weight computation must be executed in parallel for every served user. The process is repeated for every available pilot bit, so weights are continuously updated in real time.
Beamforming in downlink

Optimal downlink beamforming will minimize the interference received by other users and will enhance the useful signal power received by the desired user. Due to the frequency translation that appears in an FDD system, uplink and downlink communication scenarios are different. In the case of study, and as a consequence of the lack of downlink channel information, the calculation of transmission weights is all but an easy task.

One of the methods for estimating downlink weights from uplink channel information is the use of the uplink spatial covariance matrix [226]. Unfortunately, this approach involves the use of too many computational resources for a real-time operation to be implemented in fixed-point DSPs. In order to reduce the complexity of the implementation, the user weights obtained from uplink have been used to compute the user weights for downlink. Two approaches have been studied: using a corrected version of weights to account for the differences in frequency for uplink and downlink, or using the same weights as in uplink.
In the first approach, the correction is performed to account for different physical delays for the two frequencies (uplink and downlink frequencies). For a carrier, the spacing between uplink and downlink in the studied frequency band (band I) is fixed, with a value of $\Delta f=190$ MHz, according to 3GPP standard. The correction of weights will be given by next equation:

$$\tilde{W}_{DL} = \tilde{W}_{UL} \cdot e^{j\frac{\lambda_{UL}}{\lambda_{DL}}} \quad (3-5)$$

In the second approach, the same weights are used for uplink and downlink, as given in eq. (3-6).

$$\tilde{W}_{DL} = \tilde{W}_{UL} \quad (3-6)$$

After studying the two options, it was observed that both methods give similar results in terms of obtained array factor. Thus, the simplest approach was used for the real implementation, that is, equal weights for uplink and downlink.

As a result of frequency translation, the transmitting array factor for downlink is similar but not the same as in the uplink. Therefore, the synthesized beam will not be pointing exactly in the direction of the mobile user, and the interference received by other user will be reduced but not minimized.

In contrast to the uplink, a complete user separation can be performed in the downlink direction. Because of that, in the proposed downlink structure shown in Figure 3-40 single-user weight vectors calculated in the uplink are applied as transmit beamforming weights to each user separately. Therefore, downlink beamforming is much simpler than the uplink one, since the adaptive weight calculation is not required. Downlink beamforming is applied at bit rate, which results in a considerable reduction in the computational load, as far as the multiplier submodule is concerned, in comparison to the equivalent module for the uplink. Moreover, a total interference cancellation approach is achieved thanks to the individual user separation. As a drawback, this scheme increases the complexity of the modulation and demodulation module, as it must be performed for all the user channels (data and control).

In contrast to dedicated channels, broadcast information conveyed by common transport channels must be received by all the users in the cell. Therefore, the associated physical channels are transmitted to the whole sector through one of the antenna array elements, which is equivalent to the radiation pattern of a conventional sectored antenna. Consequently, these common transport channels are not beamformed.
3.4.3.5 Digital downconversion and ADCs

As previously mentioned, ADCs and digital downconversion are integrated in a single board or RX front-end. The digital downconversion consists in digitally mixing the input IF signal with a digital oscillator, and afterwards filtering the signal with a decimation filter. A diagram of the main parts of one receiving channel for one antenna is shown in Figure 3-41. The digital modules are programmed in the FPGA included in the board, as well as the subsequent signal formatting needed to properly communicate with the DSP. The mixer is already implemented in the FPGA, but several parameters can be selected, as the decimation rate (from 2 to 64). Of great interest is the option of synchronously resetting the digital oscillator generator, and also the different options for the sampling clock, which can be an internal crystal oscillator or an external source. This last option, together with the synchronous digital oscillator generation, allows for several RX channels to work synchronously, and was the selected option. A commercial signal generator (from Rohde&Schwarz) was used as external clock source.

The ADC sampling frequency must be carefully selected. It has to be a multiple of the UMTS baseband signal rate, $R=3.84 \text{ Mchip/s}$, multiplied by the number of samples per chip, which is $N_{spc}=4$ in this prototype. Neither 15.36 MHz nor 30.72 MHz can be used as sampling frequencies, since it would cause aliasing in the sampled signal. On the other side, the ADC features restrict the possible sampling frequency to a maximum of 100 MHz. Thus, $f_s = 61.44 \text{ MHz}$ has been chosen. The sampling frequency $f_s$ does not meet the Nyquist Theorem in a wide sense ($f_s$ is lower than $2 \cdot IF = 2 \cdot 44 \text{ MHz}$), but since the actual signal bandwidth is lower than IF (it is band-limited to 5 MHz), an undersampling strategy can be adopted, without losing
information of signal. Figure 3-42 represents the spectrum of a 5MHz signal after sampling at two different frequencies, for IF = 44 MHz.

Figure 3-41. Diagram of A/D converter and IQ demodulator for one channel

Case 1: $f_s=61.44$ MHz

Case 2: $f_s=30.72$ MHz

Figure 3-42. Representation of spectrum of sampled signal for two possible sampling frequencies (upper figure: $f_s=61.44=16\cdot R$, lower figure: $f_s=30.72=8\cdot R$).

3.4.3.6 Digital upconversion and DACs

Similarly to downconversion, an IQ modulator that performs upconversion is required before each DAC. Also the front-end solution has been adopted here, but instead of getting a flexible solution on FPGAs, an integrated circuit performs the operation. Nevertheless, the system allows to select the oscillator frequency and several other parameters (interpolation rate, etc…) at run time, which makes this solution also quite flexible. A block diagram of one channel can be seen in Figure 3-43. The interpolation rate $N$ is selected so the sampling
frequency is the same for uplink and downlink. This module also presents the option of using an internal clock to drive the DACs, or using an external source. The same source as for ADCs is used here, so that there is no drift from the two sampling clocks.

![Figure 3-43. D/A converter and IQ modulator.](image)

### 3.4.4 Code optimization and load distribution

Once the implemented algorithms and methods have been designed, developed and independently tested, a very important step is the integration of the modules and the intercommunication of information. In fact, data flows are one of the most demanding tasks in terms of computational load, as will be shown. This is due to the high rate that needs to be processed in real time, especially for uplink where beamforming is done at chip rate. As a result, code optimization to get parallel processing when possible and an optimal load distribution along the DSPs is a must in this implementation. General guidelines are given for a real-time implementation of smart antenna, and the results in terms of DSP load and solution for load distribution are presented.

#### 3.4.4.1 Maximum number of instruction per DSP

The analog received signal is sampled in the ADC to 4 samples per chip rate. Being spread with 256-bit channel codes, each DPCCH bit consists of 256 chips. Therefore, each DPCCH bit is formed of $256 \times 4 = 1024$ samples. The buffer that communicates ADC with DSP is configured so the samples are delivered in 1024-sample packets, which in general will not be a single DPCCH bit but part of two bits, since the input buffer is not necessary synchronized at bit rate. For real-time execution, the available time for processing each packet of 1024 samples is one DPCCH bit period, that is, $256$ (chips/bit)/$3.84 \times 10^6$ (chips/sec) = 66.67 $\mu$s. The clock rate of each DSP is $F_{DSP}=300$ MHz. Thus, the maximum number of DSP cycles available for processing along 1-bit time is:

$$\text{Num(Cycles/bit)}_{\text{MAX}} = F_{DSP} \cdot T_b = 300 \cdot 10^6 \cdot 66.67 \cdot 10^{-6} = 20000$$

(3-7)
Each DSP’s architecture has eight highly independent functional units (six ALUs of 32-/40-bits and two 16-bit multipliers). As a result, up to eight 32-bit instructions per cycle can be executed. Thus, each of the used DSP has a performance capability of up to 2400 MIPS on pipeline, and the maximum number of instructions that can be executed during a bit period are $20000 \cdot 8 = 160000$ instruction.

### 3.4.4.2 Evaluating computational load of each module

According to previous paragraph, in order to get a real time operation, signal processing must be performed by several DSPs in parallel so that the required computational load in each DSP to process signal received in a bit-period must be lower than 160000 instructions, or equivalently, 20000 DSP clock cycles per bit. In order to guarantee that this real-time constraint is achieved, a tool for measuring DSP clock cycles needed to execute certain code lines has been used: the STS module from the Instrumentation block, of DSP/BIOS [227].

DSP/BIOS is a scalable real-time kernel designed for applications that require real-time scheduling and instrumentation. It is provided by Texas Instrument along with Code Compose Studio, and offers several APIs which can be used to analyze and debug code for TI DSPs. It is divided into several modules, offering multiple real-time analysis tools. One of them is the Statistics Object Manager (STS), which allows to measure statistics in real-time running code. We have used it to measure the average and maximum number of clock cycles when running each of the adaptive antenna SW modules in an infinite loop, in order to check whether the real-time constraints were accomplished or not, and to optimally distribute the computational load along the DSPs.

Table 3-X shows the number of clock cycles per bit of each module after code optimization, measured by using STS DSP/BIOS module. We conclude that these values are always lower than the maximum number of clock cycles per bit for one DSP, so each DSP can perform at least one of the modules.

### 3.4.4.3 Code optimization and computational load of adaptive antenna modules

In a real-time implementation, optimizing the code to reduce the computational load is of great importance if a reasonable load and number of processing resources are to be used. We have followed the following steps for optimizing the code, including some of the recommendations of [228] and [229]:

1) **C code writing.** In order to speed up development, the C code written for simulation tests is reused as much as possible in this step.

2) **Compiling with maximum instruction reduction.** When compiling the C code, we choose the compilation options for reducing the execution time at a cost of increasing the required RAM.
3) **Use of DSP intrinsic operations.** C62x DSPLIB is a library designed for the C62x series of fixed-point DSPs from Texas Instrument that offer a number of instructions that are specifically optimized to take advantage of the parallel capabilities of the DSP core [230]. These functions are especially used in real-time applications, since their execution time is much lower than the C equivalent code. The main disadvantage is that they can only be used under certain restrictive conditions. In specific parts of code where intensive computation was done, and when using intrinsic operations was possible, they have been utilized.

The flow diagram in Figure 3-44 summarizes the previous steps.

![Flow diagram of optimization steps](image)

**Figure 3-44. Optimization steps flow diagram.**

After code optimization and programming, module load has been measured. Table 3-IX shows the complexity of each optimization step in clock cycles, time and clock cycles per second for the module of coarse synchronization. As it can be seen, the complexity of coarse synchronization module has been reduced two orders of magnitude after the third optimization step.

<table>
<thead>
<tr>
<th>Module</th>
<th>Optimization step</th>
<th>Clock Cycles/bit</th>
<th>Time (μs)</th>
<th>Millions Clock Cycles per second</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coarse synchronization</td>
<td>1</td>
<td>2571296</td>
<td>8571</td>
<td>38500</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>65732</td>
<td>219</td>
<td>986</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>8700</td>
<td>29</td>
<td>131</td>
</tr>
</tbody>
</table>

Table 3-IX. Reduction of the number of clock cycles in the coarse synchronization module.
Table 3-X illustrates the computational load of all the signal processing modules implemented on DSPs, when the three optimization steps have been applied. The achieved reduction in clock cycles when performing the three steps in code optimization on the other modules is, on average, two orders of magnitude too.

<table>
<thead>
<tr>
<th>Module</th>
<th>Clock Cycles/bit</th>
<th>Time (μs)</th>
<th>Million of Clock Cycles per second</th>
<th>%DSP capacity used</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coarse synchronization</td>
<td>8700</td>
<td>29</td>
<td>131</td>
<td>43.5 %</td>
</tr>
<tr>
<td>Fine synchronization and demodulation</td>
<td>6600</td>
<td>22</td>
<td>99</td>
<td>33 %</td>
</tr>
<tr>
<td>Slot synchronization</td>
<td>890</td>
<td>3</td>
<td>13</td>
<td>4.43 %</td>
</tr>
<tr>
<td>Single-user weight computation</td>
<td>830</td>
<td>3</td>
<td>13</td>
<td>4.17 %</td>
</tr>
<tr>
<td>Common weight computation</td>
<td>5200</td>
<td>17</td>
<td>78</td>
<td>26 %</td>
</tr>
<tr>
<td>Combiner</td>
<td>11500</td>
<td>38</td>
<td>174</td>
<td>58 %</td>
</tr>
</tbody>
</table>

Table 3-X. Computational load of synchronization, demodulation and beamforming modules.

3.4.4.4 Load distribution in DSPs

According to the required computational capacity for each module after code optimization, the distribution of load and tasks between DSPs was carried out.

As presented in the hardware description section, 6 QUAD boards have been used for signal processing, with 4 DSPs each. In order to properly design the load distribution between DSPs, several questions have been considered. To begin with, it has been taken into account that 2 independent polarizations should be processed, so the number of available DSPs for each one is 12. Despite this independence, the load cannot be divided into 3 QUADs per polarization. This is due to the need of five broadband receiver channels plus ADC and five broadband transmitter channels plus DAC per polarization. Receivers and transmitters boards consist of 2 channels each, which have to be associated to two DSPs in the same QUAD. As a result, receiving and transmitting channels must be considered as pairs, so that the possibility of using 3 independent QUADs per polarization is eliminated.

Another point to take into consideration is the association of a task per DSP as far as possible. In this way, not only the load distribution but also the data exchange between DSPs is more easily understood. As regards the data exchange between DSPs and QUADs, the tasks and load distribution has been designed aiming at reducing the number of data transfers between processors as much as possible. This makes the interconnection between DSPs simpler, since less synchronization for data exchange is needed. Data transfers between DSPs are preferred to those between QUADs, due to the higher complexity of transfer and synchronization in the last ones.

On this basis, the scheme in Figure 3-45 has been proposed and implemented. As it can be observed, the synchronization and demodulation of received signals from the 4 antennas in the uplink are processed in the same QUAD, hence avoiding the extra data exchange and the difficulties of making a correct synchronization if the signals would be received in different...
QUADs. Similarly, all the transmitted data are obtained and sent to the antennas in a single QUAD, at downlink. Due to computational cost restrictions, only 3 users can be processed with this hardware implementation. A higher number of users could be processed with more DPSs or with higher computational capacity ones. Thanks to the modular architecture, adding new users is not a complex task.

Figure 3-45. Load distribution.

3.4.5 Analysis of impairments due to non-ideal effects and proposed solutions for their mitigation

The implementation of the smart antenna prototype allows us to study the effect of some imperfections and impairments of the system due to non-ideal components or implementations, and permits the analysis of their effect in the system. These non-ideal effects
must be assumed in a real implementation of adaptive antenna, so their analysis and the study of possible solutions to mitigate them are of great interest.

**RF-IF phase and amplitude offset between branches**

The first and most likely impairment is non-identical branches. Due to array imperfections, mutual coupling, instability and differences in RF-IF modules, the signal vector that is fed into the adaptive beamforming module differs from the one received in the array from the propagation scenario. Because of that, the synthesized radiation pattern that is measured can be completely different from the expected one. For the implemented adaptive antenna, it was observed that the main problems were the coupling between chains and instability in time, since fixed differences in phase and amplitude offset can be compensated with a calibration module. In order to characterize the maximum allowable instability in the system several simulations were carried out, where variations in the signal were included, in both the phase and the amplitude. They were simulated as a Gaussian random variable, independent in each chain. The obtained radiation pattern when two sources (one desired user and one interfering user) are considered in the system was studied in two situations: with high instability in time and with low instability in time. Figure 3-46 shows the two situations, where several radiation patterns have been depicted simultaneously in order to compare the variation of the results.

As it can be noted, the instability in chains causes an important degradation of the radiation pattern, where the main effects are null filling and higher side lobes. After many simulations it was concluded that, in order to guarantee that a null filling in the radiation pattern will not be higher than -20 dB, the maximum allowable standard deviation in relative phase and amplitude was 5° in phase and 1 dB in amplitude. Measurements of the RF-IF branches in the adaptive antenna prototype showed that the system fulfills these maximum levels of instability.
In order to mitigate fixed differences between branches, a calibration module is proposed. Calibration is required because uplink adaptive weights are used to calculate downlink adaptive weights, so a good characterization of different antenna channels and coupling among these circuits is required. The calibration must be performed in the absence of interfering signals, and in a controlled scenario as a far field anechoic chamber. Calibration process provides the system with the uplink and downlink coupling matrices. The selected calibration procedure for uplink is schematically represented in Figure 3-47. Amplitude and phase differences between receiver chains are obtained by comparing the received signal with the reference signal vector, when a UMTS signal is transmitted from different angular positions. That is, if the user reaches the array through the broadside direction, the downconverted signal vector $\mathbf{x}'$ must be equal in amplitude and phase to $\mathbf{x}$, $\mathbf{x} = \mathbf{x}'$. However, due to the calibration errors, this condition is not satisfied. The objective of the calibration unit is to correct the received signal vector $\mathbf{x}'$, so that $\mathbf{x}' \approx \bar{\mathbf{x}}$ is satisfied. Measuring relative amplitude and phase offsets for different angular positions, the measured data allow us to solve an overdetermined linear system of equations, and to calculate the elements of the uplink calibration matrix. The values obtained from the calibration unit are saved and applied to the received signals before the adaptive beamforming.

![Diagram](image)

**Figure 3-47. Calibration procedure used in the uplink for correcting amplitude and phase differences between the antenna elements.**

Downlink calibration procedure is slightly different. A sinewave is used as transmitting signal, and different orthogonal combinations of weights are chosen for the four antenna branches. The receiver (Vector network analyser) compares the signal in the probe with one sample obtained from transmitting antenna 1 input through a directional coupler. Information is obtained in different angular positions of the probe, and again an overdetermined linear system is solved to obtain the downlink calibration matrix. This matrix is applied in a similar way as uplink calibration matrix.
**Frequency mismatch and module for phase tracking**

In many systems and also in the prototype presented here, a significant effect due to non-ideal systems is frequency mismatch in both the IF frequency and the sampling frequency that drives the ADC/DACs. Since the IF frequency for the input signal in the ADC is not exactly the expected value, the digital downconverter (which considers a fixed IF) offers as output a signal with a variable phase error, due to the frequencies differences. In Figure 3-49 this idea is depicted.

As a consequence of frequency mismatch, $\Delta f(t)$, a phase error appears, which is usually modelled as a simple linear equation:

$$\Delta \theta(t) = 2\pi \Delta f(t) \cdot t \quad (3-8)$$

The phase error was studied in the DPCCH demodulated bits for the desired user. Since these bits are used as a time reference for the adaptive algorithm in the beamformer, it is important to correct the possible frequency and phase error. Note that considering the expression above, if the frequency mismatch is approximately a constant value compared to the rest of the variables (as it is in a low mobility scenario with low Doppler frequency, as in our system), the phase error should be linearly dependent with time. However, when considering several users in the system, that was not the case in our system. This effect was studied with simulations, and it was observed that the demodulated bits not only suffer a constant phase error, but also a small deviation from the expected value, even when the noise level is very low.
small. This effect is mainly due to the residual signal of interfering users, which is not totally eliminated by the despreading operation in the demodulation of desired signal. To take into account this effect, a new term can be included in the above expression, so the phase error can be expressed as:

\[ \Delta \theta(t) = 2\pi \Delta f(t) \cdot t + \Delta \varepsilon_{\text{interf}}(t) \]  \hspace{1cm} (3-9)

The residual signal of interfering users, \( \Delta \varepsilon_{\text{interf}}(t) \), is a complex value and its effect in \( \Delta \theta(t) \) is not a constant value. In Figure 3-50 several subsequent demodulated bits are schematically represented with (right) and without (left) this effect.

If the signals of users are supposed to have zero mean, the residual interfering signal will have a zero mean effect as well, which means that the mean value of \( \Delta \theta \) computed in a certain number of samples will be approximately the same in both cases.

In order to track the phase variation in the demodulated bits, two schemes were considered:

- A bit-by-bit track module, which computed the phase deviation between demodulated bits (taking into account the possible sign change depending on the transmitted bit) and carried out on-time compensation.
- A block track module, which computed a mean value of the phase deviation and carried out a “late” compensation, that is, compensate the phase error in the current frame with the mean value computed in the previous frame.

Both possibilities were analysed with two methods: by direct simulation of the whole system and by processing samples taken from the real system, in an “off-line” way. The study showed that the second scheme offers a better performance, regarding the capability of adjusting the radiation pattern to obtain a null in the interference direction of arrival.

The phase track had to be implemented in the fixed-point platforms. The computation of the actual and desired phase values requires the calculation of trigonometric functions, which is not efficient in a fixed-point DSP. In order to obtain a real-time implementation, a look-up table...
with quantized values of phase and trigonometric functions was used, so the operations that imply a higher computational load are avoided. But the use of quantified values requires to select the look-up table size. Therefore some simulations were carried out to analyse the system performance with different number of quantized values.

![Global Radiation Pattern](image)

**Figure 3-51. Radiation pattern simulated for different quantized phase values in the phase track module.**

In Figure 3-51 the radiation pattern obtained by simulating the system with different number of phase values in the look-up table is shown. It can be observed that with only 12 possible values of phase (when values can be between 0° and 360°) the accuracy is not enough for the beamforming algorithm to be able to put a null in the interference direction, although the maximum gain is properly pointing at the desired user. This is due to the demodulation effect, which reduces the information about the interfering user, so with only 12 values of possible angles the adaptive algorithm loses track of it. When using a 360-value quantification, the obtained radiation pattern shows a very low side lobe in the interfering direction, so that this number of quantified phase values can be considered accurate enough for our application.
3.5 Novel methodology to evaluate adaptive antennas

3.5.1 Introduction

After design and implementation of the proposed adaptive antenna, the evaluation of its performance is the step to measure its features and it is needed to test the real advantages that a smart antenna system can offer in a mobile communication system. Simulations are of great importance in first steps of analysis of the configuration of a smart antenna system, such as simulations to evaluate different parameters to be selected in the system, or planning tools to compute the throughput or coverage in the system.

Despite their interest as analysis tool, simulations cannot include all effects that are present in real systems. Thus, measurement of features of an adaptive antenna prototype are of great value in order to evaluate the actual performance of the system.

Actual antenna measurements standards do not cover procedures to assess the performances of smart antennas and their associated adaptive algorithms. Those adaptive or reconfigurable algorithms are actually evaluated as a subsystem and the antenna array patterns are measured independently. Algorithm developers provide performances, usually based on simulations, intending to include radio channel impairments but usually forgetting important antenna features. On the other side, antenna engineers compute and measure classical parameters (gain, beam width, bandwidth, SLL, etc.) forgetting adaptive algorithms considerations. However, operators and terminal manufacturers need to know the full smart antenna system performance. Antenna pattern should be statistically measured implementing not only the real adaptive or reconfigurable algorithm but also, and what is more important, including the real modem signal. Measurements should be made not only in anechoic chambers but also in a real deployment scenario to include the radio channel impairments along with the adaptive antenna system.

To address this issue, we have proposed a novel measurement procedure for adaptive antennas, which includes two steps to allow measurement of “adaptive radiation patterns” including both the element effect and the adaptive algorithm effect. Two scenarios have been considered to carry out the evaluation measurements. The first one deals with the measurement in a controlled scenario inside an anechoic chamber, while the second set of measurements is performed under realistic conditions in an outdoor scenario. The procedure and some results for the implemented adaptive antenna are presented below.

3.5.2 Evaluation under controlled scenario: 2-steps measurements in an anechoic chamber

Conventional antennas are usually evaluated in an anechoic chamber, in terms of its radiation pattern, gain, etc. As passive elements with a fixed radiation pattern, the measurement of conventional antennas is well known and several standards for different types of
measurement conditions and antennas exist. These measurements in anechoic chamber are generally based on the use of a reference antenna or probe which transmits (or receives) a sinewave signal, while the antenna under test is rotated or moved along a grid to cover the angular dimension(s) to be measured.

Unfortunately, this method is not valid for an adaptive antenna. First of all, an adaptive antenna as the one implemented here will be able to beamform a user only if it is being served by a Node B, that is, it will point a beam to UMTS sources (not tones) that are using physical dedicated channels. Sinewave tones will be considered as interferences by the adaptive antenna, so measuring its radiation pattern in a conventional way will only give a null diagram. Secondly, the radiation pattern of the adaptive antenna would change according to the environment, so the system will not be able to measure the real radiation pattern with conventional method even if a UMTS generator is used as source.

In order to be able to measure the radiation pattern of the adaptive antenna under controlled scenarios, a new two-step measurement method is defined according to the requirements of the new proposed system. In the first step, the situation to be tested is created by using UMTS generators to simulate desired and interfering users. The desired user uses predefined physical parameters (scrambling code, etc…) that simulate the parameters that would be allocated by the Node B in a call establishment, and which are also known by the adaptive antenna. In this scenario, the smart antenna calculates the adaptive weights to maximize the SINR (Figure 3-52). Thus, this situation is equivalent to an “operation of full adaptive operation”.

**Figure 3-52. First step of procedure for evaluating adaptive antenna.**
In the second step, the adaptive weights are kept at fixed values, specifically the values that were obtained in the first step. Therefore, for this second situation a special operation is used: “situation of frozen operation”, which means that the weights are not updated any more. This is equivalent to switch off the weight computation module. This way, the synthesized radiation pattern can be measured in a conventional way.

![Figure 3-53. Second step of procedure for evaluating adaptive antenna.](image)

Measurements have been carried out in a chamber where the far-field condition is satisfied \((d>2R^2/\lambda \sim 5.2 \text{ m})\). A positioning device is used to roll the smart antenna over the azimuth domain, so the produced radiation pattern is measured in 2D (azimuth angle). In order to test the user beamforming and interference cancelling capacity of the prototype, one interfering user is introduced into the measurement system. A desired user is located in several directions with respect to the broadside of the adaptive array. Also the interfering user is placed in different angular positions relative to the desired user direction. Only one interfering user is used, mainly due to the need of expensive UMTS signal generators to simulate them.

Some results are included here to show the performance of the prototype. Figure 3-54 depicts the measured radiation patterns for one desired user in the broadside direction and one interfering user in two possible angular directions, 18 and 10 degrees. These patterns are compared with the idealized pattern of a sectored antenna showing cancellations in the interfering directions obtained with the adaptive antenna higher than 12 dB.
3.5.3 Evaluation under realistic scenario

The last set of measurements is carried out in a more realistic environment, trying to characterize the performance of the antenna with the actual propagation channel. The objective is the estimation of increase in gain ($\Delta G$) compared with a conventional sectored antenna, (working only with the desired user) and the increase in SIR ($\Delta \text{SIR}$) compared with a conventional sectored antenna (working with one desired user and one interfering user). The antenna is set in a sub-urban site and two users are located in different positions around this site (Figure 9).

![Image](image_url)

**Figure 3-54.** Measured radiation pattern with interfering users in $18^\circ$ and $10^\circ$ azimuth angle.

Once one location of interference, desired signal and adaptive antenna is chosen, the next procedure is followed.
For $\Delta G$ measurement:

A. The signal from desired user is transmitted, and the adaptive antenna performs the adaptive process, obtaining the adaptive weights in uplink and downlink.

B. The system is switched to “frozen operation” (weights are not updated) and the received power ($S_a$) in uplink and downlink is measured with a spectrum analyzer.

C. The adaptive antenna is replaced by a conventional sectored antenna and the received signals ($S_c$) in uplink and downlink are measured.

D. $\Delta G$ is obtained: $\Delta G (\text{dB}) = S_a (\text{dBm}) - S_c (\text{dBm})$.

For $\Delta C/I$ measurement is quite similar, but an interference is included:

A. Signals from desired and interfering users are transmitted, and the adaptive antenna performs the adaptive process, obtaining the adaptive weights in uplink and downlink for this new situation.

B. The system is switched to “frozen operation” (weights are not updated) and the received power in uplink and downlink is measured with a spectrum analyzer, first with just the desired user as transmitter ($S_a$) and second with just the interfering user as transmitter ($I_a$). SIR for the adaptive antenna is computed as:

$$\text{SIR}_a (\text{dB}) = S_a (\text{dBm}) - I_a (\text{dBm}) \quad (3-10)$$

C. The adaptive antenna is replaced by a conventional sectored antenna and received signals in uplink and downlink are measured. The signal to interference ratio for the sectored antenna is computed, $\text{SIR}_c$.

D. $\Delta \text{SIR}$ is obtained: $\Delta \text{SIR} (\text{dB}) = (S/I)_a (\text{dB}) - (S/I)_c (\text{dB})$

As an example, Table 3-XI shows the results obtained with a measurement following this procedure, comparing the theoretical or simulated results with the results estimated through the measurement of the radiation pattern.

<table>
<thead>
<tr>
<th></th>
<th>$\Delta G$</th>
<th>$\Delta \text{SIR}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Theoretical/simulated</td>
<td>6.2 dB</td>
<td>14.4 dB</td>
</tr>
<tr>
<td>Calculated from radiation patterns</td>
<td>-</td>
<td>14.5 dB</td>
</tr>
<tr>
<td>Measured</td>
<td>6 dB</td>
<td>12.5 dB</td>
</tr>
</tbody>
</table>

Table 3-XI. Measured values for increase in gain and SIR when using the adaptive antenna.

We see that the adaptive antenna performs in general very well, and that the measured results are very close to the expected results in terms of increase in gain and SIR.
3.6 Set-up stage and modules to operate in a real 3G network

The signal processing operation of the implemented adaptive antenna has been explained in detail in section 3.4.3. However, it was stated that before being able to process the user information it is required that one (or more) dedicated channel is associated to the user. The proposed smart antenna does not foresee beamforming users that are not using dedicated channels, since the traffic for control channels is assumed lower. Beamforming for high-speed data packet channels (HSPA) is a very interesting research line that has been left for future investigation. In conclusion, before the adaptive antenna is able to compute the optimal beamforming weights and actually perform beamforming, it is needed to read some parameters of the dedicated physical channels that are dynamically assigned by the network to the user. This task is accomplished by the set-up stage. Since it is a quite complex task and not only physical channels have to be processed but higher layer channels, we have considered it clearer to explain this stage in a separate section.

Figure 3-56 represents the modular diagram of the signal processing part, where the set-up module has been included and its input-output relations with other modules are highlighted. Inputs are the signal sent by the Node B and the received signal from the user, since several channels need to be read to obtain certain parameters. As outputs, the obtained parameters are communicated to synchronization and modem modules in both uplink and downlink.

This set-up module has been implemented as a proof-of-concept module, so that it is shown that the adaptive antenna can be deployed in a real UMTS network. A more detailed explanation of the implementation of the set-up module can be found in [231].

![Figure 3-56. Set-up module and relation with other signal processing modules.](image-url)
3.6.1 Need for set-up in the plug-and-play adaptive antenna

As stated above, in order to fulfil the desired transparent operation in the system, several user parameters must be known. These values are specified by the network when the user asks for a communication and a radio bearer is established. A first possibility would be to perform a “blind search”, which consists in testing all the possible values until finding the correct one. Despite its conceptual simplicity, this method is unfeasible in a real case, where the number of possible scrambling codes is not negligible, since the time to find the required parameters would be unacceptable for a real-time system. Thus, a set-up module like the one proposed in this paper is needed. Its objective is to “read” or decode the information sent by the network to the UE, containing certain physical configuration values. Since this information is included in radio bearer, which are entities belonging to RRC layer, the whole protocol tower, from physical layer to RRC layer, must be read.

3.6.2 Call establishment in UMTS

The establishment of a dedicated channel involves the UE transition to CELL_DCH state. This state can be reached from two possible modes: CONNECTED mode (CELL_FACH state) or IDLE mode. Thus, 2 possible transition cases can be considered, as depicted in Figure 3-57. In this figure, the channels that are involved in each message are mentioned for the three layers: logic channel, which is mapped onto a transport channel and this one onto a physical channel. Different options for mapping of channels have to be considered, as stated in 3GPP standard.

As stated in 3GPP standards, in order to establish a radio bearer the following steps are carried out by the UE. They are summarized below:

A. Obtaining system information. The System Information (SI) transports the required information to properly configure several channels in the system. The UE shall read this information to be able to demodulate or modulate other channels. The SI is periodically sent to
all the UE in the broadcast channel (BCCH). This channel is mapped onto the transport channel BCH, and this one onto the physical channel P-CCPCH. Since the SI is necessary for all the UEs aiming to enter the system, it follows a fixed and known structure so it can be understood by all UEs. MAC and RLC layers follow a transparent mode, so they do not attach any header. In RRC layer, the information is structured in different blocks. A main block, namely the Master Information Block, contains the information about the structure of the subsequent SI blocks (SIBs), such as existence, location and frequency of repetition. In our case, the specific SIB needed to perform the adaptive antenna setup is the SIB number 5, which includes the information regarding the common physical channels in the cell, such as the secondary common control physical channel (S-CCPCH) information.

B. Reception of message “RRC Connection Setup”. This message contains the information about the dedicated radio link that will be used to receive a radio bearer where the setup data can be read. It is transmitted in logical channel CCCH (common control channel), mapped onto transport channel FACH (Forward Link Access Channel) and physical channel S-CCPCH. The reception and demodulation of this physical channel is possible thanks to the previous information obtained in step 1. Layers MAC and RLC functions have to be considered in order to obtain the RRC message. After reception of this message, the UE enters in CONNECTED mode and a dedicated channel is assigned to the user. Depending on local or network configuration, the UE may enter in CELL_FACH state or in CELL_DCH state. In the former case, the FACH transport channel will be used to send the message “Radio Bearer Setup”. In the latter case, where the UE transition is directly made from IDLE mode to CELL_DCH state of CONNECTED mode, the message “RRC Connection Setup” allows to establish a temporal dedicated channel, which will be used by the network to send the message “Radio Bearer Setup” to assign a final dedicated channel to the user.

C. Reception of message “Radio Bearer Setup”. With this message, the network assigns a dedicated data channel or radio access bearer to the user. It is sent in logical channel DCCH (dedicated control channel), which can be mapped onto either a DCH transport channel (CELL_DCH case) or the FACH transport channel (CELL_FACH case). The parameters that were obtained in step 2 are used to receive and properly demodulate the corresponding channel.

3.6.3 Proposed set-up for adaptive antenna in a UMTS network

Table 3-XII shows the parameters needed by the signal processing modules in the implemented adaptive antenna, divided into two groups: uplink and downlink parameters. They are assigned by the network when a dedicated channel is established, via a “Radio Bearer Setup” message. In order to simplify the setup module, several simplifications were considered in the set-up module. They are outlined below.
Due to its lower establishment time, the case of transition to CELL_FACH state is more common than the one to CELL_DCH state in real networks. Therefore, only this possibility is considered in the setup module.

In the usual commercial implementations of real networks for operators, there are several parameters for common channels that can be considered as fixed and known values. This fact has been used to simplify the setup module, by taking them as predetermined values. Specifically, the physical parameters for the S-CCPCH (SF, code number, number of pilot bits, TFCI existence and timing offset) has been considered as fixed parameters, whose values are taken from basic test cases from 3GPP standard [232] and checked against the real implementation in an operator network, namely the version 3 of UMTS network of Vodafone Spain, which collaborated in this study.

<table>
<thead>
<tr>
<th>Link</th>
<th>Parameter</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Type and value of scrambling code (or seed) associated to a dedicated channel.</td>
</tr>
<tr>
<td>Uplink</td>
<td>DPCCH Slot format (Number of FBI bits, TFCI existence).</td>
</tr>
<tr>
<td>Downlink</td>
<td>Spreading code, associated to the user.</td>
</tr>
<tr>
<td></td>
<td>Spreading factor of data channel.</td>
</tr>
<tr>
<td></td>
<td>Time offset between data channel and primary common control channel P-CCPCH.</td>
</tr>
</tbody>
</table>

Table 3-XII. Required set-up parameters in the adaptive antenna

With these simplifications in mind, the final implementation for the proposed setup module consists in 4 steps:

1. **Decoding the physical channel S-CCPCH**, by following the recommended values in [232].

2. **Reading the layer MAC header**, with information about the user and the logical channel. In order to do so, a simplified MAC interpreter with different blocks was developed.

3. **Reading the layer RLC header**. In order to do so, a simplified RLC interpreter with different blocks was developed. Several header fields must be read, such as the ones indicating the length of each RLC block, the type of PDU, etc. A reassembly of the RLC blocks is done in order to obtain a complete unit to be sent to higher layers.

Three operation modes may be used in this layer, namely the transparent mode (TM), the unacknowledged mode (UM) and acknowledged mode (AM). Each one differs from the rest in the header fields. In the case of sending the Radio Bearer Setup module, only UM and AM are allowed by the 3GPP specifications. The actual mode is assigned by higher layers. Due to the importance of the Radio Bearer Setup message, and considering the difficulty of reading the instruction where the type of mode to be used is assigned, it has been assumed that AM will be
used. It may be noticed that UM functions are included in AM, since the latter is a more complex mode. Thus, with this implementation both modes could be received and decoded. Following the AM concept, a reception window must be included in the system, capable of storing a certain number of blocks, in order to recover a block if it is lost.

4. Decoding the RRC information. Once all the blocks that constitute a DCCH are assembled by RLC layer, the resulting PDU is sent to RRC layer. Information in RRC layer is coded following the ASN-1 code (Abstract Syntax Notation 1) [233]. ASN-1 is a very efficient codification, used to define network messages [234]. As a drawback of its efficiency, the implementation of codecs for ASN-1 is very complex.

Several commercial ASN-1 coders and decoders can be found. However, their use in the presented setup module is not feasible, due to constraints in DSP memory. A specific decoder has been developed for the particular messages that are needed in the setup module. It has been optimized for the application and the DSP platform, so the computational complexity is minimized.

The main shortcoming of this set-up is its lack of generality, since it has been designed and implemented taking into account several simplifications, as stated above. This is quite common in 3GPP due to the high complexity of the standard (even commercial base stations only consider certain options of all the possibilities stated in 3GPP standards). A shortcoming in our implementation is that the messages in upper layers have to be sent without codification (which in fact is included as an option in 3GPP standards), since the key to code and decode the signal is obtained from parameters stored in the user SIM and the network operator, and are not available for the adaptive antenna. This is required only during call establishment, and can be reconfigured later once the physical parameters have been read by the set-up module. This way the signal is secured without affecting negatively the adaptive antenna operation.

3.6.4 Real-time DSP implementation

3.6.4.1 Implemented modules

The realized set-up module consists of 4 sub-modules, each one implementing the functions of a layer. Their operation is summarized below.

- Physical sub-module

FACH is mapped onto S-CCPCH. This physical channel contains three transport channel: one PCH (Paging Channel) and two FACHs. Figure 3-58 shows the coding and multiplexing steps for downlink [235] used for our case of PCH and FACH configuration.
Thus, the aim of the physical module is to undo the steps of the multiplexing and coding structure. The TFCI bits, which are encoded using a (32, 10) sub-code of the second order Reed-Muller code, are decoded from the received frames. This field indicates the transport format combination of different transport channel (TrCHs) used in the UE. The next stage is to quantize the received bits for later soft Viterbi decoding. These values are de-interleaved and the discontinuous transmission (DTX) indication bits are extracted. Taking into account the decoded TFCI, demultiplexing of transport channels is performed and we only consider FACH 2 bits to undo the rate matching, to decode with the Viterbi soft algorithm, and to subtract the CRC field for each of the possible transport blocks.

**MAC sub-module**

It analyses the MAC header of the received transport block, and checks its main fields, such as type and number of logical channel or user identity. This module carries out a user control, since there is a number of maximum allowed users processed by the prototype. When this number is achieved, no more users are accepted in the system.

**RLC sub-module**
It analyses the RLC header. Its main task is to assemble the received segments that form a logical channel block, and to manage the order in which they arrive. In order to prevent losing some segments while waiting for a delayed one, a reception window has been implemented.

- **RRC sub-module.**

Once the whole DCCH channel has been received and properly assembled, RRC sub-module is run. A specific ASN-1 decoder has been implemented, so the module carries out the decoding of the Radio Bearer Setup message and obtains the required parameters (Table 3-XII), which are stored in global variables.

### 3.6.4.2 Real-time operation

For transparency purposes, the setup module must be able to intercept and demodulate all the Radio Bearer Setup messages sent by the network to the users. The module must work in a real-time basis, so the adaptive antenna can be updated every time a new user enters the system. Thus, it is clear that the setup module must analyse the received signals continuously.

The setup module has been implemented in another QUAD with TI C6203 DSPs, as the ones presented in hardware platform section. Two DSPs have been used: the physical layer has been implemented in the first one while the MAC, RLC and RRC layers have been implemented in the second one.

Taking into account that an S-CCPCH physical frame lasts 10 ms, and that in a frame up to 2 blocks must be read and decoded, we must guarantee that the jointly execution time of MAC, RLC and RRC modules does not exceeds 5 ms. Table 3-XIII summarizes the maximum execution time required for each module, which has been measured with the same tool as the other signal processing modules (STS, from DSP/BIOS, as explained in section 3.4.4.2). The results show that the required time to analyse a block is much lower than 5 ms, so the real-time constraints are fulfilled.

<table>
<thead>
<tr>
<th>Module</th>
<th>Max. execution time (μs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MAC layer</td>
<td>3.05</td>
</tr>
<tr>
<td>RLC layer</td>
<td>200</td>
</tr>
<tr>
<td>RRC layer</td>
<td>184.43</td>
</tr>
<tr>
<td>Total</td>
<td>387.48</td>
</tr>
</tbody>
</table>

**Table 3-XIII. Maximum execution time per module**

### 3.6.5 Test results of set-up module

After testing each module separately and measuring the execution time for each one, the whole setup module was tested under a real environment and with a real signal. Two types of tests were carried out: preliminary off-line tests and final on-line tests.
To perform the off-line tests, the received signal from a UMTS Node-B was stored, so several frames were obtained to be processed in an off-line basis. A UE for measurements and tests (TEMS) was used to realize several calls, so the messages sent by the network to initiate the call could be stored. The post-process offered as result the required parameters specified in Table 3-XII, which were checked against the real values assigned by the network.

The on-line tests were carried out in collaboration with Vodafone Spain, one of the Spanish 3G operators. In the on-line tests, the setup module was switched on in a cell with a Node-B base station, and several calls were made with the TEMS, which was configured with the parameters required by Vodafone network. At the moment of tests, this network allows the use of uncoded messages in higher layers, as needed for proper working of the implemented set-up module. The set-up system had to recognize each call and collect the required parameters in an on-line way. As a result, the values of the parameters were obtained during the establishment of the call, and stored to be used by the adaptive antenna. Several tests were performed, to check the set-up module operation with up to 3 calls (users). The tests were successful in this network. As an example, in Table 3-XIV the acquired values for a call are shown; matching up with the actual parameters in the network was checked.

<table>
<thead>
<tr>
<th>Downlink</th>
<th>Uplink</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frame offset</td>
<td>Code number</td>
</tr>
<tr>
<td>5632</td>
<td>128</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Uplink</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scrambling</td>
</tr>
<tr>
<td>Code Type Code seed Num. DPDCHs SF TFCI existence</td>
</tr>
<tr>
<td>1 1199596 1 64 1</td>
</tr>
</tbody>
</table>

Table 3-XIV. Acquired set-up values, for one test.
3.7 General conclusions, contributions and further research on adaptive antennas for mobile system

A prototype of an adaptive antenna for UMTS applications has been designed, implemented and evaluated. Along this work, different aspects have been analysed, multiple parameters and algorithms have been studied and selected and several lessons have been learnt. The main conclusions are drawn here, as well as the contributions of this thesis and the proposed further aspects to be investigated.

In a first step, some adaptive algorithms were theoretically studied in this thesis, in order to know which one would be suitable for implementation in the prototype. Our analysis shows that both the normalized mean square error (NLMS) and recursive least square (RLS) are easily applicable in UMTS scenarios, if the dedicated pilot bits of the user are known. Simulations of a UMTS scenario with several users and interferences show that using the adaptive antenna with four elements and NLMS or RLS offer a significant reduction in BER when compared with the results from a conventional sectored antenna, with up to an order of magnitude of improvement (up to 10 times lower BER) in the best cases. Although RLS slightly outperform NLMS in terms of average BER in static scenarios, it has been observed that in order to obtain a fast and good evolution a high forgetting factor is needed in static scenarios, which is not interesting for fast changing scenarios since the ability to adapt to changes is highly reduced. Moreover, we have shown that the number or required operations for RLS in a 4-antenna array is approx. 20 times higher than the ones required for NLMS, which means that a much more powerful hardware would be needed to implement an RLS-based adaptive antenna. As a conclusion, NLMS was chosen as the adaptive algorithm for the implementation of this prototype. Obviously, as novel powerful SDR hardware appear in the market, they could be used in the future to improve the adaptive algorithm module in an adaptive antenna with RLS or more complex algorithms.

Conventional adaptive antennas for UMTS assume that the beamforming process takes place after demodulation (before modulation for downlink), that is, at bit level. As a result, a specific “smart” Node B is needed to be used with a conventional adaptive antenna. To overcome this issue, we have presented a novel architecture that uses standard Uu link to communicate with the Node B, reducing cost and time to deployment. This concept (plug-and-play), together with the partial interference cancellation, has been analysed and its significant interest shown. We have shown, via simulations, that the expected average performance of the adaptive plug-and-play antenna is always better than the one for a sectored antenna, in terms of BER. When a high number of intracell users are present, the BER improvement is lower, while the best improvement is achieved in high-interference level scenarios, as hot spots.

After theoretical study of algorithms and simulation of results, different aspects of implementation were addressed. A Software-Define Radio architecture was chosen, and a DSP-
FPGA-based platform was used to implement the signal processing modules. DSP development showed a high flexibility and offered a rapid prototyping solution once the specific HW platform is known. However, DSPs presented a poor processing capacity for very demanding operations as correlations and filtering at chip level, which makes us recommend the use of FPGA for these tasks and DSP for more complex but less exhaustive operations. The experience showed us that a quite demanding task that is not always considered when estimating computational load is data communication between input/output buffers and DSPs, as well as between DSPs when needed. This data streaming meant close to a fifth of the computational load for some DSPs in our implementation, after code optimization. Thus, it is of paramount importance to properly distribute the computational load in DSPs in order to reduce the requirements of large packets of data to be exchanged.

The prototype has given the chance to study impairments due to non-ideal implementation, of great importance in order to analyse possible problems that may appear in a real system. Several effects were studied, such as frequency mismatch and non-ideal RF modules. Analysis of measured signals and the use of simulations to emulate the non-ideal aspects have shown as a powerful tool to search for solutions to mitigate the problem.

After implementation of the real-time prototype, evaluation of the performance was required. We noticed that evaluation methods for conventional antennas were not applicable for adaptive antennas, should their “smart” operation be tested. A novel two-step methodology has been presented and the results in a controlled scenario (anechoic chamber) and a realistic scenario (outdoor) were measured. The implemented adaptive antenna showed an improvement of up to 6 dB in gain when compared with a conventional sectored antenna, and the measured radiation patterns showed nulls at the direction of arrival of interferences of at least 15 dB cancellation level.

Finally, the required steps to integrate the adaptive antenna in a real UMTS network were studied, and a set-up module was proposed and implemented on the same SDR platform. Results of integration tests with a real operator showed that the integration was possible for the network that was deployed at that moment, when some simplifications are assumed. As a shortcoming of the set-up module, this work showed us that a problem to be solved is the way to communicate the user cipher code to the adaptive antenna. Higher level communications may be encrypted by a user-specific code, which is only known by the user and the specific operator that serves the user. Some communications are allowed to work without this cipher, but those that are encrypted would not be properly processed by the adaptive antenna.

**Contributions**

The contributions of this chapter of the thesis to the current state of the art can be summarized as follows:
A novel adaptive antenna has been implemented, aiming at addressing some of the main reasons that prevent smart antennas to be widely adopted: high complexity and difficulty of integration with current base stations. The plug-and-play concept has been applied to the prototype. A beamforming scheme for UMTS that is based on partial interference cancellation has been proposed, which has been firstly studied by means of simulations and afterwards implemented in an SDR platform. Several adaptive algorithms applied to beamforming have been studied, and their advantages and drawbacks to be used in the prototype have been highlighted. Several simulation results in terms of improvement of BER and SINR when using the adaptive antenna have been presented, as well as some measurement results. Many details of the real implementation and the most interesting aspects to be taken into account have been given.

The real-time prototype has been used to analyse some impairments that appear due to non-ideal characteristics of the real implementation. We have studied some of them, and have proposed some ways to mitigate their effect or evaluate it. Of key importance was the frequency mismatch, due to different frequencies in transmitter and receiver. It was solved with a tracking module that operated at a user-level, but calculated the average frequency drift by computing an average value to reduce the effect of interferences in frequency estimation.

A novel methodology to evaluate and measure adaptive antennas has been proposed. It is based on a two-step method, first computing the optimal weights to beamform the signal in a “running state” and afterwards measuring the antenna performance in a “frozen state”. This way the problem of adding interferences to the antenna operation while measuring is avoided. The implemented prototype was measured with this method, showing the good operation of the method.

A setup module including the steps and processing of messages required to allow for the adaptive antenna to be integrated in a real network operator has been proposed and implemented. The setup module recognizes each new user in the system, it retrieves some user parameters that are necessary for the adaptive antenna to work properly, and it reports them to the rest of the processing modules in the system. The setup operation has been tested with a real UE and Node-B, showing that the module is capable of extracting the required values for the adaptive antenna to work transparently in the system.

**Further research**

After some years of research in smart antennas, it seems clear that they show good properties in some scenarios (as the ones with transmitter and receiver in line of sight, or the
ones with small angular spread and few multipath contributions), but it does not work properly in scenarios with high angle spread in transmitted or received signals or scenarios with many multipath contributions with small time spread (such as indoor, or dense urban scenarios). This implies several effects, such as an introduction of some uncorrelation between antennas, which causes a significant reduction in performance of the beamforming algorithms or even its malfunctioning. To take advantage of multipath properties and other spatial characteristics of some environments, some algorithms as spatial multiplexing and spatial diversity algorithms have recently been proposed, and are currently being standardized for some applications as WLAN networks and new versions of 3GPP.

Regarding the plug-and-play adaptive antenna, the main shortcoming to be overcome is the way to read or communicate the cipher code from the user to the adaptive antenna. Since this parameter is known by the specific operator and the SIM of the user, it seems difficult to be known unless the Node B informs the adaptive antenna, but that would require some special communication between Node B and adaptive antenna, which reduces the transparent operation. Thus, this is still an open issue to be solved.

Finally, novel adaptive algorithms are been proposed every day. The adaptive antenna could benefit from them if a good effort in optimizing some of them and reducing their complexity is added.
4

RAPID PROTOTYPING FOR MIMO:
Design, implementation and measurements with a WLAN MIMO testbed

4.1 Introduction and motivation ................................................................. 150

4.2 Design and implementation of a MIMO testbed ............................... 151
  4.2.1 Design for multi-purpose scenarios and application .................... 151
  4.2.2 General architecture..................................................................... 153
  4.2.3 Modular description................................................................. 155
    4.2.3.1 Antenna modules .............................................................. 155
    4.2.3.2 RF-FI modules................................................................. 159
    4.2.3.3 Signal processing modules ................................................. 162
    4.2.3.4 User interface and post-processing.................................... 165
    4.2.3.5 Main features of final prototype ....................................... 170

4.3 MIMO channel measurements: Indoor and Outdoor-to-Indoor channel characterization including polarization effect ........................................ 173
  4.3.1 Motivation and progress beyond the state of the art ................. 173
  4.3.2 Measurement set-up................................................................. 173
  4.3.3 Approaches for capacity computation considering channel normalization ................................................................. 176
  4.3.4 Power computation and coverage maps .................................... 178
  4.3.5 Capacity results for indoor and outdoor-to-indoor scenarios ....... 180

4.4 General conclusions, contributions and further research on MIMO prototyping and dual-polarized MIMO channel measurements. .......... 183
A vast contribution to the theoretical field in MIMO system has been made by the research community since the idea was first presented. Multiple algorithms have been proposed, designed for taking advantage of the spatial diversity offered by the use of multiple elements, from the information theory point of view. The theoretical capacity of these systems has also been studied, under many different assumptions. Moreover, complex processing modules for full communication systems with multiple antennas have also been designed, and different channel models have been developed.

However, in order to have a better knowledge of the MIMO system and to study the effects of real components and implementation, measurements with prototypes are of great interest. This chapter presents the design and implementation of a MIMO testbed, and the results from a measurement campaign accomplished with it are also shown.
4.1 Introduction and motivation

Since MIMO systems begin to become a “hot topic”, many research institutions and universities have been working towards developing some type of MIMO testbed or prototype, with sometimes different objectives: while some of them aim to take measurements of the channel when multiple antennas are used at transmitter and receiver, some others are designed for showing the performance of specific algorithms, or even complex full systems.

It seems clear that the growing interest in MIMO systems has also been translated to increasing interest in real implementation and measurements. While theoretical aspects and studies are needed in order to optimally design codes, algorithms and methods to transmit information in the multiple input multiple output channel, it is undeniable that real measurements are of great importance. Channel simulations could be used to evaluate MIMO algorithms, but they are always based on some simplification and can never be considered as accurate as a real channel. Moreover, despite the great effort in knowing the MIMO channel features and the enormous number of channel measurements available at the literature, there are still some open issues to be clarified regarding some MIMO channel characteristics, such as the relationship between polarization and time delay in different scenarios, the signal depolarization depending on the indoor or outdoor type of scenario, etc.

Currently, there is also a great interest in studying different possibilities for the antenna elements and the array configuration to be placed in the user terminal or in the base station side. The antenna effect in the MIMO system and algorithms may be of key importance, and has not been thoroughly studied yet. Also possible reconfiguration of the antenna array, depending on the scenario characteristics, may be a good way to optimize the MIMO system.

After the initial step in this thesis of implementing a real-time smart antenna, it seems natural to go further and get introduced in the MIMO world. From the previous considerations, we observe that there are multiple aspects to be studied in a MIMO system that require a MIMO testbed or real implementation. Therefore, we have considered of paramount importance to design and implement a multi antenna test bed to be used in the study and analysis of several MIMO aspects. This chapter presents the design and implementation of a novel MIMO testbed, based on the same rapid prototyping premises used for the smart antenna presented in chapter 3.

Despite the great amount of MIMO implementations already shown in the literature, we have observed that they focus on a specific objective (channel measurement, algorithm testing...). However, the MIMO testbed we present here has a multi-purpose design, as novelty with respect to the existing ones, which allows us to use it for multiple studies. Next section details the testbed design and implementation, as well as novelties regarding previous testbeds. Section 4.3 describes a measurement campaign carried out to contribute to a better knowledge of the MIMO channel when multi-polarization is included, which was performed by using the novel multi-purpose MIMO testbed.
4.2 Design and implementation of a MIMO testbed

As a first part of the research work on prototyping for MIMO channel, a custom MIMO testbed was designed, implemented and evaluated. It is described in detail in the first part of this chapter, devoted to explain the innovative aspects that were addressed with this MIMO testbed, as well as several aspects dealing with implementation and possible impairments.

4.2.1 Design for multi-purpose scenarios and application

The idea of developing a MIMO testbed is not at all a novel one. Indeed, at the time of beginning this project on designing and implementing a MIMO testbed, several equipments for measuring MIMO channels have been presented and some channel measurements have been reported. Also some implementations of algorithms for MIMO systems could be found in the literature. Currently, MIMO systems are envisioned not only as a future possibility but as the promising technology to be used in the near future. As a result, many new MIMO implementations have been reported in the literature, in multiple forms (channel sounders, demostrators, testbeds, prototypes…), and with different purposes (to measure MIMO channels, to test different MIMO algorithms and schemes, to analyse different antenna configurations, to measure system throughput, to study integration of MIMO techniques with existing communication systems…). A thorough study of the state of the art in existing MIMO prototypes and their main characteristics is included in Chapter 2.

Despite the prolific work in MIMO testbeds, most solutions are designed for a specific objective, and not much flexibility is foreseen when defining the prototype. Conversely, the testbed we propose here is a system that was designed with a three-fold purpose: MIMO channel sounding, MIMO algorithm testing and Multi-element antenna evaluation for MIMO.

- MIMO channel measurement and characterization: the MIMO testbed was designed so multiple transmitters could send different signals to the channel, while multiple receivers could simultaneously capture the received signals after being modified by the channel effects (attenuation, fading, scattering, reflections…). Currently, several solutions have been presented as MIMO channel sounders, some of them even at a commercial level. One of the most world-accepted MIMO channel sounders and with an excellent measurement accuracy and broadband operation is the RUSK channel sounder with MIMO extension, by MEDAV [236]. Despite its undeniable good performance, an important drawback of this channel sounder is its very high cost. Moreover, the system has been designed in a non flexible way, thus not allowing other uses but channel characterization. There is no option to select the type of signals to be transmitted or the domain where signals are measured (this is fixed to the frequency domain, and results are afterwards converted to time domain). Conversely, the MIMO testbed proposed here allows for a flexible implementation, by means of using programmable DSPs in order to
let the researchers change the characteristics of the transmitted signals. Furthermore, off-the-shelf and inexpensive components were used to build the RF modules, so that a cheap implementation was achieved. Finally, our MIMO testbed uses different hardware chains for each of the transmitters and receivers, as opposed to the RUSK and other solutions, which are based on fast switches to change from one antenna to another. This way our system accomplishes the multiport transmission (and multiport acquisition) actually in the same time instant for all ports.

- **Testing of algorithms and signal processing schemes for MIMO**: conventionally, MIMO algorithms and TX-RX schemes are firstly analysed from a theoretical point of view, and then their performance is evaluated by means of simulation, by using more or less simplified MIMO channel models. However, the optimal evaluation would be obtained by measuring their performance in real conditions regarding channel and hardware. The main drawback that prevents algorithms designers from testing their methods in real environments is the high effort that programming these algorithms in real signal processing platforms involves. Having to consider real-time constraints and reduced memory of DSPs and FPGAs, as well as the required knowledge to program them, makes this solution quite time consuming in general. As an attractive solution, we propose here a system that allows the user to easily test algorithms and schemes for MIMO in multiple realistic channel conditions, by using a user-friendly window interface. The user will generate the data to be transmitted, and will obtained the received data after the signal has travelled through the MIMO channel.

- **Evaluation of multi-element antennas and their configurations in MIMO systems**: a third purpose of the proposed MIMO testbed is to be able to test different antenna arrays to be used in MIMO systems. It has not been until recently that the effect of antenna elements are being considered as an important part in MIMO systems, which have typically been predominantly studied from a “system and information theory point of view”. Due to this increasing interest, we think that including the possibility of reconfiguring the antenna elements to be used in the testbed is very attractive. Thus, several antenna elements were built for the system, also allowing for other possible elements to be connected. The antenna modules were implemented to permit different antenna polarizations and elements spacing.

To summarize, the MIMO testbed presented here has been designed with a novel flexible perspective, trying to cover multiple areas of current interest: understanding broadband MIMO radio channel, testing MIMO algorithms in real channel conditions, and evaluating the antenna in MIMO systems.

The detailed architecture and implementation of the proposed testbed is presented in next sections.
4.2.2 General architecture

The developed testbed is based on a modular architecture, which allows updating or improving each module in a separate fashion, as well as obtaining a flexible implementation. By making each module as independent as possible, the system can be improved in different phases (as has been the case for the RF modules, for example), according to the specific needs.

An important feature of the design is to follow the Software-Defined Radio concept [237], as the smart antenna system presented in the previous chapter. Therefore, modulation and demodulation is performed in digital signal processors, and the digital part of the system is pushed forward the RF part, in order to simplify reconfiguration. This way, a software update would allow an improvement or change in the system (as in changing the IF frequency, the type of modulation scheme, etc).

Figure 4-1 presents the general scheme of the system and the modular implementation.

Figure 4-1. Main scheme and modular architecture of the MIMO testbed.
The system consists of one transmitter node and one receiver node. Each node comprises three main modules: the signal processing module, the radio frequency module and the antenna module, which are represented in the bottom part of previous figure.

The top part of Figure 4-1 shows the general scheme of the system. It consists of one multi-antenna transmitter module, with \( N_{\text{TX}} \) transmitter channels and antenna elements, and one multi-antenna receiver module, with \( N_{\text{RX}} \) transmitter channels and antenna elements. The system can be configured for supporting up to 4 elements in each link end, that is, \( N_{\text{TX}} \leq 4 \) and \( N_{\text{RX}} \leq 4 \).

The transmitter (TX) node is controlled by a PC, which is connected to the MIMO transmitter modules through a standard PCI bus. The same applies to the receiver (RX) node, which is controlled by another PC. The software implementation of the control application offers the option of remotely controlling both nodes from another PC, as long as the control PCs are connected to a LAN network (as depicted in Figure 4-1). However, in order to facilitate the measurements in different locations, the option of local control of the nodes is also available, so they do not require to be continuous connected with a LAN cable.

The general operation of the system can be summarized as follows: the user application (typically, but not limited to, a Matlab program) generates the data to be transmitted (baseband signal), which depend on the purpose to be covered (channel sounding, algorithm testing, etc...). A certain data structure should be followed. The generated data is transmitted from the control PC to the TX node, where the signal is first stored, and after that transmitted. Since the storage capabilities in the DSPs are limited, and the transmission rate from DSPs to the analog modules is very high, in order to get a continuous transmission the same data are repeated until new data are stored in the DSPs memory. The signals are analog converted, and then up-converted from intermediate frequency (IF) to radio frequency (RF). After being transmitted by the antenna module, the signals undergo the channel propagation effects, and the resulting signals are received by the RX node. The converse operations are carried out in the RX modules: the received signal is down-converted to IF and then sampled and digitally converted. Finally, the digital data are baseband converted and stored in the DSPs memory, from where the information is sent to the user application in order to be off-line post processed. As can be noticed, this scheme includes some modules that operate in an offline basis (generation and preprocessing of signal, and the final postprocessing of signal), although some others work in real-time (DSP-FPGA processing: up- and down-conversion, filtering...). Therefore, the system is not meant to show a fully real-time operation. The off-line operation was preferred for one main reason: this way the real-time constraints do not apply here, thus highly simplifying the algorithm testing and avoiding the time-consuming design for optimizing memory and processing requirements. As a drawback, the system does not allow to test algorithms that are based on instantaneous feedback schemes, since TX and RX are not communicated in real time.
The transmitter and receiver modules are independent, so different clock sources are used at TX and RX for mixing the analog signals and also for sampling the signals. The benefits of using different clock references at TX and RX are an easier implementation and a more flexible structure when doing channel measurements, since TX and RX can be carried along different routes without the need of using very long cables to connect them. Moreover, an interesting aspect of this feature of the system is that it allows studying some realistic impairments such as the need for performing frequency tracking and correction and the effect of small frequency or phase drifts between TX and RX. This aspect may be of interest in real systems, where having the same clock reference at TX and RX cannot be assumed. On the other hand, using different clocks at TX and RX would certainly involve some error when measuring the MIMO channel, thus a lower performance should be accepted when using the MIMO testbed as channel sounder, as the price to be paid in order to have an easy to move and multi-purpose system.

In the following paragraphs the detailed features of each of the modules that constitute the MIMO system are described.

4.2.3 Modular description

4.2.3.1 Antenna modules

The antenna module has been designed so that different antenna array configurations can be evaluated, and some antenna features as element spacing and antenna polarization can be analysed from a MIMO perspective. Therefore, two antennas arrays have been designed to study how they influence in the MIMO channel capacity depending on the environment.

The first designed antenna array consisted of four $\lambda/4$ monopoles. The elements are situated on an aluminium ground plane of dimensions $15 \times 53$ cm$^2$, which corresponds to $1.22\lambda \times 4.24\lambda$ at the central operating frequency $f_0 = 2.45$ GHz, so that the effects for not having an ideal infinite ground plane can be neglected. Each monopole is a brass rod of 3mm diameter, and whose length was optimized to get optimum reflection coefficient at the operating frequency. Ideally, the theoretical monopole length should be $d_{th} = \lambda/4 = 3.06$ cm, but it is well known that in order to account for currents on the top of the rod the actual length should be slightly reduced. Thus, the proper length value was computed with the Tai equation [ref]. Finally, in order to account also for the other effect of the real implementation (such as narrower bottom part of the rod so it can be inserted in the SMA connector), the monopole length was adjusted by measuring the reflection coefficient within the operating frequency band, with a network analyser. The final implemented element is 2.8 cm long.

It is well known that correlation among received signals in a MIMO system have an effect in the available channel capacity and the suitability of processing algorithms to be used. Moreover, coupling among elements also affects the system performance, since it changes the
elements impedance and the antenna efficiency. Obviously, spacing among elements varies both
the antenna coupling and the signal correlation; therefore, it is of interest to evaluate its effect in
the algorithm and system performance. Aiming at allowing different array configurations,
several positions to locate the antenna elements can be selected, thanks to several holes to
connect the elements. The minimum and maximum possible distances are \(0.1\lambda\) and \(\lambda\),
respectively. An SMA connector is used to connect each monopole element to the
corresponding RF module. Figure 4-2 shows one of the implemented monopole element to the
corresponding antenna module. Two
similar monopoles arrays were built, in order to use them both at the receiver and the
transmitter.

![Figure 4-2. TX monopole array, to be used as one possible antenna module with variable spacing
among elements. RX array was very similar.](image)

Figure 4-3 shows the module of \(S_{11}\) for the 4 elements of the TX (left) and RX (right)
monopole arrays. As expected for this type of antennas, the impedance matching is quite good
for the operating frequency \((S_{11} \text{ lower than } 15 \text{ dB for } 2.4\text{GHz to } 2.5\text{GHz frequency band}).

![Figure 4-3. Reflection coefficient for monopole arrays at TX (left) and RX (right).](image)

Figure 4-4 shows the coupling coefficients between antenna pairs, for different antenna
spacing. As expected, higher spacing is translated into lower coupling coefficient. Conversely,
closely spaced elements are highly coupled. The lowest allowed spacing in the antenna module
\((d=0.1\lambda)\) means obtaining a quite high coupling of approximately \(|S_{21}| = -6\text{dB}\). For the results
presented in this chapter regarding channel measurements with different antenna combinations,
the case with highest coupling is \(d = 0.5\lambda\), with a coupling of approx. \(|S_{21}| = -13 \text{ dB}. According
to some results in the literature [238] the coupling effect in this case can be considered as negligible.

**Figure 4-4. Coupling level for different spacing between antenna pairs in the monopole array.**

The aforementioned omnidirectional elements were designed in order to allow studying different antenna spacing, coupling and correlation levels. However, other possibilities can be implemented so as to increase directivity or study other effects. One of them is polarization. Slant cross-polarized dipoles (±45º) have been designed and implemented to compare polarization and spatial diversity in MIMO channels with the monopole array. Figure 8 shows a photograph of the four crossed dipoles. The spacing between the element pairs is \( d = \frac{\lambda}{2} \).

Similarly to the case of the first antenna array, the slant dipoles were theoretically designed and afterwards their lengths were adjusted by measuring the reflection coefficients. Each dipole is fed by using a \( \frac{\lambda}{4} \) balun structure, as typically done for this type of antennas. Each pair of crossed dipoles is mounted on a methacrylate structure, designed to hold the elements at a fixed location and with a fixed relative spacing among the two antenna pairs. The material was selected to be as ideal as possible (low dielectric constant and losses), so its effect on the MIMO channel can be neglected. Figure 4-5 shows a photograph of the TX module with two slant antennas (two input ports for each antenna, that is, 4 input ports for the module).

Figure 4-6 and Figure 4-7 present the measured S matrix for each 2-port 2-dipole antenna, at TX and RX respectively. As expected, we obtain a good reflection coefficient for all cases (approx. 15 dB or lower for the operation band, in grey). The coupling coefficient differs from some antenna pairs to others. This can be blamed to implementation issues, due to the difficulty of soldering the dipole branches so that the disturbance to the counterpart dipole is low. Nevertheless, the measured coupling coefficient is lower than -15 dB for the worst case (antenna 2), being much lower (close to -35 dB) for the best case (antenna 3). Moreover, an extra decoupling due to polarization orthogonallity can be assumed. We see that coupling for
this antenna module is much lower than for the monopole array with conventional spacing of \( \lambda/2 \). However, coupling between co-polarized elements is expected to be quite high also for this second antenna module. Still, it is low enough as not to require a matching matrix to mitigate its effect.

![Figure 4-5. Slant dual-polarized antennas based on crossed dipoles.](image)

![Figure 4-6. S parameters of TX antennas.](image)

a) Antenna 1 (ports 1.1 (+45°) and 1.2 (-45°))  
b) Antenna 2 (ports 2.1 (+45°) and 2.2 (-45°))
The interface between the antenna module and the RF-IF module is an SMA connector, which is quite commonly used for RF applications. Therefore, other antenna modules can be designed and easily connected, should the system require it.

### 4.2.3.2 RF-IF modules

As previously mentioned, the MIMO prototype was designed on a modular basis. This way, each module could be updated or improved independently. That was the case with the RF-IF module.

A first version of the RF modules was implemented in a first stage, as a proof of concept implementation. Cheap components, as surface mount filters and amplifiers for the operation frequency band were used, and several plug-in submodules were designed and implemented. A detailed description of this first radiofrequency stage can be consulted in [239].

In order to improve the first RF-IF module, a second version was designed and built. The aim was to increase reliability, to improve stability and also to obtain a higher input power (high amplification) and lower noise level at RX (with low noise amplifiers). The details of this second version of RF module are given below.

The RF module consists of four RF-IF chains, built with plug-in and commercial coaxial components. A heterodyne scheme (typically used for these systems) was chosen for both the transmitter and the receiver. The selection of amplifiers and power levels has been carried out taking into account the required power levels to perform outdoor/indoor measurements. The RF-IF modules were mounted in a metal box with standard dimensions (2U), so it can be integrated in a 19” rack.

For the transmitter module, the general scheme is shown in Figure 4-8, while Figure 4-9 shows photographs of the final implemented module.
The signals to be transmitted are fed to the RF module modulated over intermediate frequency (IF) from the processing module. Up-conversion is accomplished by 4 mixers, which use the signal from the same local oscillator (LO) to up-convert the signal from IF to RF. IF frequency was chosen as $f_{IF} = 40$ MHz, as explained in next section. The LO signal is a very stable sinewave obtained from a commercial signal generator, which is split into 4 branches in order to be get the same LO signal for the 4 paths. Its frequency is chosen to obtain the carrier frequency of 2.45 GHz, which is the central frequency of standardized bandwidth for 802.11 WLAN communications. The upper frequency for the LO was chosen, that is, $f_{LO} = 2.49$ GHz, since the BP filters were able to properly filter at that frequency any possible LO leak due to non ideal LO-RF isolation, while $f = 2.41$ GHz is not filtered (it is within the bandpass of the used filter). A power amplifier was included in each RF chain, so that up to approx. 25 dBm per antenna can be transmitted, fair enough for an indoor/outdoor system.

At the receiver, the RF signal is amplified with the aim of increasing power level received as input of the processing module, since the DSPs sensibility is around -40 dBm. The
overall scheme for the RF module at RX is depicted in Figure 4-10. A chain of 4 amplifiers was used to get a high amplification of signal. In order to reduce the final noise at the receiver, a low noise amplifier was used a first component of each chain. In the current implementation, the sensibility of the receiver module at the antenna connector is -80 dBm. Similarly to the transmitter side, a commercial signal generator produces the signal for the local oscillator at the receiver. This sinewave is sent to four paths through a 1 to 4 divider. Since the local oscillator is a different different at transmitter and receiver, frequency tracking in reception is required, and a residual effect due to the remaining frequency error must be accepted. Despite this disadvantage, the system design including two different oscillators instead of one was preferred, since this scheme simplifies the independent movement of TX and RX along different measurement routes and it allows including more realistic effects in the algorithms. Figure 4-11 shows photographs of the final implemented module.

Figure 4-10. Scheme of RF module at RX.

Figure 4-11. Photographs of the RF module at RX.
A more detailed description of the design steps in the RF module is out of the scope of this thesis, as well as the models of selected components or their features. Interested readers can refer to [239],[240] for more details.

4.2.3.3 Signal processing modules

As stated in previous sections, the test bed has been designed on the basis of the Software Define Radio (SDR) approach, that is, a reconfigurable and reprogrammable Software Radio platform is used to accomplish baseband up- and down-conversion, so the system can be reconfigured regarding selected modulation and coding, bandwidth and data rate. That means that the signal processing modules may change from one application to another, but thanks to the SDR-based implementation the required updates are easily carried out. In the MIMO testbed proposed here, a basic transmitter-receiver scheme is considered, since we aim to obtain an architecture as general as possible. The SDR technology uses the commercial standard VME bus structure to host the processing boards. The four digital transmit channels are performed in a Pentek 4292 Quad DSP (with four TMS320C6203 DSPs) board and two Pentek 6229 digital upconverter daughter boards. At the receiver, the four digital channels are implemented in another Pentek 4292 Quad DSP (also with four TMS320C6203 DSPs) board and two 6235 digital downconverter daughter boards. The upconverter (downconverter) boards are mainly based on an FPGA and two digital-to-analog (analog-to-digital) converters, so that two upconverter (downconverter) channels are achieved with each board. As can be observed, the same DSP and FPGA models are used as in the smart antenna implementation presented in chapter 3, thus taking advantage of previously gained experience. However, we may note that the operation mode here is completely different from that one proposed there: while for the smart antenna a fully real-time operation was aimed, for the test bed presented here the processing module works in a mixed off-line and on-line (or real time) operation mode. The source signal is previously generated and pre-processed off-line. Afterwards, it is synchronously sent to the SDR platform and transmitted through the MIMO channel and then received and stored in memory (all this process obviously in a real-time basis). Finally, the received signal is sent to a PC, where an offline processing can be realized.

The main scheme for the software radio module at transmitter is shown in Figure 4-12. The host PC communicates with the digital signal processing boards by means of a dedicated connection. A conventional PCI connection is used as interface between the PC and the DSP QUAD, thus allowing to use any conventional server or PC as control and preprocessing PC. Signals from PC are transferred to the DSP board by means of a TCP/IP connection, and the preprocessed data are stored in SDRAM blocks of 8 Mbytes, one block associated to each DSP.

Once all the data to be sent are stored in the SDRAM blocks, the operation is quite similar to the one presented in chapter 3, since the same hardware is used. A short explanation is
given here, so the main ideas are reviewed. More details on the specific implementation can be consulted in chapter 3.

The four transmitter channels are synchronized in order to start the transmission of signals simultaneously from the four DSPs. Specific libraries from the QUAD manufacturer are used to synchronize the signal transmission, which is basically based on a ring communication with one master DSP and the rest of the DSPs being slaves. A 4-kB FIFO bus is used between each DSP and its associated upconverter channel to allow for real-time operation and not losing samples. In the upconverter module, I and Q signals are fed to an interpolation filter and are digitally upconverted to IF. Then, a digital to analog converter (DAC) carried out the conversion to send analog signals to corresponding input at the RF module. 12-bit converters are used. A synchronization trigger is sent to the buses and the digital upconverters in order to get a fully synchronized upconverted signal, which is extremely important in a MIMO system (as it also was in the smart antenna implementation) in order to keep phase references between transmitters. Also the same clock signal is used to trigger the four DACs at the transmitter, so the four channels are analog-converter at the same time. The clock signal is obtained from a very stable crystal oscillator and some multipliers at the transmitter. Input sampling frequency for DACs is reprogrammable up to 200 MHz. The selected sampling frequency at the DACs for the user interface deployed here (see section 4.2.3.4) was $f_{s,TX} = 200$ MHz, although the input data rate to the upconverter is lower, e.g. $f_{IN,TX} = 6.25$ MHz for channel measurement application (subsequent interpolation is performed).

Figure 4-12. Scheme of signal processing module at TX, implemented over DSPs and FPGAs.
At the receiver, conversely operations are performed, as shown in Figure 4-13. The signals received by antenna and RF modules are fed at the IF input of the software radio platform, and then sampled by ADCs. The used platform supports 12 bits in ADC and hence a good dynamic range. The sampling rate at the ADCs is reconfigurable up to 105 Msamples/s. The obtained digital IF signals are converted to baseband by the digital downconverters, which obtain two branches each (I and Q). Finally, the baseband signals are filtered with a decimation filter in order to remove undesired copies of signal after downconversion. The decimation rate can be selected by the user. Finally, the signals are fed to the FIFO buses and read by the DSPs, which store the information into their 8MB local SDRAM. Also here a trigger signal is sent to synchronize the downconversion process and the FIFO buses, and a master-slave scheme is carried out so that the four DSPs receive the information synchronously. A single clock is used for all the ADCs, being obtained from a crystal oscillator (a different one from that used at TX). The sampling frequency at ADC is selected as $f_{S,RX} = 100$ MHz. The decimation value is selectable. For example, for the channel sounder application, the decimation value was chosen as $N_{DEC} = 16$, so the signal received at the DSPs had a data rate of $f_{IN,RX} = 6.25$ MHz.

**Figure 4-13. Scheme of signal processing module at RX, implemented over DSPs and FPGAs.**

More details on the specific implemented software to control the DSPs and the up- and down-converters, as well as the parameters to configure the system are not included here, since this information was not considered relevant for explaining the contributions of this chapter. They can be consulted in [239].
The communication between the FPGAs-DSPs modules and the PCs is carried out by means of the specific libraries for the selected SDR platforms, which allow the communication with Matlab via TCP/IP protocol. The software development of the testbed is based on Pentek Swiftnet environment, which supports remote access through TCP/IP connection. This is possible thanks to stream API of Swiftnet, which provides a means for moving streams of data between host and target. In our case, Matlab is the client application which is running in the host, hence a flexible performance is possible.

4.2.3.4 User interface and post-processing

In order to simplify the user interface and to make it easier the use of the implemented testbed, especially for algorithm testing, a Matlab-based application was developed, with a windows-like (also called “GUI” in Matlab environment) format. This interface is presented at the user’s PC. The communication between the testbed (control PCs, or FPGAs-DSPs modules) and the PCs is carried out by means of the specific libraries for the selected SDR platforms, which allow the communication with Matlab via TCP/IP protocol, as represented in Figure 4-1. For this purpose, a streaming solution and its API functions for developers, based on the Swiftnet solution for communication, from Pentek, was used. SwiftNet is a powerful communications protocol that links a workstation and software development tools to the VMEbus DSP processors. This solution treats DSP processors as network devices and supports all possible combinations of platforms, tools and boards thereby improving the overall development environment from Swiftnet to be used with Pentek SDR. As a result, a transparent communication between the user’s PC and the testbed is achieved. In our case, Matlab is the client application which is running in the host (the user’s PC).

Two main user applications were developed: MIMO channel sounder application and MIMO algorithm testing application. They are presented below.

**MIMO CHANNEL SOUNDING APPLICATION**

The first one was implemented in order to make the channel measurements in a more automatic way, so that the planned measurement campaigns (see next section) were easy to carry out. Figure 4-14 shows the main window of this application, to be run in a Matlab environment.
Figure 4-14. Main window of the client application in Matlab, for channel sounding.

The main window is divided into two sets of parameters: the transmitter and the receiver parameters. The number of antennas is a parameter to be chosen by user, between 1 to 4 elements. The spacing between elements is included just as extra information to be saved with the data, since it does not affect the measurement procedure.

The measurement procedure is summarized as follows: once the file where the signals to be sent is selected, the data must be loaded in the DSPs, by clicking the “load” button at TX. To transmit the signal, the “transmit” button is clicked. If both operations are successfully performed, the corresponding ticks should appear. After this, the signals to be transmitted are sent through the TX modules in a periodical way (once the 8MB are sent, the transmission begins again), after the user selects the “transmit” button again (the signal to be transmitted is updated with the new information in the edit box for file of transmit signal) or when the user clicks the “halt” button to stop the TX DSPs.

To receive the signals, the DSPs are loaded with the receiver program (by clicking the “load” button), and the file where the resulting signals will be saved is specified. Then the system is ready to acquire the 8 MB of received data. We may note here that the reception is not synchronized with the transmission, and hence the received data will not begin at the same position as the transmitted data. Nevertheless, since the data are periodically sent, no information is lost. In order to received and save the data, the button “receive” must be clicked. If a new 8MB block of data is to be received, the user just needs to change the file name and
click “receive” again. Different receptions of data can be performed until the “halt” button is pressed to stop the RX modules.

From the explanations above, we see that the channel sounding application is very simple and easy to understand for the users, as required by the initial specifications that we had set to the system.

**OPTIONAL POSTPROCESSING FOR CHANNEL SOUNDER**

The user may prefer to implement his or her own pre- and post-processing modules to estimate the H parameters for channel matrices. However, in order to simplify the channel sounding tasks, some modules have already been implemented and can optionally be selected by the user when the default transmitted signals have been selected. Specifically, the channel sounder application allows the user to select the synchronization of the received signal with the transmitted blocks, and also to perform frequency tracking to mitigate errors due to possible frequency drift or differences between frequency at TX and RX. Moreover, the option of estimating the H matrices from the received signals and the used transmitted signals is presented to the user. Figure 4-15 shows the window to be presented to the user when “Estimate H matrix” is clicked. The option of plotting the estimated H matrix as a function of instant time t and time delay τ is given. Also a report can be generated, which comprises data from measurement (number of TX and RX antennas, spacing between elements and estimated H matrix). Finally, once the H matrix is estimated, the theoretical capacity of the MIMO system can be computed. The window dialog gives the option to select “Compute capacity”, which graphically presents the capacity assuming no channel state information at the transmitter (see eq. (4-2) below).

![Figure 4-15. Window dialog for estimation of H matrix (Channel Sounder application).](image-url)

The estimation of channel matrix is performed by comparing the received signal with known bits (reference bits). Four pseudonoise (PN) codes of 256 bits in length with appropriate are used to be sent at the transmit antennas, one for each antenna. The codes are chosen so that they show good correlation properties (very low correlation between different PN codes, and
very low autocorrelation for \( \tau \neq 0 \), so the 4 transmit antennas can be differentiated at the receiver, and a good synchronization stage can be carried out at the receiver.

Before sending the PN codes, they are oversampled by a factor of 4, in order to simplify synchronization at the receiver. Thus, 1024 samples are sent for each PN code. The code is repeated 2048 times, in order to get a full bunch of 8MB data, being each sample of 4 bytes (2-byte I branch, 2-byte Q branch). The final 8MB packet of data is stored in a .mat file, so the user can choose to use it if he/she wants to use the provided method to estimate H matrices. We may note that when the provided preprocessing data and postprocessing option are chosen, and taking into account the selected sampling frequencies at TX (\( f_{\text{IN,TX}} = 6.25 \text{ MHz} \)) and RX (\( f_{\text{IN,RX}} = 6.25 \text{ MHz} \)) as explained above, then the used bit rate is 1,5625 Mbps. This data rate limits the bandwidth of channel matrix to be estimated to 1,5625 MHz.

In the off-line processing at reception, the received signal is first synchronized to the known PN codes (in order to find the time instant for the beginning of the PN code) and then each received signal (for the 4 RX antennas) is correlated with each of the known PN codes. The synchronism stage is divided into two parts (similarly to the performed process for synchronization in the smart antenna, see chapter 3). In the first one, a coarse synchronism is carried out by correlating the received signal for each antenna with the transmitted PN code. For all the blocks, the maximum value is calculated, so the \( t_o \) instant is obtained. Then, in the fine synchronism the correlation for each corresponding 256 bits is calculated for the previous and following instants (early and late samples), Thus, a comparison to choose the optimum instant is implemented.

Finally, an instantaneous phase correction block has been designed to mitigate the frequency error effect, by estimating the average frequency drift and correcting the corresponding phase shift at each time instant. The frequency error is estimated as an average value for the overall bunch of 8MB received data. Thus, we assume that instantaneous changes in IF frequency are negligible compared with the frequency error due to differences in IF frequency caused by using different local oscillators at TX and RX. This assumption was verified by measurements with the implemented testbed.

**ALGORITHM TESTING**

The second implemented application is the MIMO algorithm tester, whose main window is shown in Figure 4-16. Its structure is quite similar to the one developed for the channel sounder, since the principles are the same. However, for this application the option of selecting different sampling rates (and therefore different signal bandwidths) is included. Three possible bit rates, or better say sampling frequencies are offered: fast, medium (default) and slow. The aim here is to let the user choose the data rate he or she wants to consider for the
algorithm to be tested, and thus how the possible implementation impairments may affect it depending on the bandwidth and data rate.

![Figure 4-16. Main window of the client application in Matlab, for MIMO algorithms testing.](image)

We may observe also that 4 different files are included for each TX and RX antenna, so it can be easier to change just one of the antenna modules. Another difference with the channel sounder application is that the option of including channel estimation or frequency correction is not included here: the user should include it in his/her algorithm. However, in order to consider these postprocessing processes, a structure of frame where a reference signal is sent at the beginning of each block is recommended. This way, the estimation of frequency drift and the estimation of channel matrix is simplified. As an example, a $2 \times 1$ Alamouti scheme was tested with the MIMO algorithms testing application, following the structure shown in Figure 4-17. Figure 4-18 depicts the received constellation for a BPSK signal before and after Alamouti scheme at receiver. It can be seen that a residual frequency error still remains. More details on this test were presented in the Master Thesis [241], where an example of using the MIMO testbed for testing MIMO algorithms can be consulted.
Figure 4-17. Example of used frame structure for algorithm testing.

Figure 4-18. Constellation of received signal before (upper figure) and after (lower figure) Alamouti demodulation.

4.2.3.5 Main features of final prototype

All the modules that constitute the MIMO testbed were integrated in two main subsystems: the TX and the RX. The TX modules were mounted on a large rack, since this subsystem is assumed to be located at a static position, and only moved when different scenarios are measured (TX at indoor or outdoor, etc). The RX modules are also integrated on a 19” rack, but this one is a quite small one, only containing the SDR part. All the other modules and the rack are mounted on a trolley, so the RX can be carried along different routes and multiple measurement points can be easily taken. General tests were accomplished to validate its operation: measurement of transmitted signals with a signal analyzer (several parameters were evaluated, as power levels and signal bandwidth), evaluation of RX parameters when an ideal signal from a signal generator is used as input, and finally overall system operation when both transmitter and receiver are used, and the channel is an ideal one (fully uncorrelated 4 subchannels: cables and attenuators are used to connect TX and RX).

The final system was named UMAT, which stands for UPM MultiAntenna Testbed. Figure 4-19 shows a photograph of the final TX subsystem in the cabinet-rack, and Figure 4-20 shows one of the RX subsystem mounted on the trolley.
The main characteristics of the UMAT system have been stated along the above sections. They are collected in next table, in order to summarize them and make it easier for the reader to have an overall view of the testbed.
### MAIN FEATURES of UMAT (UPM MultiAntenna Testbed)

<table>
<thead>
<tr>
<th>Feature</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of antennas at TX (NTX)</td>
<td>4</td>
</tr>
<tr>
<td>Number of antennas at RX (NRX)</td>
<td>4</td>
</tr>
<tr>
<td>Type of antennas</td>
<td>· Option 1: Omnidirectional single-polarized monopoles.</td>
</tr>
<tr>
<td></td>
<td>· Option 2: Omnidirectional dual-polarized slanted dipoles.</td>
</tr>
<tr>
<td>Maximum TX power</td>
<td>25 dBm per TX antenna</td>
</tr>
<tr>
<td>Operating frequency</td>
<td>( f_0 = 2.45 \text{ GHz} )</td>
</tr>
<tr>
<td>Sampling frequency at TX (Maximum)</td>
<td>( f_{s,TX,\text{max}} = 25 \text{ Msp} )</td>
</tr>
<tr>
<td></td>
<td>Used for channel sounding: ( f_{s,TX} = 6.25 \text{ Msp} )</td>
</tr>
<tr>
<td></td>
<td>Minimum: ( f_{s,TX,\text{min}} = 0.781 \text{ Msp} )</td>
</tr>
<tr>
<td>Sampling frequency at RX (Maximum)</td>
<td>( f_{s,RX,\text{max}} = 25 \text{ Msp} )</td>
</tr>
<tr>
<td></td>
<td>Used for channel sounding: ( f_{s,RX} = 6.25 \text{ Msp} )</td>
</tr>
<tr>
<td></td>
<td>Minimum: ( f_{s,RX,\text{min}} = 0.781 \text{ Msp} )</td>
</tr>
<tr>
<td>Supported signal bandwidth (at baseband)</td>
<td>Up to ( B_{\text{max}} = 12.5 \text{ MHz} ).</td>
</tr>
<tr>
<td>Maximum stored data</td>
<td>8MB per antenna</td>
</tr>
<tr>
<td>Operation mode</td>
<td>Off-line pre and postprocessing + real-time up and down conversion and general filtering.</td>
</tr>
<tr>
<td>User interface</td>
<td>Matlab “GUI” user-friendly windows</td>
</tr>
<tr>
<td>Developed Applications</td>
<td>· Channel Sounder</td>
</tr>
<tr>
<td></td>
<td>· Algorithm testing</td>
</tr>
<tr>
<td>Interface protocol with testbed</td>
<td>TCP/IP + Swiftnet</td>
</tr>
<tr>
<td>Advantages for the user</td>
<td>· Easy testing of algorithms under real channel conditions.</td>
</tr>
<tr>
<td></td>
<td>· Matlab interface, easy to use.</td>
</tr>
<tr>
<td></td>
<td>· Offline processing, avoiding real-time constrains and enhancing time to evaluate MIMO schemes.</td>
</tr>
<tr>
<td>Advantages and contributions</td>
<td>· Designed and developed for multipurpose operation (channel sounder, algorithm testing, antenna evaluation).</td>
</tr>
<tr>
<td></td>
<td>· Both single-polarization and dual polarization were considered.</td>
</tr>
<tr>
<td>Contributions to current testbed</td>
<td>· Realistic implementation impairments can be studied and their effect evaluated in MIMO algorithms (frequency offset, different gains at each channel...).</td>
</tr>
<tr>
<td></td>
<td>· Realistic antenna modules can be tested to evaluate their effect in MIMO algorithms.</td>
</tr>
</tbody>
</table>

Table 4-I. Summary of UMAT main features.
4.3 MIMO channel measurements: Indoor and Outdoor-to-Indoor channel characterization including polarization effect

4.3.1 Motivation and progress beyond the state of the art

One of the applications of the presented MIMO testbed is the possibility to measure the radio MIMO channel, thus allowing the evaluation of different characteristics of the channel. As part of the research work presented in this chapter, several results are shown regarding radio channel characterization.

A special emphasis is put on the performance comparison when single and dual-polarized antennas are used in the MIMO system.

4.3.2 Measurement set-up

A measurement campaign was conducted at the Faculty of Telecommunication Engineering, ETSIT-UPM. With the objective of evaluating different scenarios and MIMO channels, multiple locations were considered for both the transmitter and the receiver modules. Moreover, two antenna configurations were sequentially used, in order to compare the system performance when polarization diversity is added to the conventional spatial diversity offered by MIMO systems.

Regarding the receiver module, office and also corridor scenarios have been considered. The former is usually considered a richer multipath scenario, thus offering higher spatial diversity gains in general. However, it may be noticed that in the presented case the corridor scenario includes sections with line of sight (LOS) between transmitter and receiver, thus leading to a higher received power (and higher signal to noise ratio, SNR), as it will be seen later. Figure 4-21 shows the positions for the receiver, labelled as route 1-2 for the office scenario and 3-4 for the corridor.

Regarding the transmitter module, three different locations were considered: two indoor (A and B in Figure 4-21) and one outdoor (C in Figure 4-21) location. The first TX indoor location (A) is used when office-like indoor positions are considered for the RX. In this case, the transmitter is located in a room with several laboratory equipments and computers, as well as office furniture. Therefore, a scenario with high multipath richness may be expected. The two other transmitter locations (B and C) are used in combination with the corridor route 3-4 for the receiver. The receiver was carried along the same route for the two TX locations, in order to obtain a fair comparison with the two options. For position B, the transmitter is situated at the end of a corridor, pointing at the other end. For position C, the transmitter was located on the rooftop of a nearby building, roughly pointing at the corridor route for RX, and at a one-floor higher position than for the indoor location. We may note that, with the proposed TX locations, two types of scenarios are covered: an indoor one and an outdoor-to-indoor one.
For each combination of transmitter and receiver locations, two types of antennas have been used: firstly the monopole arrays (Figure 4-2) were utilized at both the receiver and the transmitter, in order to evaluate the system performance when only vertical polarization is used. Afterwards, dual-polarized antennas were used (Figure 4-5), so polarization diversity is included in the system, to the cost of reducing spatial diversity (since the dual-polarized dipoles are co-located).

As a result, the complete measurement campaign includes 6 sets of measurements or “cases”. Table 4-II summarizes the measurement setup for the 6 measured scenarios. Figure 4-21 depicts the TX locations and RX measured routes on the floor map of the building. Figure 13 shows the antennas viewpoints from the indoor and outdoor locations in the corridor scenarios (cases 3 to 6). Figure 14 shows the office were the transmitter is located for cases 1 and 2 (the receiver was moved along an office in non-line of sight, NLOS, situation).

<table>
<thead>
<tr>
<th>Case (scenario)</th>
<th>Case 1</th>
<th>Case 2</th>
<th>Case 3</th>
<th>Case 4</th>
<th>Case 5</th>
<th>Case 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>RX Route</td>
<td>Office route (1Æ2)</td>
<td>Corridor route (3Æ4)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>TX location</td>
<td>Indoor (A)</td>
<td>Indoor (B)</td>
<td>Outdoor (C)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>polarization</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>TX Power (for</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>each branch) [dBm]</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of</td>
<td>7</td>
<td>7</td>
<td>17</td>
<td>17</td>
<td>17</td>
<td>17</td>
</tr>
<tr>
<td>measured points</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 4-II. Summary of measurement setup for each of the six scenarios under analysis
Figure 4-21. Floormap of measured scenario. TX locations are shown in red (letters), RX routes are depicted in blue (numbers).

Figure 4-22. Tx viewpoint for indoor (left) and outdoor (right) location for different antenna arrays.
4.3.3 Approaches for capacity computation considering channel normalization

The collected data were used to estimate the channel coefficients by correlating the received signal with a known PN code, as explained in previous section. Although the MIMO testbed has the possibility to operate in a broadband basis, narrowband measurements were considered in this analysis, for simplicity reasons. This type of measurements is fair enough to evaluate the theoretical narrowband capacity with different Tx and Rx setups.

The main mathematical theory and equations regarding MIMO capacity computation were already presented in chapter 2. However, we include them again here, in order to extend the study by considering also the power computation and different ways to normalize the channel matrix. This analysis is of great interest, since depending on the type of considered normalization a different assumption will hold in the system, as we will see later.

Let us assume a narrowband system with $N_{TX}$ antennas at the transmitter and $N_{RX}$ antennas at the receiver. Telatar [7] derived the general expression for the capacity in this system, which is given by

$$C = \max_{Q} \log_2 \left| \frac{1}{N_{RX}} \mathbf{I}_{N_{RX}} + \frac{\rho}{N_{TX}} \mathbf{Q} \mathbf{H}^H \mathbf{H} \mathbf{Q} \mathbf{H}^H \right|$$

(4-1)

where $\mathbf{Q}$ is the covariance matrix of transmitted signals, such that $\text{Tr} \{ \mathbf{Q} \} \leq N_{TX}$ to account for power constraint, $\rho$ is the signal to noise ratio at the receiver, $(\cdot)^H$ denotes Hermitian and $|\mathbf{A}|$ is the determinant of matrix $\mathbf{A}$. Two cases were considered in this analysis: no channel state information (CSI) at transmitter and total CSI at transmitter. In the first case, the power allocation strategy is assumed to be uniform, so that the channel capacity expression may be simplified to

$$C_{UP} = \log_2 \left| \frac{1}{N_{RX}} \mathbf{I}_{N_{RX}} + \frac{\rho}{N_{TX}} \mathbf{H} \mathbf{H}^H \right|$$

(4-2)
When total CSI at transmitter is considered, the optimum waterfilling scheme is assumed to allocate power, so the singular value decomposition (SVD) of $H$ is realized, and the capacity is computed as

$$C_{WP} = \sum_{i=1}^{K} \ln(\mu \lambda_i^+)$$  \hspace{1cm} (4-3)

where $(\cdot)^+$ denotes taking only those terms which are positive, and $\lambda_i$ is the $i^{th}$ (out of $k$) non-zero eigenvalue of the correlation channel matrix $R=HH^H$. The parameter $\mu$ is chosen to satisfy the power constraint

$$\rho = \sum_{i=1}^{K} \left( \mu - \frac{1}{\lambda_i} \right)^+$$  \hspace{1cm} (4-4)

Conventionally, in order to remove the path loss effect and study the diversity characteristics of the MIMO propagation channel, the channel matrix $H$ is usually normalized to obtain a fixed local signal to noise ratio for each measured point, method that was firstly presented in [242]. The use of this normalization is equivalent to considering a perfect power control in the system. This is interesting to characterize the multipath richness and diversity offered by the propagation environment, but it does not take into account the path loss, shadow fading and penetration losses. The normalized channel matrix for each instant $H(n)$ is then computed as

$$H(n)_{norm} = \sqrt{N_{TX} \cdot N_{RX}} \frac{H(n)}{\|H(n)\|_{Frob}} = \sqrt{N_{TX} \cdot N_{RX}} \frac{H(n)}{\sum_{i,j} h_{ij} \cdot h_{ij}^*}$$  \hspace{1cm} (4-5)

Also path loss analysis is of interest in these systems, especially from the deployment and operator point of view, since this study gives an idea of coverage and required TX power to guarantee a certain power level and thus a certain service quality. The Frobenius norm can be used for this purpose, so that the average power is obtained as

$$P_{avg} = \frac{\|H\|_{Frob}^2}{N_{TX} \cdot N_{RX}} = \frac{\sum_{i,j} h_{ij} \cdot h_{ij}^*}{N_{TX} \cdot N_{RX}}$$  \hspace{1cm} (4-6)

We observe that eq. (4-6) represents the conventional normalization that is used to compute capacity (as in eq. (4-5)).

Conversely to typical capacity analysis, we have considered that computing capacity by also including the effect of path loss and shadow fading is of great interest, especially when comparing channel measurements in different environments. Otherwise, the analysis would neglect some effects of importance, as the LOS-NLOS distribution and capacity variance due to it. In order to include this effect, a new normalization is considered for the MIMO channel: $H$
matrices are not instantaneously normalized, but they are normalized to a single value for each set of measurements, so that the average received SNR is the same for all sets. This way, the comparison between different sets does not depend on transmitted power or path loss differences (as when comparing indoor and outdoor Tx location), but on the available diversity. The new normalized H matrix is now computed as:

\[
H(n)_{\text{norm,new}} = \frac{H(n)}{\sqrt{\frac{1}{N} \sum_{n=1}^{N} P_{\text{avr}}(n)}}
\]  

(4-7)

where N is the number of collected points during the measurement in a certain scenario.

The obtained power and capacity results are shown below, and some results are drawn from them.

4.3.4 Power computation and coverage maps

In order to analyse the path loss differences for indoor and outdoor-to-indoor scenarios, the average received power for the cases 3 to 6 has been computed, and the results are depicted in a similar way to coverage maps in Figure 4-24.

Twenty measured values of H were used, and a linear interpolation is done to represent the power for the whole route. From Figure 4-24 we observe that the power is more evenly distributed for the outdoor-to-indoor scenarios (cases 5 and 6), which is reasonable since not direct LOS situations are found in this type of environment (all the measured positions are in NLOS). For the indoor transmitter location (cases 3 and 4) the remarkable difference in received power between the LOS and NLOS routes in the corridor is clear, as expected. Thus, the coverage is improved for routes in direct line of sight with the transmitter module. It is also interesting to note that, for the indoor transmitter location, the received power level suffers higher variance in the NLOS routes, showing a higher fading effect. From a deployment operator point of view, it may be of interest an even distribution of power as the one obtained with Tx outdoor locations. Nevertheless, it may also be considered that a higher transmitted power is required in these scenarios: while a transmitted power of 2 dBm is used for the indoor locations, 23 dB more are necessary to obtain similar received power levels when an outdoor Tx site is used.
Figure 4-24. Received power computed from the estimated H matrices (squared Frobenius Norm).
4.3.5 Capacity results for indoor and outdoor-to-indoor scenarios

By using the conventional normalization of fixed SNR at the receiver (eq. (4-5)), the achievable capacity was obtained for non CSI at the transmitter (eq. (4-2)), and the results are shown in Figure 4-25. An SNR of 20 dB was assumed. Similarly to the maps of received power, a smooth representation in the figures is obtained by interpolating some extra values to the measured ones. It is interesting to note that both NLOS and LOS situations are included in the measurements. In general, for NLOS cases a lower spatial correlation may be assumed, which involves higher available spatial diversity and thus higher capacity for a fixed SNR at the receiver.

Firstly, the results for indoor scenario for the two antenna cases are compared (Figure 4-25, a) and b)). From these figures, we clearly observe that a higher capacity is achieved when the dual-polarized arrays are used for the measured corridor scenario. We may notice that, due to the normalization performed on the $H$ matrix, the path loss effect along the route is not included. Figure 4-25, c) and d) compare the achievable capacity for outdoor Tx location for the two considered antenna arrays. As in the indoor case, the dual-polarized arrays offer a higher capacity than the single-polarized ones. As a result, we conclude that the dual-polarized system offers better performance for the measured corridor scenario when optimum power control may be assumed in the system. When comparing results from different Tx locations, no remarkable differences can be observed in average.

In order to gain a better insight into the capacity results, we have computed the cumulative distribution function (CDF) of the capacity for the four cases under study, by considering the “fixed” normalization (“new” normalized $H$ matrix), as stated in eq. (4-7). Assuming a mean SNR = 20 dB, Figure 4-26 shows the obtained CDF of capacity for the 4 cases.

We should consider that only 17 measured points are used to compute the CDF, so the distribution curves are not smooth. From Figure 17 we may observe that the effect of LOS and NLOS situations is quite significant: the positions where there is LOS imply a higher received power, and thus a higher capacity. This behavior is more remarkable for the indoor Tx position, as expected. Again, in the figure we observe that the dipoles offer a higher capacity for both indoor and indoor-outdoor scenarios, while a slightly better performance is attained in the indoor scenario. We may note, however, that by normalizing to equal mean received SNR the effect of different antenna gain is not included.
Since the same environment was considered for the three antenna configurations, the H matrices were normalized to the same reference value: the mean Frobenius norm for the set of
measurements obtained with monopoles spaced \( d=\lambda \). Therefore, the comparison takes into account the gain of each antenna configuration, as opposed to the previous results. We observe that with this consideration, the monopoles achieve a higher diversity, since they offer a higher gain in the horizontal plane (where most of the rays may be assumed to impinge) that the dipoles.

When the two power allocation schemes are compared, we observe that an increase of approximately 1 b/s/Hz may be achieved with the waterfilling scheme thanks to the exploitation of channel knowledge at the transmitter.

![Figure 4-26. CDF of capacity for the four sets of measurements. The H matrix is normalized to a fixed value so the mean SNR is 20 dB for the 4 sets.](image)

![Figure 4-27. CDF of capacity for office-like scenario (Tx at location A, Rx at location 3), for different antenna modules.](image)
4.4 General conclusions, contributions and further research on MIMO prototyping and dual-polarized MIMO channel measurements.

A great effort has been carried out in the research community to gain a better insight into the MIMO channel characteristics and their effect in MIMO algorithms. However, it seems clear that this is still an open research area and some aspects remain to be clarified or improved.

Many channel sounders and MIMO testbeds have been previously presented in the literature, with different objectives and covering different frequencies bands. However, most of them lack reconfigurability and flexibility, and some of them are based on very expensive platforms. We have proposed a simplified MIMO testbed architecture, which allows a simple testing of MIMO algorithms, antenna architectures and channel sounding. An inexpensive and easy to update system have been implemented, which allows to perform channel measurements without having to acquire very expensive or difficult to operate equipment. As a price for simplicity, the system works in an off-line basis, as opposed to the smart antenna presented in Chapter 3. However, this gives the user the option to pre- and post-process the signal off-line in a dedicated PC, so no real-time constraints have to be taken into account in the algorithms design. By comparison with the design and implementation stages in both systems, it is clear that the new testbed is a good example of some of the good design practices to be considered when simplicity and rapid prototyping is preferred to real time operation.

We have presented the implementation of two types of antenna modules, given the interest in studying their effect in the MIMO systems: single-polarized monopole antenna arrays and dual-polarized dipole antenna arrays. Both of them present similar features in terms of low directive radiation pattern, high efficiency and good impedance matching. However, the former type of element shows a slightly higher directivity (due to the ground plane of monopoles) while the second one may offer a better diversity due to the dual polarization usage.

Recently, dual polarized antennas have been proposed as a solution to reduce size of multiantenna user devices to be used in future communication systems. However, the characteristics of MIMO channels when multiple polarizations are used are not yet clear for all scenarios. In order to get a better knowledge of the MIMO channel when dual polarization is used, a measurement campaign was carried out, by using the implemented testbed and antenna modules. Different routes along indoor hallways and offices scenarios were covered, including indoor and outdoor TX locations. From the path loss analysis, we conclude that an outdoor-to-indoor scenario may yield a smoother distribution of power thanks to the avoidance of LOS between transmitter and receiver. The capacity analysis when perfect power control is assumed (as typically done from a information theory perspective) reveals that in general the indoor-to-outdoor scenario gives a lower variance in capacity, while the LOS-NLOS changing conditions in indoor scenarios yield to a more variable capacity, which may be less interesting in order to guarantee a certain quality to the users. However, the price to be paid is a higher transmitted...
power required. Moreover, a higher capacity was achieved when using dual-polarized antennas in outdoor-to-indoor case. However, this conclusion is based on the assumption of perfect power control at the transmitter and no limitations of transmitted power, as it is usually in a dense urban environment of wireless systems.

Finally, we have shown the interest in evaluating the path loss effect and antenna influence in the MIMO capacity. When the channel matrix is normalized assuming a fixed transmitted power, the analysis of measurements shows that a higher capacity is in general obtained for the single polarized elements used in the measurement campaign. This may be explained by the higher gain of these antennas. Therefore, the antenna gain should also be considered when evaluating the antenna performance in the MIMO systems, since perfect power control cannot always be assumed.

**Contributions**

The contributions of this chapter of the thesis to the current state of the art can be summarized as follows:

- A novel MIMO testbed has been designed, implemented and its main features tested. Although the implementation of a MIMO testbed itself is not new in the research field, some specific concepts as rapid prototyping based on off-line pre and post-processing, reconfigurability and simplicity have been applied in a novel way, aiming at obtaining an improved system regarding usability and multi-purpose scheme. It is worth mentioning that three objectives or applications have been integrated and tested with the MIMO testbed: channel sounding, algorithm testing and antenna array configuration for MIMO.

- By using the implemented testbed, a new measurement campaign was conducted to compare the MIMO system performance with either single-polarized or dual polarized antennas. As a novelty, both indoor and indoor-to-outdoor scenarios were measured, by selecting two types of locations for the transmitter module (“base station”): an indoor one (similar to the conventional WiFi network deployment with indoor access points) and an outdoor one (maybe more typical for cellular networks, but also of interest for new WiMAX communications). The interest in studying these two locations is undeniable for system operators and network deployers, which must choose the best location for base stations or TX-RX fixed units. Some conclusions have been presented from the measurement campaign, as the lower variance in capacity that is obtained with the outdoor location, to the cost of a higher transmitted power required, or the higher capacity in MIMO systems when no CSI is assumed and dual polarized antennas are used, compared with using single-polarized ones.
Different normalization methods to be considered with measured channel matrices $\mathbf{H}$ in a MIMO system have been proposed, showing the interest of each of them and how the results are also different. While conventionally the channel matrix is normalized to its Frobenius norm (assuming perfect power control in the system) we have shown that it is also important to consider the different antenna gains in the system, especially when several antenna arrays for MIMO are to be compared. Also the differences with the case when no power control is used are presented.

**Further research**

Despite the great effort in developing MIMO testbeds and prototypes, and also in improving the MIMO channel knowledge and its main features, there are still some open issues to be tackled before extensively deploying multi antenna solution in radiocommunication systems.

First of all, it is important to mention that agreement in the standardization is a must. Several standardization works are currently being carried out within both the 3GPP body and the IEEE. Several studies have been published by the 3GPP working groups, and the second version of draft 802.11n has recently been presented. However, no commercial MIMO equipments have been launched yet, since most manufacturers are waiting for final standards to be compliant with. Nevertheless, some pre-commercial MIMO equipments and devices can be already found.

Some more channel measurements would be desirable in order to improve channel knowledge. As an interesting way to reduce the antenna array size, multi polarization in MIMO is of great importance. Measurements in outdoor environments to better characterize this aspect are of interest. Also of special interest is the characterization of MIMO channels from the multi-polarization point of view when not only 2 but also a third polarization is considered. Especially for MIMO indoor scenarios, multiple scattering and multipath behaviour can be expected, so the radio waves experience depolarization in every angle or polarization angle. A 3D channel model was presented in [142] and a 3-polarization antenna was proposed in [156], but no measurements with this type of scenario have been presented to the knowledge of the author.
PROPAGATION ASPECTS OF MIMO SYSTEMS:
Analysis with multiple base station locations and multiple scenarios

5.1 Introduction and motivation ................................................................. 190

5.2 Measurement system and considered scenarios ................................. 193
  5.2.1 MIMO testbed used in the measurement campaign ..................... 193
  5.2.2 Data processing and channel estimation procedure ...................... 197
  5.2.3 Measurement environment .......................................................... 201

5.3 Evaluation of measurements and analysis of results: general channel characteristics and single-user analysis ................................................................. 204
  5.3.1 General model ............................................................................ 204
  5.3.2 Analysis of spatial correlation in MIMO systems for multiple scenarios .. ................................................................. 205
  5.3.3 Path loss, fading and coverage analysis ........................................ 211
  5.3.4 Capacity analysis and different methods of combining base stations .. 215

5.4 Multi-user-oriented evaluation of measurements and analysis of results... .................................................................................................................. 220
  5.4.1 Multi-user MIMO system model and sum capacity problem .......... 220
    5.4.1.1 Multi-user MIMO downlink ................................................... 220
    5.4.1.2 Multi-user MIMO uplink ....................................................... 222
  5.4.2 Evaluation of multi-user MIMO at downlink: studied scenarios and capacity analysis ................................................................. 223
    5.4.2.1 Downlink scenario and capacity expressions ............................... 223
    5.4.2.2 Results for multi-user MIMO downlink ..................................... 224
  5.4.3 Evaluation of multi-user MIMO at uplink: studied scenarios and capacity analysis ................................................................. 227
    5.4.3.1 Uplink scenario and capacity expressions ................................... 227
    5.4.3.2 Results for multi-user MIMO uplink ......................................... 228

5.5 General conclusions, contributions and further research on propagation aspects for MIMO systems ................................................................. 231
The importance of properly knowing and characterizing the MIMO channel is undeniable, as has previously been stated in previous chapters. It allows a better design and optimization of MIMO schemes and algorithms, taking into account the characteristics of multipath propagation and delay, frequency dependence, temporal fading, etc. Moreover, novel methods to improve the achievable diversity in the system can be proposed, as using combined methods to include spatial and polarization diversity.

The previous chapter presented some results regarding capacity computation and power maps, obtained from a measurement campaign were several measurement points were included, and where a broadband MIMO testbed was used. In this chapter, a different measurement campaign is presented, where the emphasis is put in studying different locations and combination methods for base stations to be used in MIMO systems.
5.1 Introduction and motivation

When defining and developing novel communication systems, experience has taught us that knowing the propagation issues and the channel characteristics is a must if a successful development and fast deployment is desired. This way the system parameters to be tuned can be properly optimized at an early step before deploying the network and the system. This is of utmost importance to get a more effective network and a better coverage. For example, the delay in final launching and deployment of 3G systems (UMTS in Europe) was caused, among others, by the reduced coverage of the system and the lack of detailed knowledge of the best locations and distribution of base stations. Moreover, hand-over between UMTS and already deployed networks such as GSM was an issue not properly solved at first stages of the deployment of the system [243]; the time to overcome this problem could have been reduced if thorough simulations of the system including proper channel models had been done when designing it.

The lesson has been learnt and now the introduction of MIMO systems for 3G and later 4G systems is being supported by a wide effort in better knowing the MIMO channel, where spatial diversity is considered. As an example, the 3GPP body has promoted some working groups that have worked on defining an adequate channel model for 3G-MIMO communications. Despite this, it is common to make several assumptions to simplify the study and evaluation when designing algorithms and schemes for MIMO systems, such as ideal antenna arrays and adequate richness of separated multipath. However, in order to be able to predict the performance of a MIMO system in a realistic scenario, either detailed propagation simulations or measurements in real environments are required. Thus, the interest in realizing new MIMO measurements to better characterize the channel is clear.

From the point of view of an operator and system deployer, one key issue is having enough information about parameters that have to be considered to select the base stations locations. Aspects such as achievable coverage and required transmit power as a function of type of scenario, available capacity and throughput of the system, and so on, are of great importance to offer a good service to users. In spite of its interest, this aspect has not been deeply treated so far, as we will discuss later.

Regarding MIMO channel characterization, a current topic of discussion is the trade-off between received power and rich multipath. A high SNR, as in line of sight (LoS) situations, may imply a low degree of scattering and spatial diversity [244]. On the other hand, non-line of sight (NLoS) cases suffer from higher path losses and thus lower received power than LoS ones, which may involve lower capacity for similar measured scenarios [96]. It is known that both factors (power and multipath richness) contribute to ergodic capacity, but it is not clear how to characterize their impact and importance, depending on the environment. In most works, it is common to see normalization of the channel matrix $H$ to the instantaneous received power (or
fixed signal to noise ratio, SNR, at the receiver) [242]. This is equivalent to assuming ideal power control in the system, so the path loss effect is not included. In our opinion, it is also of interest to consider the channel path loss and its relation to the transmitter and receiver location. In [64] the capacity for LoS and NLoS fixed indoor positions is compared with and without power normalization. In [63] the authors proposed the normalization of the MIMO channel based on the average received SNR for the whole route in an outdoor scenario, so a fixed transmitted power is assumed. Comparison of different normalization methods and their analysis in different scenarios are currently open issues. In this chapter, we address this comparison. In order to complete the analysis that was presented in chapter 4 based on measurements with a multipurpose testbed, a new and extensive measurement campaign was realized, which covered many routes and different scenarios.

Many measurement campaigns aiming to characterize the MIMO channel have been reported in the literature (see chapter 2, section 2.2.5 for a summary). Despite the remarkable effort in characterizing and measuring MIMO channels, most of the previous works focus on either the indoor or the outdoor case. However, the outdoor to indoor scenario has important applications for data transmission in third generation cellular systems, as well as wireless local area networks (WLAN). The user equipment may be indoor while the base station may be located on a rooftop. One of the few examples that includes this type of scenario is [102], where a measurement campaign conducted to validate a channel model is presented. Both indoor and indoor-to-outdoor measurements are included in the study, mainly aiming to check the proper behaviour of the channel model in different scenarios. In [245] measurements and data evaluation for an outdoor to indoor case is presented, where the work is focused on the statistical distribution of the signal and direction of arrival. A similar study is [246], which also includes outdoor scenarios and compares the angle and path distance distribution for both types of environments. Although some preliminary studies of indoor-outdoor environments have been done, most of them focus on a specific scenario and do not consider multiple transmitter locations, and the capacity analysis are scarce. Moreover, most of the previous works consider a single BS in the MIMO system. Thus, a more complete capacity analysis is needed, with the aim of examining several options for the BS location and their configuration scheme at system level. We propose in this chapter some methods to combine the signal of multiple-antenna base stations (BS), with and without channel state information (CSI) at the transmitter, and different BS locations are compared.

Moreover, in order to apply the MIMO concept to realistic MIMO systems, it is interesting to study not only link-level aspects, but also system-level ones, as multiple access methods, higher level algorithms, interference due to other users, etc. Including multiple users in the study, their effect and the type of algorithms to mitigate the intercell and intracell interference is of key importance from a system operator point of view. Some previous work has been done in this area, mainly theoretically. In [63], a multicell MIMO system is analysed,
where the channel is modelled by using the double-directional channel model, but no
measurements or co-channel interference are presented. A work that study multi-user MIMO
throughput is presented in [247], where only co-channel interference is considered (only one base
station is included), and the channel is again a simulated one. Thus, evaluating the multi-user
MIMO schemes with real channel measurements is of interest to complement multi-user MIMO
studies. We have used several of the routes that were measured in the measurement campaign
presented here to analyse the effect of intracell and intercell interference for different base
station locations and different signal processing approaches.

To sum up, the objective in this chapter is twofold: to investigate the use of multiple
base stations in an indoor environment, and to contribute to a better characterization of the
properties of outdoor-to-indoor propagation. In all cases, we have measured the channel matrix
as a function of location. This allows us to study not only the statistical properties of the
channel, but also how the coverage varies with the exact office environment. Channel matrix
normalization assuming fixed transmitted power or fixed received SNR power were considered,
in order to give a better insight into the effect of received power and available spatial diversity
in the capacity. Several schemes with one or two base stations are also investigated, and the
system improvement obtained when channel state information is available at the transmitter is
also shown. Also multiple users are considered in order to study the effect of intracell and
intercell interference in real indoor and outdoor-to-indoor scenarios.
5.2 Measurement system and considered scenarios

A narrow-band MIMO testbed, developed in the Department of Signals, Sensors and Systems, KTH, was employed to perform a measurement campaign for different scenarios, all of them having place in KTH University, Stockholm. The obtained data were used to evaluate several parameters and characteristics of the MIMO channel. A general description of the measurement system and studied environments is given below.

5.2.1 MIMO testbed used in the measurement campaign

The measured data that were used to study and evaluate different base station locations and configurations were collected with a 4 by 8 DSP-based testbed, whose general architecture is depicted in Figure 5-1 (transmitter modules) and Figure 5-2 (receiver modules). We may note that both figures depict the RF chains in a schematic way: in the actual implementation 2 frequency conversions and several amplification and filter stages were used. The testbed is a simplified version of the one presented in [183]. The system operates in an off-line basis, so the received signal is first stored and after that post-processed in a personal computer. The system bandwidth is only 9.6 kHz, which allows narrow-band channel measurements. The off-line and narrow-band features simplify the system operation, since neither real-time constrains nor broadband equalization need to be considered. Moreover, these are sufficient features considering the studied characteristics of the MIMO radio channel.

![Figure 5-1. Illustration of hardware transmitter modules for the used narrowband testbed.](image)

![Figure 5-2. Illustration of hardware receiver modules for the used narrowband testbed.](image)

The carrier frequency of the system is 1766.6 MHz. The radiofrequency chains are implemented with plug-in commercial components, including amplification and frequency conversion stages. A heterodyne scheme with 2 intermediate frequencies is used, for both the
transmitter and the receiver chains. For a thorough explanation of the radiofrequency hardware, [248] can be consulted.

In order to study different transmitter configurations, the 4 transmitters were split into 2 groups of 2 transmitters each. The interest in studying this configuration at the transmitter side will be explained later. The digital signals to be transmitted by each 2-TX group were synchronously generated in a TI 6713 DSP, which was controlled by a laptop. However, we may note that the transmitted signals from different TX groups are not synchronous. Therefore, there is a phase uncertainty in the measured channel matrix for transmitters in different groups.

The generated signals are digitally up-converted to the lower intermediate frequency, \( f_{\text{IF,low}} = 10 \text{ kHz} \), and then analog-converted with 16-bit precision and a sampling rate of 48 ksps. The radiofrequency chains up-converts the signal from IF to RF, and the signal is also amplified and filtered. Since signals from different transmit groups are not synchronized at the digital stage, so there is no need to preserve phase synchronization in the analog stage, two different sources are used as sinewave oscillators to mix and upconvert the signals in the two TX groups. This way the two TX groups can be placed in separated locations without the need of connecting the oscillator sources to get synchronized signals, which gives higher flexibility when testing different locations for base stations. Figure 5-3 shows a photograph of the transmit modules, for the outdoor location.

![Figure 5-3. Photograph of transmit modules. The 4 output channels were mounted together for the outdoor location of base station.](image)

In the receiver side, the received RF signal is amplified and filtered. A downconversion stage is performed. Conversely to the TX scheme, at the RX the same local oscillator is used for the 8 receiver inputs, so the received signals are synchronized. The IF signal is analog to digital
converted and then collected by a data acquisition board with up to 8 analog inputs, namely the NI-PCI6071E from National Instruments. The sampling rate is 40 ksps, with 12-bit precision. Here the same clock is used to control all the ADCs and choose the sampling instant, so there is no phase uncertainty among the received inputs. A dedicated PC is used to control the board and store the raw data, which is received via PCI bus. After data collection, the files are post-processed using Matlab™.

The receiver modules are mounted on a trolley, so the receiver side is carried along different routes. A photograph of the receiver modules is shown in Figure 5-4.

The transmitted power per chain was limited to -5 dBm in the indoor to indoor environment, which is sufficient for the studied picocell scenarios, presented in next section. In order to analyse an outdoor to indoor environment, extra amplification was included for the measurements realized in these scenarios. With this extra amplification, a maximum output power per chain of 25 dBm was obtained.

![Figure 5-4. Photograph of receiver module with 8 receive inputs. The system was mounted on a trolley so it can be easily moved along indoor routes.](image)

Regarding the antenna arrays, different options are considered in each link end. For the transmitter end, Huber-Suhner dual-polarized planar antennas with slant linear polarization (±45°) are used for indoor locations. These antennas are typically used for communications in indoor environments, since thanks to their low profile and small size they are easily placed at walls or on equipments or furniture. On the other hand, Powerwave broadband dual-polarized ±45° antenna arrays are used for outdoor transmitter locations. These antennas are common ones to be used in radio links or outdoor communication scenarios.

The nominal gain for the indoor antennas is 8 dBi and the 3dB beamwidths are 85° and 75° in the horizontal and vertical plane, respectively. The 2 antennas for indoor locations were
spaced 0.6λ. The antennas for outdoor locations have a nominal gain of 15.1 dBi and 3dB beamwidths of 67º (horizontal) and 14.7º (vertical), and were situated with a spacing of around 4.7λ (~ 0.8 m). Figure 5-5 shows a photograph of each of the antenna types, mounted on one of their location for the measurement campaign. Appendix A gives more information about the characteristics of the TX antennas.

![Antennas](image1)

Figure 5-5. Antennas used for the TX side, for indoor (left) and outdoor (right) BS location. Although three 2-port antennas were available for the indoor location, only 2 were used (one for each TX group).

For the receiver end, two 4-element antenna arrays were designed and implemented. The first one is a conventional linear λ/4 monopole array with element spacing d = λ/2, which may be used as a reference antenna. The second one is a compact antenna array that consists of 4 PIFA elements. Since this chapter focuses on the channel characterization for different transmitter locations, only the received signal from the monopole antenna array will be considered thereafter. Details on the design and implementation of the compact array, as well as a comparison of performance from a point of view of MIMO system for the two arrays can be found in chapter 6.

The monopole array was designed in a similar way as the one used for the measurement campaign presented in chapter 4. Since a lower frequency (1.766 GHz) than the one used in previous measurement campaign (2.45 GHz) was considered, a larger array was obtained. The ground plane is made of a brass plate measuring 170 × 425 mm², that is, approximately 1λ × 2.5λ at the central frequency f₀ = 1.766 GHz. This size is big enough so the ground plane can be considered as an ideal infinite ground plane for far field measurements in the upper semispace (z > 0). Each monopole element is made of a brass rod with 3 mm of diameter, and was fed by an SMA coaxial connector. They were placed on the ground plane with a spacing d = λ/2. The length of each element was selected as slightly smaller than the ideal dth = λ/4 = 4.25 cm, in order to take into account the currents on the top of the rod (due to not having a zero-diameter “rod” or current line). The Tai equation was used to initially calculate the length to be used for the monopole arrays. Afterwards, simulations of the monopoles with the 3D electromagnetic...
simulation software CST Microwave Studio were performed, in order to optimize the length of the monopoles and the connector size so good return losses were achieved. Finally, the implemented monopole arrays were characterized by measuring its main radiation and matching parameters. A return loss coefficient lower than -15 dB was achieved for all the elements, as well as a quite omnidirectional radiation pattern, typical for a monopole antenna. Since this chapter is devoted to propagation and channel measurement, details on the monopole are skipped here. More information about both the design and the features of the monopole array are given in chapter 6, where the aim is put at evaluating antennas in MIMO systems. Figure 5-6 shows a photograph of the RX antenna array mounted on its placement (a movable trolley) for the measurement campaign.

Figure 5-6. RX antenna on the trolley used to carry out the measurement campaign.

5.2.2 Data processing and channel estimation procedure

There are different methods to estimate the MIMO channel matrix by using radio channel measurements. The final procedure and also the results depend not only on the post-processing modules but also on the transmitted signal. Next paragraphs explain the methodology that was followed in this measurement campaign to estimate the $H$ matrix.

A digital sinewave was chosen as transmitted baseband signal for the measurements. A different frequency was used for each transmitter in order to be able to separately detect each transmitted signal in the receiver, and thus properly estimate all the elements in the channel matrix $H$. The used frequencies for the baseband sinewaves were chosen as $f_{TX1} = 1 \text{ kHz}$, $f_{TX2} = 0 \text{ kHz}$, $f_{TX3} = -1 \text{ kHz}$ and $f_{TX4} = -2 \text{ kHz}$, as represented in Figure 5-7 and summarized in Table 5-I.
Since very close frequencies were chosen, the frequency channel response can be considered flat in the whole measured bandwidth. By using a simple sinewave instead of pseudonoise codes or more complex signals, the required signal processing to estimate the channel matrix $\mathbf{H}$ is simplified, but it is still accurate to analyse the narrowband properties of the measured scenarios. We may note that complex signals are represented in the DSPs, which allows the use of a spectrum that is not symmetric around $f = 0$ Hz for the digital baseband signals. Figure 5-8 depicts the digital signals to be transmitted by each TX antenna, divided into the two TX groups, and the actual received signal, both at baseband.

Figure 5-8, Frequency allocation for the 4 sinewaves that were transmitted (one for each TX output) and actual received signal, where frequency drift and the filter effect can be observed.

In Figure 5-8 a), not ideal Dirac deltas are obtained due to the window effect when computing the Fourier Transform with a finite number of samples (19800 samples are used in
the representation). The two TX groups were represented separately to make it clearer that each TX group has a different local oscillator to upconvert signals from the upper intermediate frequency $f_{\text{IF}} = 70$ MHz to radiofrequency $f_{\text{RF}} = 1766.6$ MHz. This fact introduces two main effects:

* Since the transmitter and the receiver use different local oscillators (LO), a certain frequency drift is expected. However, since different LOs are used at different TX groups, a different drift may be observed at the receiver for each TX group. This effect can be observed in Figure 5-8, where we see that, for the presented measurement, the received baseband sinewaves are centered at -2091 Hz and -1087 Hz for group 2 (a frequency offset of approximately -90 Hz) but at -351 Hz and 658 Hz for group 1 (a frequency offset of approximately -350 Hz). Obviously, those values are only shown as example and they may vary with the time instant when measurement is accomplished (due to warming of equipments, etc). This effect is considered when estimating the $H$ matrix.

* There is a phase uncertainty between TX signals that belong to different TX groups, that is, the estimated values of $H$ matrix will experiment an unknown phase rotation due to the frequency offsets in the local oscillators of the TX groups. Furthermore, the different D/A sampling clocks at each TX group introduce another degree of uncertainty in the phase of different TX groups. As a result, the relationship between the estimated $H$ matrix and the real $H$ matrix could be expressed as:

$$H_{\text{meas}} = H_{\text{real}} \cdot D_{\Delta f}$$

(5-1)

where $D_{\Delta f}$ is a diagonal matrix expressing the frequency offsets from the frequency at the receiver and the actual transmitted frequency:

$$D_{\Delta f} = \begin{bmatrix}
e^{j2\pi f_1} & 0 & 0 & 0 \\
0 & e^{j2\pi f_1} & 0 & 0 \\
0 & 0 & e^{j2\pi f_2} & 0 \\
0 & 0 & 0 & e^{j2\pi f_2}
\end{bmatrix}$$

(5-2)

Several steps were performed in order to estimate the channel matrix $H$. The process is based on estimating the received frequencies and after that correlating the received signals for each of the RX inputs with the ideal signal to be received. Figure 5-9 shows graphically the steps that were carried out.
Chapter 5

Estimation of received frequencies (at baseband)

Initialization

File with collected data from one measurement

Loading “raw” digital received data

Digital downconversion from lower IF to baseband

Nominal TX frequencies $f_{Rx1}, f_{Rx2}, f_{Rx3}, f_{Rx4}$

Digital baseband filtering

Correlation of received signals with complex exponentials $e^{j2\pi f}$

Calibration tables (Tx and Rx)

Estimation of received frequencies (at baseband)

Calibration

H downsampling

H_{MIMO}

Figure 5-9. Data processing to estimate H matrix from radio measurements

After a previous initialization of the process, the digital data that were collected in one route are loaded. This signal was digitalized and stored at the lower IF. Thus, a digital downconversion is required to obtain baseband information. In order to remove the produced signal at the double of IF frequency, a digital low-pass filter is used, with a 3dB bandwidth of approximately 5 kHz. Its effect can be observed in Figure 5-8 b), where we see how the non-desired signal copies at $f = 2 \cdot f_{IF,low}$ are highly attenuated. After that, the estimation of the received baseband frequencies is done, by using the Fast Fourier Transform (FFT) and averaging over each second. This way the possible frequency drifts along the time (due to warming of equipments) are corrected. The nominal transmitted frequencies are used as input to find the received frequencies more easily.

The actual estimation of the H matrix is performed by correlating the received signal with a complex digital exponential for each transmitter $i$, $e^{j2\pi f_{Rx}}$. This operation is done with the signal received by each receiver (each RX input). Therefore, the estimated H matrix is an $N_R \times N_T$ matrix, as expected.
Due to small differences in the hardware components, the receiver chains present slightly different gains and phase offsets; the same effect appears in the transmitter chains. In order to mitigate its effect in the estimated $H$ matrix, a calibration stage is performed before processing the collected signal. The calibration values are previously obtained from back-to-back measurements on the testbed and stored in a calibration table. It was computed every time the system was switched on, that is, one calibration table was obtained for each measurement day.

Finally, in order to simplify the post-processing of data with the estimated $H$ matrices and also for the sake of reducing storage requirements, the estimated $H$ matrices were downsampled from having 400 samples per second to having 100 samples per second. Since the MIMO channel was measured at a pedestrian speed, 100 samples per second are more than enough to observe fast fading with a good accuracy. Figure 5-10 shows an example of estimated $H$ matrix obtained for one route where the receiver was moved at c.a. 1 m/s and there was NLOS between TX and RX.

![Estimated H Matrix](image)

**Figure 5-10. Example of some elements of the estimated H matrix for one NLOS route**

Many routes were measured and different TX and RX locations were considered. Next section explains all the considered scenarios and presents the measured environments.

### 5.2.3 Measurement environment

The measurement campaign was carried out in the S3 building and surroundings, in the KTH campus, Stockholm. Several scenarios were included, with especial emphasis on the consideration of different locations and configurations for the transmitter.

The measurements were divided into 4 groups, regarding the transmitter locations. The receiver was moved along indoor routes for all the cases. The measured cases were:
- **Case A**: the 4 transmitter antennas (A12, A34) were located at one end of the 4th floor in the S3 building.

- **Case B**: the transmitters were split into 2 groups (B12 and B34) or base stations (BS) and each BS was placed spatially separated at the same end of the 4th floor in the S3 building.

- **Case C**: the transmitters were split into 2 BS (C12 and C34) and each one was located at a different end of the 4th floor in the S3 building (maximal spatial separation).

- **Case D**: the 4 transmitter antennas (D12, D34) were located at the flat roof of the Q building (in front of the S3 building).

Case D consists in an outdoor to indoor scenario, while the other cases are examples for indoor scenarios. Thus, 2 different scenarios are addressed in the measurement campaign.

The receiver modules were carried along 22 indoor routes covering 3 different floors of S3 building (floor 3, 4 and 5) at a pedestrian speed (approximately 0.9 m/s). While floor 4 and 5 include mainly hallways and offices, laboratory rooms and a hall are found in floor 5, with a slightly different plan. Similar routes were conducted for the 3 floors. The measurements included situations of line of sight (LOS) and non-line of sight (NLOS), as well as routes in the offices. The same routes were repeated for each case of transmitter location, which allow to assume that very similar points are measured for each TX location. Figure 5-11 represents the floorplan for floor 4, where the main routes and the position of TX for the different cases are also depicted.

![Floor plan (4th floor) and locations for base stations 1 and 2 for the 4 cases in the measurement campaign.](image)

To give an idea of the required resources that were need to accomplish the measurement campaign, the time to carry out the measurement campaign for all the routes and the TX
locations was 4 days, including every-day calibration and set-up of the equipment. Post-processing the raw data required also a great amount of processing requirement, needing about 3 hours to just estimate the H matrices of one day of measurement with a conventional desktop PC of 2.4 GHz and with RAM 1GB (once the software to estimate H was ready).

Table 5-II summarizes measurement setup characteristics for the considered environments. Figure 5-12a) presents a picture of the indoor scenarios from the point of view of the transmitter for Case A (very similar to the one in cases B and C), while Figure 5-12b) shows the outside view of S3 building from the outdoor location of transmitters (Case D).

<table>
<thead>
<tr>
<th>Cases A, B, C</th>
<th>Case D</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Tx Location</strong></td>
<td>Indoor</td>
</tr>
<tr>
<td><strong>Tx power / branch</strong></td>
<td>-5 dBm</td>
</tr>
<tr>
<td><strong>Tx antenna elements spacing (for each BS)</strong></td>
<td>0.6λ</td>
</tr>
<tr>
<td><strong>Rx antennas spacing</strong></td>
<td>0.5λ</td>
</tr>
<tr>
<td><strong>Polarization at Tx</strong></td>
<td>slanted ±45° linear</td>
</tr>
<tr>
<td><strong>Polarization at Rx</strong></td>
<td>vertical (linear)</td>
</tr>
<tr>
<td><strong>Carrier frequency</strong></td>
<td>1766.6 MHz</td>
</tr>
<tr>
<td><strong>System BW</strong></td>
<td>9.6 kHz</td>
</tr>
<tr>
<td><strong>Time resolution after channel estimation</strong></td>
<td>0.01 s</td>
</tr>
<tr>
<td><strong>MS speed</strong></td>
<td>0.9 m/s</td>
</tr>
<tr>
<td><strong>Number of Tx and Rx elements (NT x NRx)</strong></td>
<td>4 × 4</td>
</tr>
</tbody>
</table>

Table 5-II. Main characteristics of measurement setup

Figure 5-12. Scenario seen from TX antennas for indoor (left) and outdoor (right) locations
5.3 Evaluation of measurements and analysis of results: general channel characteristics and single-user analysis.

From the measured $\mathbf{H}$ matrices several parameters have been computed and analysed. In this section, results regarding general channel characteristics (spatial diversity, path loss...) and single-user aspects (link level results) are computed and discussed. Our study has been focused on some of the identified open issues that were mentioned in the introduction of this chapter, as the effect of considering non-ideal or even non-existing power control in the system, or the channel capacity when multiple base stations are used to cover the indoor area.

Firstly, we have calculated the propagation characteristics of the MIMO channel for different BS locations, comparing the indoor-to-indoor scenario with the outdoor-to-indoor one. Specifically, the cross-correlation between antennas at TX (and also at RX) has been computed, in order to evaluate the spatial diversity that can be achieved for the two types of scenarios.

Secondly, the capacity that can be achieved with two completely opposite methods has been computed: capacity with no channel state information at the transmitter and capacity with perfect channel state information at the transmitter. Several options for combining the signals at the base stations level are proposed and compared here. This may be of great interest for future radio communication systems, where a certain collaboration or cooperation between BSs or access points can be expected, in order to improve the system performance. This analysis is done at link-level, that is, for a single user, as opposed to the one done for multi-user cases presented in next section.

5.3.1 General model

For the rest of the chapter, we will consider an $N_{Tx}\times N_{Rx}$ MIMO system, being $N_{Tx}$ the number of transmit antennas and $N_{Rx}$ the number of receive antennas. Let us remind that the input-output relationship for a narrowband MIMO channel is expressed as the well known expression:

$$y(t) = \mathbf{H}(t)x(t) + \mathbf{n}(t)$$

where $y(t)$ and $x(t)$ are the received and transmitted signals respectively, and $\mathbf{n}(t)$ is a vector of additive white Gaussian noise with variance $\sigma^2$. The channel matrix $\mathbf{H}(t)$ consists of $N_{Rx}\times N_{Tx}$ elements, $h_{ij}(t)$, which are the complex channel gains between transmitter $j$ and receiver $i$. We must observe that this expression is only valid for narrowband ("flat fading") channels, as in the case of the measurement campaign presented here.

In order to study some of the properties of the channel, the correlation between $\mathbf{H}$ matrix elements is of interest. This could be computed as the so called the channel correlation matrix, $\mathbf{R}_H$:

$$\mathbf{R}_H = \text{vec}(\mathbf{H}^H)\text{vec}(\mathbf{H}) \in \mathbb{C}^{(N_{Rx}N_{Tx})\times(N_{Rx}N_{Tx})},$$

204
where \( \text{vec}(A) \) is the operation of situating all the elements of the matrix \( A \) in a single column (column-wise ordering). In \( \mathbf{R}_H \) all the correlation information about the channel (including relationship between TX and RX statistics) is kept. We see, however, that keeping all this information has the cost of having to use a very big matrix (\( N_{Rx} \times N_{Tx} \) rows, \( N_{Rx} \times N_{Tx} \) columns).

To simplify the operation and use of the correlation matrix, sometimes the correlation matrices at the transmitter \( \mathbf{R}_{Tx} \) or at the receiver \( \mathbf{R}_{Rx} \) are calculated, whose mathematical expression was presented in Chapter 2, eq. (2-24), and is rewritten here to refresh the concept to the reader:

\[
\begin{align*}
\mathbf{R}_{Rx} &= \mathbf{H}^H \mathbf{H} \in \mathbb{C}^{N_{Rx} \times N_{Rx}} \\
\mathbf{R}_{Tx} &= \mathbf{H}^H \mathbf{H} \in \mathbb{C}^{N_{Tx} \times N_{Tx}}
\end{align*}
\]

(5-5)

Roughly speaking, the correlation matrix \( \mathbf{R}_H \) represents how similar/different the elements of the \( \mathbf{H} \) matrix are one from the other, thus giving an idea of the available spatial diversity in the MIMO system thanks to using multiple antennas at TX and RX. \( \mathbf{R}_{Tx} \) and \( \mathbf{R}_{Rx} \) has a similar meaning, but only including statistics at one side of the link.

For the forthcoming analysis, we have assumed a single user MIMO scenario. Since the fixed link end (base stations) are the transmitters and the mobile unit consists of receivers, we can considered that the system is a downlink one; however, since the \( \mathbf{H} \) matrices were computed, we may assume a certain reciprocity between transmitter and receiver (as long as the frequency band is kept the same).

For all the analysis we have assumed that the noise level during the measurement campaign was low enough to be neglected, so the estimated \( \mathbf{H} \) matrices do not include a relevant noise level.

### 5.3.2 Analysis of spatial correlation in MIMO systems for multiple scenarios

We begin the propagation analysis by studying the spatial correlation properties of the measured scenarios. From a point of view of MIMO system, the spatial correlation properties of the channel are of paramount importance, since they represent the extra spatial diversity that is introduced with the use of multiple antennas, and thus they show the possible advantages of using these systems.

It has been shown [102] that the spatial correlation matrices at the transmitter \( \mathbf{R}_{Tx} \) and receiver \( \mathbf{R}_{Rx} \) can be used to estimate the correlation matrix \( \mathbf{R}_H \) of a MIMO system in some cases such as indoor NLOS, which in turn gives a direct insight into the achievable spatial diversity and MIMO capacity. This approximation of \( \mathbf{R}_H \) by using \( \mathbf{R}_{Tx} \) and \( \mathbf{R}_{Rx} \) assumes that the correlation matrices at the transmitter and the receiver are separable, that is, the statistics of signal at the receiver side are independent of the ones at the transmitter: it supposes that, on its
travelling along the MIMO channel, the radio signal “forgets” the conditions at the transmitter (direction of departure, angle spread, scattering position and density…) when it arrives at the receiver. With this assumption, the correlation matrix $\mathbf{R}_H$ can be expressed as:

$$\mathbf{R}_H \approx \mathbf{R}_{Tx} \otimes \mathbf{R}_{Re}$$

This is a quite rough simplification, but it has been widely used and it allows the use of a very interesting simplification of the channel. This model has been presented in [102], including indoor measurements to support it. It has been usually called the Kronecker model, and since its presentation it has gained many supporters and also detractors (see chapter 2, section 2.2.6 for a brief introduction on the Kronecker model and some references to its advantages and shortcomings). The model has been proven to be accurate enough for indoor scenarios, as the one we study here. Based on this, we study here the correlation at the transmitter and receiver side.

When computing the spatial correlation coefficients between two antennas, two options may be considered: either the complex information is taken into account or else the phase is discarded and only the power (envelope) information is used. In the context of modelling, the complex correlation coefficient $\rho^{\text{cplx}}$ is preferred, since it carries the full information (amplitude and phase) of the radio channel, which is required to properly combine the modelled multipath scenarios. However, the power correlation coefficients $\rho^{\text{pow}}$ have a clearer engineering interpretation than the complex correlation coefficient, which make them suitable for analysing correlation properties of a measured MIMO channel. Since we are interested in the analysis of the signal, studying the power correlation is fair enough. Moreover, it has been shown [249][250] that both correlation coefficients are linked. In the case of Rayleigh distributed signals, their relationship is given by the following expression ([251]):

$$\rho^{\text{pow}} = |\rho^{\text{cplx}}|^2$$

For indoor environments (as the one analysed in this work), we may assume multipath richness and Rayleigh distributed signals in general, so the expression above will hold in our case. For clarity reasons, we will hereafter refer to the power spatial correlation coefficient simply as correlation coefficient, $\rho$. We may note that the measured routes are mostly NLOS, thus (2) is applicable in most cases.

**Spatial correlation at the transmitter**

The spatial correlation at the transmitter is of great interest for our analysis, since it allows the study of spatial diversity offered by the different studied locations for the transmitter.

The correlation coefficient between transmitters $i$ and $j$ is then computed as:

$$\rho^{\text{pow},Tx}_{i,j} = \left( |h_{m,i}|^2, |h_{m,j}|^2 \right)$$
where \(\langle \cdot, \cdot \rangle\) denotes the correlation operation, defined as:
\[
\rho_{a,b} = \langle a, b \rangle = \frac{E\{ab\} - E\{a\}E\{b\}^*}{\sqrt{E\{a^2\} - E\{a\}^2} \sqrt{E\{b^2\} - E\{b\}^2}}
\]  \hspace{1cm} (5-9)

where \(E\{\cdot\}\) denotes expectation and \((\cdot)^*\) is the conjugate operation. The slow fading is removed by local averaging of the power over a distance of 1 m.

The correlation coefficients for each transmitter pair were computed for the 4 transmitter locations under study, considering all the measured routes. The computed cumulative distributed functions are depicted in Figure 5-13.

Figure 5-13. CDF of the spatial power correlation coefficients for transmit antennas. 4 locations are considered for the 2 base stations: indoor co-location (A), indoor medium spatial separation (B), indoor opposite location with larger separation (C) and outdoor location (D).
As expected, the correlation is much smaller when considering antennas in spatially separated base stations (cases B, C) than for cases where both base stations are closely located (cases A, D). The highest correlation scenario is found to be case A, where the mean correlation coefficients vary from 0.45 to 0.55. We may notice that in this case the polarization diversity in transmitters offers some interesting decorrelation, reducing the mean value of \( \rho \) from \( \rho_{13} = 0.53 \) to \( \rho_{14} = 0.45 \) and from \( \rho_{24} = 0.53 \) to \( \rho_{23} = 0.5 \).

When the base stations are spatially separated but in the same end of the office floors, as in case B, the correlation between antennas in different base stations is substantially reduced, due to the increase in spatial diversity obtained by separating the antennas. Moreover, the new location for the base stations causes that when one BS is received in LOS, the other one is received in NLOS, which also reduces correlation between base stations. The same effect holds for case C (base stations placed at opposite ends of the office floor), where the correlation is even lower. In this case, the antennas in different base stations are highly uncorrelated; the average \( \rho \) is close to 0.1 and there is a small variance around this value. Nevertheless, the use of different polarizations does not introduce extra decorrelation in these 2 cases, mainly due to the already low level of correlation.

When the outdoor location is considered for the transmitters (case D), a slightly lower correlation than in the indoor case (case A) is observed, which can be explained by the fact that the antenna groups were more closely located in case A \( (d=\lambda/2) \) than in case D \( (d=4.7\lambda) \). However, it is quite interesting to notice that, compared to the case of indoor TX location (case A), for the outdoor location the extra decorrelation obtained by using dual-polarized antennas is more important than the one obtained due to spatial separation, even though the spacing is larger: while the lowest average correlation coefficient is obtained for antennas pairs with spatial and polarization diversity (1-4 and 2-3), the highest one is observed for antenna pairs with the same polarization (1-3 and 2-4). Thus, lower polarization correlation was obtained for outdoor to indoor cases than full indoor ones.

**Spatial correlation at the receiver**

Similarly to previous analysis, the spatial power correlation coefficient between receivers \( i \) and \( j \) is computed as:

\[
\rho_{i,j}^{\text{pow,Rx}} = \left| \frac{\langle |h_{i,m}|^2 |h_{j,m}|^2 \rangle}{\sqrt{\langle |h_{i,m}|^2 \rangle} \sqrt{\langle |h_{j,m}|^2 \rangle}} \right|
\]  

(5-10)

It is expected that the results from the receiver point of view are less significant: since the receiver antennas were carried along the same routes for the four studied cases A to D, few differences will be found in the correlation coefficient studied for the overall measurement routes of each case.

Figure 5-14 presents the results for the spatial correlation at each receiver pair, for the four studied cases. We observe that the closer the elements are, the higher the correlation is
Propagation aspects of MIMO systems

\( \rho_1 < \rho_3 < \rho_2 \) and so on, as expected. Similar statistical distributions of \( \rho \) are obtained for the four measured situations. However, it is interesting to notice that the average \( \rho \) value is found to be slightly smaller for the outdoor location (see Table 5-IV), since there are less LOS (highly correlated) routes. This result demonstrates that the assumption of independent correlations at TX and RX is not always valid.

To sum up the previous results for TX and RX, and in order to study the average spatial correlation for each of the locations independently of the antenna, the CDF as a single curve considering all the transmit antennas (in the case of TX correlation) or the receiver antennas (in the case of RX correlation) is computed. The resulting curves as shown in Figure 5-15. Table 5-III shows the average power correlation coefficient for the transmitter, while Table 5-IV contains the average power correlation coefficient for the receiver.

Figure 5-14. CDF of the spatial power correlation coefficients for receiver antennas. 4 locations are considered for the 2 base stations: indoor co-location (A), indoor medium spatial separation (B), indoor opposite location with larger separation (C) and outdoor location (D).
As conclusions for the previous analysis of spatial correlation, we can summarize the next main points:

- For the **indoor** scenarios that were measured, we have found that a **higher mean correlation** is obtained at the receiver than at the transmitter for the three BS locations. There are two main reasons for that: first of all, the transmitter antennas take advantage of the use of polarization diversity (cross-polarized antennas are used for antenna modules at BS), while the receiver antenna array is a single-polarized one. Secondly, the antenna elements spacing at the receiver is only 0.5 \( \lambda \), while for the transmitter the spacing for different antenna elements is 0.6\( \lambda \) for case A, and much larger for cases B and C with split antenna groups. As a conclusion, we see that even at the indoor scenarios where a high decorrelation due to the existence of many scatterers could be expected, an antenna module designed for low correlation is desirable for a conventional MIMO scheme (separated antenna elements, dual polarized antennas…).

- **Statistics for different transmitter locations were shown to be very different.** When the four transmitter antennas are placed very close to each other (cases A and D), their correlation coefficient follows approximately a Gaussian distribution, while for the cases
where antenna modules were split into two antenna groups or base stations, the CDF follows a non-gaussian distribution. Practically, that means that different correlation levels will be encountered, so studying different locations and its spatial correlation levels makes sense.

- **Case C**, where the two base stations are located further apart, offers the **lowest TX correlation** level, with a mean value for the spatial correlation coefficient that is 24 points lower than the case with highest correlation, case A (indoor co-located base stations).

- Although both **case A and case D** consider co-located base stations, the **latter outperforms the former one**. This is thanks to the higher flexibility in separating the antennas for the outdoor location (case D), and also to the better use of polarization diversity that is done in the outdoor-to-indoor scenario. However, before choosing an outdoor location for base station, other considerations (as required transmitted power, achievable capacity…) must be done, as we will see in next sections.

- The spatial correlation at the receiver follows a very similar distribution for the **four studied cases** (approximately Gaussian). However, even if the same routes were followed and the same receiver antenna was used, some differences are obtained (maximum difference of 9 points in mean value). This fact agrees with some other studies [ref] that show the deficiencies of the Kronecker model when assuming that spatial correlation at transmitter and receiver are separable.

- At the receiver, where antenna spacing is kept constant for all the cases, the slight differences among correlation levels as a function of studied case are mainly due to the different location of TX. The **lowest RX correlation is obtained for case D**, which again supports the idea of higher polarization diversity for the outdoor-to-indoor case. Moreover, the fewer LOS routes for case D also help in reducing the correlation level.

Knowing the spatial correlation at transmitter and receiver is useful to get an idea of available spatial diversity in the system. However, it may be of interest to consider the comparison of achievable capacity for each case. Moreover, it has been shown [252] that in some cases, such as keyholes [253], low correlation at transmit and receive ends does not involve a high capacity. In order to complete the analysis, next sections show path loss and received power as a function of location, eigenanalysis and capacity results for the measured scenarios.

**5.3.3 Path loss, fading and coverage analysis**

As it was presented in chapter 2, section 2.2.2, many different parameters are used when studying and characterizing the MIMO channel, such as the spatial and temporal correlation, angle spread, coherence time and Doppler spectrum, etc. One of the fundamental parameters for
any propagation channel (either SISO or MIMO) is the time-variant “gain” (or losses) of the channel. For modelling and analysis, this effect is usually broken down in two types of fading or changes in the channel gain: long-term fading (or slow fading) and short-term fading (or fast fading). The first one usually includes both the path loss and the shadow fading. Path loss is due to the power losses that the signal suffers when travelling along the distance to be covered between transmitter and receiver (usually free space path loss is assumed). The shadow fading are random slow variations of received power that are not directly dependent on the separation between TX and RX, and may be associated to the effect of physical shadowing caused by objects, walls, buildings, etc, which the signal finds when travelling along the propagation path. Finally, the fast fading is due to the (constructive or destructive) combination of multiple copies of the signal at the receiver, produced by multipath propagation.

The channel matrix that was estimated from the radio measurements contains the information of the gain (or losses) for the measured routes, as a function of RX location. Since multiple antennas are used at the transmitter and the receiver, several channel gains (in our case, $4 \times 4 = 16$ channel elements) are obtained. In order to represent the average channel response (path loss and fading), the received power averaged over the 16 channel elements was computed as:

$$PL(dB) = 10 \log_{10} \left( \frac{1}{N_{Tx} N_{Rs}} \|H\|_{Frob}^2 \right) = 10 \log_{10} \left( \frac{1}{N_{Tx} N_{Rs}} \sum_{i,j} |h_{ij}|^2 \right)$$  \hspace{1cm} (5-11)

where $\|A\|_{Frob}$ is the Frobenius norm of matrix $A$. This expression is the instantaneous path loss, that is, it includes both the slow fading terms and the fast fading effect.

As a first example, Figure 5-16 shows the path loss computed as (5-11), for one of the measured routes. An NLOS route following the northern hallway in the 5th floor was selected, with base stations co-located in an indoor position (case A).
We can see the two main types of fading in the signal: slow fading, which approximately follows a decrease in power with a log-distance dependency, and the fast fading representing multipath effects. In addition, some shadow fading seems to be also present, thus appearing some small fading not following the conventional power dependency with Tx-Rx distance.

In order to separate the fast fading and the slow fading, the first one can be filtered by a window, which is equivalent to average out the small scale or fast fading. The window size must be selected big enough to remove all the fast fading effect, but is must be small enough to avoid filtering the slow fading effect (free space slow fading and shadow fading). Several window sizes were tested, from 0.5 m to 5 m of measurement. After analysing the obtained path loss over several routes, the best window size for the studied indoor environment was chosen to be around 1m (100 samples, which is equivalent to 0.9 m), that is, approx. $6\lambda$. This is in a similar order to the proposed window size in other works, as in [254], to average out small scale fading without distorting large scale fading. Figure 5-17 shows the path loss that is obtained after removing the fast fading effect by filtering with the 1m-sliding window, for the same route as the one studied in Figure 5-16 (5th floor, NLOS route, northern corridor).

Even after averaging out the fast fading, we can see that the resulting curve is not a distance-log one, but other effect is added up: it is the shadow fading, due to objects that could appear between transmitter and receiver (in this case, the floor that the signal has to go through, for example).

This averaging operation by filtering with a sliding window was performed with all the measured routes and for the 4 studied BS location (cases A to D), so the slow fading could be analysed.
Figure 5-18 shows the path loss including slow and shadow fading as a function of location on the 4th floor for cases A-D. The values shown in Figure 5-18 are the average in each square. Note that for cases B and C, we actually see the average path loss from one mobile location to two different base stations. The average path loss for the whole floor was computed and its value is shown in Table 5-V.

<table>
<thead>
<tr>
<th>Path Loss (dB)</th>
<th>case A</th>
<th>case B</th>
<th>case C</th>
<th>case D</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>59.2</td>
<td>56.9</td>
<td>53.9</td>
<td>83.15</td>
</tr>
</tbody>
</table>

Table 5-V. Average path loss for all measured routes and the four BS locations (A to D)

Case A. BS co-located in NE corner.

Case B. BS located in different corridors but on the same side of building.

Case C. BS located in different corridors in opposite sides of building.

Case D. BS co-located on a different building to the NE.

Figure 5-18. Path loss $\sum |h_{ij}|^2 / N_T / N_R$, averaged over each 1m, for cases A-D with different locations for base stations.

The path loss plots indicate that for a total transmitted power, the power is better distributed for cases were the base stations are not co-located (cases B and C), which seems
reasonable. These cases provide more LoS situations (due to the propagation in the north and south hallways), and thus a better coverage. The outdoor location also offers a good distribution of power over the whole floor, but at the cost of higher losses, since the signal has to cross through the building walls, and the TX-RX distance is larger. Therefore, higher transmitted power will be required in the outdoor location if the same SNR level as in the indoor case is required.

5.3.4 Capacity analysis and different methods of combining base stations

Once the correlation and path loss analysis have been carried out, next step in characterizing the MIMO environment is to compute the achievable capacity and compare the results for different BS locations, for a single user (multiuser capacity or sum-capacity will be presented later). Firstly, some results regarding floor 4 are plotted, including capacity maps of this floor. Secondly, CDFs considering all the measured routes are presented.

To give a fair comparison between the studied scenarios, we have normalized the transmit power for the indoor cases A, B, and C so that the average SNR for each case over the whole floor 4 is 10 dB. The transmit power is thus:

\[ P_{\text{Norm}} = \frac{N_t B N_r \sigma^2 \text{SNR}}{E\|H\|^2_{Frob}} \]  

where each of the four cases is normalized with a different value, so H represented the measured channel matrix for each case (A, B, C and D) and four different values of \( P_{\text{norm}} \) are obtained. Since the path loss is much greater for the outdoor case, the normalization value for this case will be very different from the other ones, while \( P_{\text{norm}} \) is very similar for cases A, B and C. Thus, we still take into account the different path loss to different locations and the effect of more or less even geographical coverage. For each scenario the BS is assumed to transmit at full but fixed transmit power \( P_{\text{Norm,A}} \), \( P_{\text{Norm,B}} \), \( P_{\text{Norm,C}} \) or \( P_{\text{Norm,D}} \) regardless of the number of transmitter antennas. This leads to a variation in the received signal to noise ratio as the mobile moves along its trajectory. We then evaluate the capacity in b/s/Hz as [7]:

\[ C = \log_2 \left| I + HQH^H \right| \]  

where \( Q \) is the transmit covariance matrix such that \( \text{Tr}(Q) = P_{\text{Norm,\{A...D\}}} \). To compare cases A-D, we consider the following options of signal processing at a system level:

- **Option 1:** no information is shared between the two BSs and the mobile can only see one BS during the whole time. This will give the capacity for a 2x4 system. The BS has full channel state information (CSI) and allocates power to its antennas according to the water-filling scheme [255].

- **Option 2:** no information is shared between the BSs, but the MS makes a selection between the BSs based on the strongest received power. This will give the capacity for
a 2x4 system with BS selection. Both BSs are assumed to have full CSI and allocate power using water-filling scheme.

- **Option 3**: both BSs share all information and transmit power. Full CSI is assumed and water-filling over the full 4x4 channel is calculated. The total available transmit power is $P_{\text{Norm},[A..D]}$ and the system is not limited to use half the power on each BS.

Options 1 and 2 are reasonable to assume for all measurement cases, while option 3 is probably only reasonable when both BSs are closely located. However, the third option is still interesting since it will serve as an upper bound on the achievable capacity for this setup.

Let us first study the variation of the capacity with location for cases A-D. Figure 5-19 shows the local average capacity including path loss effect for the hallways and some of the offices on the 4th floor. Thus, these plots can be interpreted as coverage plots for the MIMO system. Starting with case A, we have a very high capacity close to the base station, but a poor coverage in the south hallway.

Case B, on the other hand, provides a more even coverage since we get propagation along both hallways. We also see that the option 2 (selection of 2 x 4) is practically equivalent to option 3, the full 4 x 4 system. The capacity drops below 10 b/s/Hz at approximately 20m in both cases. Only in the open area around the BSs (to the far right), where we can receive substantial power from both BSs, do we see a slight improvement using full water-filling. Taking into account that the full 4 x 4 system option requires that the 2 BSs share the CSI at any moment, the 2 x 4 system with BS selection seems a very interesting solution.

Next, looking at case C, we see that we have the same range of coverage, >10 b/s/Hz up to 20m, as case B. Also in this case option 2 (selection) and option 3 (full 4 x 4 water-filling) yield almost identical capacity results. Again the difference is seen only in the far ends (right or left), where water-filling provides 2-3 bits higher capacity.

For case D we have used a separate normalization as mentioned above, equivalent to using 24dB higher transmit power than the indoor cases. This gives a larger region that has capacity above 10 b/s/Hz, and we now cover the offices in the northern corridor. The coverage is more evenly distributed, but this is to the cost of higher transmit power.

Moving on to the capacity statistics, we first see in Figure 5-20 the CDF of the capacity for the whole floor for 2 x 4 systems with and without BS selection. The same power normalization is used as in Figure 5-19. The fixed systems represent single 2 x 4 systems with BS at the four different transmitter locations. Judging from the symmetry of the 4th floor, we might expect identical capacity for the indoor cases A, B, and C. However, our choice of routes combined with unavoidable changes in the propagation conditions from measurement to measurement result in the slight difference seen in Figure 5-20.
a) Case A, Option 3. Capacity on a $4 \times 4$ system using water filling over all channels

b) Case B, Option 2. Capacity on a $2 \times 4$ system using water filling at the BS. The Ms selects the BS from which it receives the strongest power.

c) Case B, Option 3. Capacity on a $4 \times 4$ system using water filling over all channels

d) Case C, Option 2. Capacity on a $2 \times 4$ system using water filling at the BS. The Ms selects the BS from which it receives the strongest power.

e) Case C, Option 3. Capacity on a $4 \times 4$ system using water filling over all channels

f) Case D, Option 3. Capacity on a $4 \times 4$ system using water filling over all channels

Figure 5-19. Capacity maps including path loss in the H matrices. The transmit power is chosen so that the average SNR = 10 dB for the whole floor, in the four studied cases.
Figure 5-20. Capacity CDF for Option 1 (a fixed 2 × 4 system) and Option 2 (selection between two 2 × 4 systems).

Case D, compensated with a 24 dB higher power, is clearly superior. However with BS selection case C is superior. This is due to the better power distribution over the whole floor and macro diversity gains. For case A and D there is only a slight improvement when using BS selection, due to a limited spatial diversity.

Next, Figure 5-21 shows the CDF of the capacity for Option 3, a full 4 × 4 system, and Option 2, selection between 2 × 4 systems. Cases B-C show no improvement using a full 4 × 4 system, indicating that the system will be making a selection of a 2 × 4 system. For cases A and D, we have a slight improvement attributed to beamforming gain and some spatial diversity (c.f. Figure 5-13, Figure 5-14). The most important conclusion from Figure 5-22, however, is that 2 × 4 selection in cases B and C vastly outperform the full 4 × 4 system of case A. Thus, for indoor base stations we are much better off using separate 2 × 4 systems and simple selection compared to a single 4 × 4 system. The reason is both lower average path loss (c.f. Table 5-V) and a lack of spatial gain due to the hallway propagation, see below.

Finally, we have studied the capacity for a fixed local average SNR to see how ideal our MIMO channel is. Figure 5-22 shows the CDF of the capacity with local SNR = 10 dB averaged over a 1 m distance. We see both the case of water-filling assuming perfect CSI, and the case of no CSI at the transmitter. The result shows that the outdoor case D provides the highest degree of multipath. With a fixed SNR neglecting path loss, case A will also outperform case B and C since the latter will have quite un-equal eigen values due to different path loss to the two base stations. Comparing with e.g. the mean capacity of 10.9 b/s/Hz for a 4 × 4 i.i.d. channel with no CSI, it is also clear that our channel is non-ideal.
Figure 5-21. Capacity CDF for Option 3 (full 4 × 4 system) and Option 2 (selection between two 2 × 4 systems).

Figure 5-22. Comparison of water-filling (Full CSI at TX) and no CSI at TX for a 4x4 MIMO case at a local average SNR = 10dB.
5.4 Multi-user-oriented evaluation of measurements and analysis of results

After thoroughly analysing the main characteristics of the MIMO channel and the achievable capacity of the system for a single user case, it is of great interest to study the effect of intra-cell and inter-cell interference for indoor and indoor-to-outdoor scenarios. Real scenarios will have to cope with this type of interferences, which have been previously studied and theoretically characterized, but hardly analysed from measurements in real scenarios.

We show here results in terms of MIMO capacity for single users when interferences are present, and also in terms of MIMO sum-capacity. Since exactly the same routes were measured for each TX location, we may assume that multiple users are present in a scenario and that the received signal is the sum of the transmitted signals from different TXs through different measured channel matrices. Therefore, the extrapolation to multi-user system is straightforward. We may note that channel matrices were measured with one one TX location at each time, thus avoiding interferences that would have distorted channel $H$ estimation.

5.4.1 Multi-user MIMO system model and sum capacity problem

We summarize here some of the most important characteristics of multi-user MIMO systems from a mathematical point of view, for both uplink and downlink. We may note that for this study we consider that the system operates in the same bandwidth for uplink and downlink, and thus both channels are equivalent, that is, $H_{UL} = H_{DL}$. For simplicity reasons, we will skip the subscripts $(\cdot)_{UL}$ and $(\cdot)_{DL}$, being clear from context. Next section will show results in terms of the parameters presented here.

5.4.1.1 Multi-user MIMO downlink

The downlink MIMO channel is usually referred to as MIMO broadcast channel (MIMO-BC), since the same signal is transmitted to multiple users. Let us consider a MIMO downlink system, where $K$ users with multiple-element antennas (at RX) are present. The received signal vector at the receive antennas of user $k$ ($k=1..K$) can be written as:

$$y_k = H_k \left( x_k + \sum_{i=1, i \neq k}^{K} x_i \right) + n_k = H_k x_k + \left( \sum_{i=1, i \neq k}^{K} H_i x_i + n_i \right) = H_k x_k + z_k \quad (5-14)$$

where $H_k$ is the $N_{Rx,k} \times N_{Tx}$ channel matrix between user $k$ and the BS that is serving this user $k$, being $N_{Rx,k}$ the number of antennas at the user $k$, and $N_{Tx}$ the number of antennas at the BS. $x_k$ is the vector of transmitted signal for user $k$, so that the transmit covariance matrix is $R_{x_k} = E[x_k x_k^H]$. The vector $n_k$ represents both the noise and the intercell interference. Thus, the term $z_k$ represents the total interference (both intercell and intracell) for user $k$ plus the noise. Intracell interference is due to the signal sent from the BS to other users, while intercell
interference is caused by BSs located in close cells with frequency reuse. In order to simplify the analysis, the system is sometimes simplified so that $z_k$ is Gaussian distributed. However, this is not true in a general case (for example, it does not hold for cases with few interfering users), since the interferences are not spatially white, but highly depend on the location of interferences and the user. We will study the capacity obtained from measured $H$ matrices, so the real interference is computed.

In order to characterize the system and maximize its performance, several parameters can be considered to be optimized. Some of them are the quality of service or SINR in the system (as discussed in [256], [257]). However, the most popular parameter usually studied is the capacity of the MIMO system. For a multi-user system, the parameter to be optimized is the sum capacity (addition of capacity for each user), with a certain constrain (usually limited transmitted power). The optimization problem is thus expressed as [247]:

$$C = \max_{R_{z_k z_k}} \sum_{k=1}^{K} \log_2 \frac{\det(R_{z_k z_k} + H_k R_{n_k n_k} H_k^H)}{\det(R_{x_k x_k})} \quad (5-15)$$

subject to power constraint $\sum_{k=1}^{K} \text{trace}(R_{n_k n_k}) \leq P_{\text{max,BS}}$

The covariance matrix $R_{z_k z_k}$ represents the covariance of the interference (both intracell and intercell) plus noise signal, and is given by:

$$R_{z_k z_k} = R_{n_k n_k} + \sum_{i=1, i \neq k}^{K} H_i R_{n_k n_k} H_i^H \quad (5-16)$$

As in single-user MIMO problem, several cases regarding channel knowledge at TX may be considered. The two obvious cases are perfect channel state information (CSI) at TX and no CSI at TX.

When there is perfect CSI at TX, the joint optimization of capacity for downlink channel is a quite complex problem. Some possible algorithms have been proposed to this problem, both non-linear and linear. A non-linear (and thus quite computationally complex) solution is the dirty-paper precoding [258]. A linear precoding technique is block diagonalization [259], with its main disadvantage being that the total number of RX antennas (adding all users) must be lower than the total number or TX antennas, which makes it not suitable for the case studied here, as we will present later. Some works, as [260], present suboptimal solutions for higher number of RX antennas. However, we will focus here on the uplink system due to its simpler solution, and we will limit the downlink results to the ones with no CSI at TX, as detailed in next section.
5.4.1.2 Multi-user MIMO uplink

In the uplink MIMO multi-user system, several users transmit to the same BS, while each user sees a different channel matrix $H_k$. This channel is usually known as multi-access channel (MIMO-MAC). The signal received by the base station can be written as:

$$y = \sum_{k=1}^{K} H_k x_k + n = H x + n \quad (5-17)$$

where the matrix $H$ represents all the channel matrices for the $K$ users, and $x$ is the transmitted signal for the $K$ users, that is:

$$H = \begin{bmatrix} H_1 & H_2 & \ldots & H_K \end{bmatrix}$$

$$x = \begin{bmatrix} x_1^T & x_2^T & \ldots & x_K^T \end{bmatrix}^T \quad (5-18)$$

As in the downlink case, $n$ is the vector that includes noise and intercell interference. We observe that, contrary to the MIMO-BC, the MIMO-MAC may be viewed as a large MIMO system represented by the MIMO channel matrix $H$, of dimensions $N_{BS} \times \sum N_{\text{user},k}$. However, since the users are independent, we must assume that signals $x_k$ from different users are uncorrelated. Therefore, $R_{xx} = E\{x, x^H\}$ is block-diagonal. The sum capacity for this case is then written as [261]:

$$C_{\text{sum}} = \max_{R_{xx} \geq 0} \log \frac{\det \left( R_{xx} + \sum_{k=1}^{K} H_k R_{x_k x_k} H_k^H \right)}{\det (R_{xx})}$$

subject to power constraint $\text{trace}(R_{x_k x_k}) \leq P_{\text{max},k}$, $P_{\text{max},k}$ being the maximum transmit power for the $k^{th}$ user.

As in the downlink case, two main examples are the most representative regarding CSI assumption: perfect CSI at TX and no CSI at TX, giving each one different capacity solutions. We may note here two issues: firstly, in order to assume perfect CSI at TX, channel information must be sent through a feedback channel to the users. Secondly, thanks to the fact that the BS knows all $H_k$ matrices and sent signals $x_k$, this information can be used to perform successive interference cancellation (SIC) at the receiver, so that user 1 deals with interference from all other users, but user 2 suffers only interference from users 3 to $K$, etc.

Both perfect and no CSI at TX for uplink have been evaluated. The analysed TX-RX schemes are presented in next section, together with the studied scenarios and the capacity expressions and results for each one. We have divided the results in downlink scenario and uplink scenario, since expressions, studied scenarios and results are different for the two of them.
5.4.2 Evaluation of multi-user MIMO at downlink: studied scenarios and capacity analysis

By using the measured $H$ matrix realization presented in section 5.2.3, the effect of multi-user interference in the capacity and the overall throughput of the system has been analysed for downlink. Thanks to the large number of measured points and to having measured exactly the same routes for all the TX locations, and assuming negligible changes due to small changes in the scenario for the narrowband measurements, we can assume that different $H$ matrices realizations are equivalent to having different simultaneous users or base stations. First of all, downlink studied cases are presented in this section. Uplink evaluation will be presented in next section.

5.4.2.1 Downlink scenario and capacity expressions

For downlink scenario, we focus on the effect of intercell interference and compare the achievable capacity for different BS locations. To do so, we assume that each user in a cell will use different frequency (FDMA) or time (TDMA) from the other co-cell users, so no intracell interference has to be studied (this will be studied later for uplink scenario). The frequency reuse in the cells is 1, so intercell interference is significant. The aim here is to study the effect of intercell interference in an indoor scenario when the interference comes from a base station located outdoor, for different interference levels and different indoor base station locations. This is important for design of system deployment.

In our measured environment, we consider the outdoor BS location as an interfering BS that is serving a cell different from the one where the user is located (“indoor cell”), but which still interferes significantly. Three possible locations for the indoor BS are studied: cases 1, 2 and 3 in Figure 5-11. Non-ideal power control is assumed for the indoor cell: the power transmitted by the indoor BS is adjusted so that the power received by user and averaged over a 1-second window is kept constant (as in [5]). The power transmitted by the outdoor BS is assumed to be fixed, so that the average interference level $I$ yields to a selected average signal to interference ratio $\overline{SIR}$ (for evaluation purposes). Figure 5-23 depicts one of the studied scenarios for downlink.

![Figure 5-23. Scenario for downlink MU-MIMO with intercell interference, for BS location case 1.](image-url)
For this scenario where there is no intracell interference and there is only one user per cell, the problem to be optimized is reduced to the following capacity problem:

$$C = \max_{\mathbf{A}} \log_2 \frac{\det(\mathbf{R}_m + \mathbf{H R}_x \mathbf{H}^H)}{\det(\mathbf{R}_m)}$$

subject to power constraint $\text{trace}(\mathbf{R}_m) \leq P_{\text{max,BS}}$

where $\mathbf{R}_x$ is the covariance TX matrix and $\mathbf{R}_z$ is the covariance matrix of the signal that describes the intercell interference from outdoor BS and the noise:

$$\mathbf{R}_z = \sigma^2 \mathbf{I}_{N_{tx}} + \mathbf{H}_{x} \mathbf{R}_{x,x} \mathbf{H}_{x}^H$$

(5-21)

The matrix between the outdoor BS and the user is $\mathbf{H}_{i}$, the covariance matrix of the signals transmitted by the outdoor BS is $\mathbf{R}_{x,x}$ and the noise power is $\sigma^2$.

When there is no CSI at TX (the BSs do not know about the downlink channel), it is known that the optimum solution is transmitting multiple independent complex Gaussian streams with equal power allocation. In this case, the capacity of the system is given by:

$$C = \log_2 \frac{\det\left(\sigma^2 \mathbf{I}_{N_{tx}} + \frac{P_t}{N_{tx,j}} \mathbf{H}_{j} \mathbf{H}_{j}^H + \frac{P_{\text{indoor}}}{N_{tx,\text{indoor}}} \mathbf{H}_{x} \mathbf{H}_{x}^H\right)}{\det\left(\sigma^2 \mathbf{I}_{N_{tx}} + \frac{P_t}{N_{tx,j}} \mathbf{H}_{j} \mathbf{H}_{j}^H\right)}$$

(5-22)

$P_t$ is the total interference power, while $P_{\text{indoor}}$ is the total transmitted power from the indoor BS.

When CSI at TX is assumed, more optimal solutions may be used (as iterative waterfilling). However, that would mean that indoor and outdoor BS should cooperate to reduce interference, thus feedback from one BS to another is needed. This case is not analyzed here. An intermediate step may be to use waterfilling for the indoor link, thus optimizing the user capacity (although not reducing interference).

### 5.4.2.2 Results for multi-user MIMO downlink

For downlink case, the capacity for the system where channel intercell interference is present has been studied. Two main parameters have to be selected: noise level, by means of selecting an SNR value, and interference level, by means of selecting an SIR value. No CSI is assumed for all results here.

Three representative cases are presented (other examples are not included for space reasons):

- High noise scenario (SNR = 2 dB). The interference is chosen so that $\text{SIR} = 10$ dB.
- Medium noise and interference scenario: we chose $\text{SNR} = \text{SIR} = 10$ dB. A noise level so that SNR=10dB is quite reasonable for a communication system. An interference of $\text{SIR} = 10$ dB may be considered a high value for intercell interference.

224
- Low noise and interference scenario: \( \text{SNR} = \frac{\text{SIR}}{30} \text{dB} \) are chosen, thus emulating a “good” channel.

The CDF of the capacity was computed using eq. ( 5-22 ), taking into account the 22 measured routes that covered the user for each BS location. The results are shown in Figure 5-24, Figure 5-25 and Figure 5-26, where we have depicted both the capacity when the intercell interference is considered (“MU” label), and the capacity for the single-user MIMO channel (“SU label”), for comparison purposes.

For the BS location of case B (two groups of Base Stations, with 2 antennas each, at opposite corners of the building), we have included two options, similarly to what was studied for single-user in previous section: either the two BSs groups (at East end and West end of the corridors, respectively) act as a single BS so the receiver estimates the four channels (which is equivalent to a \( 4 \times 4 \) MIMO system), or we assume that the user selects the TX group from which he receives the higher power, thus converting the system in a \( 2 \times 4 \) one. This way the user receiver is simplified since the required channel estimations are reduced. The first one (\( 4 \times 4 \) case) has been called case B1, while the \( 2 \times 4 \) one has been called case B2.

From these figures, we observe that for all the cases, the best BS location in terms of capacity is case A (CoLocated elements). This is reasonable if we take into account that power control is assumed, thus giving a higher importance to the location that offers higher diversity. Since there are less line-of-sight routes for this configuration, it may be expected a higher capacity with normalized channel. Also reasonable is the fact that the \( 2 \times 4 \) system offers the lowest capacity.

---

**Figure 5-24.** CDF of capacity for a multiuser MIMO system with two BSs, low SNR case (\( \text{SNR}=2\text{dB}, \text{SIR}=10\text{dB} \)).
For the studied downlink scenario, we observe that when the noise effect is very important, the interference effect becomes less important. Thus, for the high-noise scenario (Figure 5-24), including the interference reduces capacity around 0.5 b/s/Hz for all BS locations, while the same interference level causes an effect of reduction of capacity in 2.2b/s/Hz (please note different scaling at x axis for the medium-noise scenario (Figure 5-25).

We also see that cases of 4×4 spatially separated antennas (cases B1 and C) give very similar results, for low or high SNR and also for different interference levels. A slightly higher capacity (increase of around 1 b/s/Hz) is obtained with case B1 (TX groups at opposite corners of the building), only noticeable for the case of high SNR (Figure 5-26).
It is of interest to see that the capacity obtained with all the BS locations for the studied 4×4 system (cases A, B1, and C) is reduced in a very similar value when including intercell interference. The studied 2×4 system with BS selection shows lower reduction for the high SNR case.

Finally, we mention that the reduction of capacity obtained due to the introduction of intercell interference is more important for high SNR scenarios, as expected.

5.4.3 Evaluation of multi-user MIMO at uplink: studied scenarios and capacity analysis

After evaluating the capacity of a multi-user indoor system when outdoor-to-indoor interferences (intercell interferences) are included, we aim at analyzing the effect of intercell interferences. With this objective, we have focus on the uplink scenario and have compared several indoor and outdoor locations for base stations. The interferences are considered at the corresponding base station, and they are computed from some of the measured routes, as explained below.

5.4.3.1 Uplink scenario and capacity expressions

For uplink scenario, we focus on the intracell interference. This way, we can study whether an indoor location is better than an outdoor location for BS, when multiple users are included in the system. These results are valuable to complement the ones presented in the single-user analysis.

In our measurement environment, we consider now one BS each time and 2 routes as users in the system. In order to get a generalized result regarding the studied indoor scenario, 50 iterations are done for each BS location, where the routes of the two users are randomly selected for each iteration from the 22 measured routes (but kept the same when evaluating different BS locations, for comparison purposes). As explained before, each route is approximately 1-minute long. Identical routes for both users are avoided when randomly selecting the routes. Figure 5-27 shows one example for this scenario.

![Figure 5-27. Example of case for uplink MU-MIMO with co-channel interference, for BS location case 1.](image-url)
In this uplink scenario, we consider the same realistic power control (thus channel normalization). The interference level is directly given by this power control. We may note that since this is intracell interference, its value is expected to be higher than the intercell interference studied in downlink.

Two possibilities are considered:

1. **No CSI is available at TX.** That means that there is no feedback channel from the BS. The best the users can do is using equal power allocation and Gaussian transmitted vectors. The capacity for one user is given by eq. (5-22). The sum-capacity in this case is given by:

\[
C = \log_2 \left( \frac{\text{det} \left( \sigma^2 I_{N} + \frac{P_1}{N_{T,1}} H_1 H_1^H + \frac{P_2}{N_{T,2}} H_2 H_2^H \right)}{\text{det}(\sigma^2 I_{N})} \right) \tag{5-23}
\]

2. **Perfect CSI is available at TX.** For this case, it has been shown [262] that the maximum sum-capacity is obtained by using an iterative waterfilling solution, where the optimal \( R_{xx} \) are computed as:

initialize \( R_{xx,i} = 0, k=1..K \)

repeat

for \( k = 1 \) to \( K \)

\[
R'_{z_i z_i} = R_{mn} + \sum_{j=1}^{K} H_j R_{x_j x_j} H_j^H
\]

\[
R_{x_j x_j} = \arg \max_{\forall k} \sum_{k=1}^{K} \log_2 \left( \frac{\text{det} \left( R'_{z_i z_i} + H_k R_{x_j x_j} H_k^H \right)}{\text{det}(R'_{z_i z_i})} \right)
\]

subject to power constraint \( \text{trace}(R_{xx}) \leq P_{\text{max,BS}} \)

until desired accuracy reached

Each problem \( k \) in eq. (5-24) is equivalent to solving a conventional single-user problem, whose solution is given by the conventional water-filling problem [263]. The final sum-capacity is computed from eq. (5-19) with the optimal \( R_{xx} \).

### 5.4.3.2 Results for multi-user MIMO uplink

As previously stated, results for uplink has been focused on the effect of co-channel (intracell) interference and its effect on each of the four possible locations for BSs. SNR = 10 dB is assumed is all cases, and the interference level is computed so all users transmit the same average power (as previously explained).

First of all, the capacity for one user when there is another user as co-channel interference is obtained. It is important to note that this is not the sum-capacity of the system,
but the capacity available for a single user. We assume no CSI at TX, so uniform power allocation is performed. Figure 5-28 shows the obtained CDF of the obtained capacity.

As shown in Figure 5-28, when no interference is included (“SU” labels), the best BS location is the outdoor one (case D), which again agrees with the case where non line-of-site routes are, and thus a higher spatial diversity is available for a normalized channel. The indoor BS locations behave similarly to the downlink case for no interference, so cases B1 and C perform very similarly, and the lowest capacity is obtained with the 2×4 system.

However, when the interfering intracell user is included (“MU” labels), interesting results arise. According to Figure 5-28, the BS scheme which achieves highest capacity is the 2×4 BS selection one. This can be explained by observing that for this case 2 BS are available: the interferer may be served by the BS that is not serving the user. This situation will occur when the interferer is receiving more power from the “free” BS. The average power transmitted by the interferer that will be received by the BS serving the desired user will be lower in this case. Thus, the average interference level for the 2×4 system is much lower than for the other cases, giving a higher capacity for the same SNR. We may note that capacity for a single user is computed here.

When sum-capacity is computed, we obtain Figure 5-29 for SNR=10dB. Both the optimal case for no CSI at TX and perfect CSI at TX are computed. We observe that the capacity increases significantly when the iterative water-filling scheme is used, clearly outperforming the uniform power allocation system. The BS locations that imply two groups of TX have a less Gaussian distribution for the capacity, although no big differences are found for average capacities.
Figure 5-29. CDF of sum-capacity for uplink MIMO. 2 users in the system. SNR=10dB.
5.5 General conclusions, contributions and further research on propagation aspects for MIMO systems

We have presented a new channel measurement campaign that has been performed aiming at studying some open issues in MIMO systems. The campaign has covered an extensive indoor area with multiple routes, floors and offices, and the measured channel matrices have been used to analyse several aspects in the MIMO channel and system that are considered as not fully characterized yet.

Regarding the characterization of MIMO channel, we have compared the spatial and polarization diversity available in these channels when indoor locations are considered for the base station (transmitters), with the diversity obtained when the base station is located outdoors. Since the multi-antenna receiver is moved along indoor routes, the first case involves having a fully indoor scenario, while the second one consists in an outdoor-to-indoor scenario. The analysis of correlation between transmit antennas has shown that a higher spatial diversity is obtained for the outdoor-to-indoor environment than for the indoor environment, when a single 4-element BS is used. This has been found to be due to two reasons: different LOS situations and polarization diversity in both scenarios. LOS routes involve an increase of signals correlation, and they are more likely to happen in indoor than in outdoor-to-indoor scenarios. Moreover, an interesting reduction in TX antenna correlation has been found when polarization diversity is used in outdoor-to-indoor scenarios, while it is much smaller in indoor scenarios. We have attributed this behaviour to the depolarization that the signal suffers when traversing building walls in outdoor-to-indoor scenario. As a result, dual-polarized antennas are definitely recommended for outdoor base stations in MIMO systems. Some reduction in TX correlation is also found in indoor scenarios when different polarizations are used, but it is not as significant as in the outdoor-to-indoor case. Thus, dual-polarized antennas can also be interesting for indoor scenarios, but their advantages are not as patent as in outdoor-to-indoor cases.

To compare other options for location of BS’s, we have also analysed MIMO channel characteristics when splitting TX modules into two groups with two antennas each. The correlation analysis has shown that a reduction in antenna correlation is achieved, as expected. Moreover, if the two base stations are properly located, they can cover different areas, thus giving a better coverage in terms of average received power. Thus, the option of using spatially separated antennas (as multiple base stations that cooperate in an optimal way) in the indoor scenario to be covered is also of great interest. Nevertheless, the main drawback of this solution is that cooperation between the base stations is needed, and if a full 4×4 MIMO system is desired where CSI is used at the transmitter this information must be shared between BS’s, which may be unfeasible for real scenarios.

In order to overcome the latter problem, we have proposed a scheme where a 2×4 MIMO system is considered, instead of the optimal full 4×4 MIMO system. By computing the
available capacity when a CSI algorithm is used, we have shown that a 2×4 scheme where the
BS (Tx group) is selected by the terminal (RX) with a simple algorithm of highest received
power reaches almost the same performance as the full 4×4 MIMO system, and adding a
significant reduction in system complexity. Other results are also present, as the reduction in
capacity when comparing our real-channel measurements with ideal i.i.d. channel.

Finally, by using measured MIMO channel realizations, we have analysed different base
station locations when intercell and intracell interference is present, for both uplink and
downlink, and cases with no CSI and full CSI at TX. The study shows that both indoor and
outdoor locations with close antenna elements performs very similar when CSI at the transmitter
and power control is assumed, being slightly higher the achieved capacity for indoor location.
Spatially separated locations for BS shows slightly lower average capacity and non-Gaussian
distribution for the measured environment.

Contributions

The contributions of this chapter of the thesis to the current state of the art can be
summarized as follows:

✓ Novel results on the characterization of outdoor-to-indoor MIMO scenarios have been
presented and compared with results in similar indoor scenarios. A complete
measurement campaign was undertaken to acquire data records that were used to
estimate the MIMO channel matrix for many different locations of TX and RX, thus
obtaining a great number of $\mathbf{H}$ matrices. Specifically, we have analysed the main
differences between outdoor-to-indoor and indoor scenarios in terms of antenna
correlation, coverage (path loss) and available capacity, showing that the outdoor-to-
indoor scenario outperforms the indoor scenario regarding spatial and polarization
diversity at the cost of a higher path loss (and thus requiring higher transmitted power).

✓ Several locations for the MIMO base station have been measured and evaluated,
assuming an indoor scenario for the receiver. This issue has hardly been considered in
previous works, but we think it has a great importance when deploying a MIMO
system. Not only outdoor and indoor BS locations have been considered in this study,
but also the use of two cooperative BS with lower number of antennas and located at
indoor but spatially separated locations. The available capacity for each BS location has
been computed for two different assumptions: no CSI at TX and perfect CSI at TX,
showing that outdoor BS location offers the highest capacity in both cases when power
control is assumed. We have also proposed another method to normalize the channel
matrix, so that the propagation characteristics are better included in the H matrix:
assuming fixed transmitted power so that the average received SNR is equal for all BS
locations (in order to fairly compare them). With this assumption, we have shown that
the BS locations that achieve better coverage can also offer a higher capacity. With this same assumption, a very interesting result is that by using a simple BS selection method based on the received power, we have shown that a $2 \times 4$ MIMO system with CSI at TX can obtain almost the same capacity as a full $4 \times 4$ MIMO system with CSI at TX, but significantly reducing the complexity of the system since in the $2 \times 4$ system with BS selection CSI does not have to be shared by spatially separated BS’s (as it has in the $4 \times 4$ system).

Also the multiuser case has been evaluated for different BS locations and outdoor-indoor scenarios. We have computed capacity for one and multiple users when both intracell and intercell interference is included, showing the degradation due to interference. All the results were calculated by using the measured H matrices, thus adding to the study the important fact that a real channel was used. Among the presented results we highlight the significant reduction in the effect of interference that is obtained when the iterative water-filling scheme is used.

**Further research**

Some of the open issues regarding MIMO channel characterization and propagation that were mentioned at the beginning of this chapter have been addressed along this thesis. However, there are still some questions that were not treated (or were superficially treated) here and are still of importance. We emphasize here the most significant ones.

Our work has focused on indoor and outdoor-to-indoor environment. However, if MIMO systems are to be widely deployed in cellular systems as 3G-WCDMA or WiMax outdoor characterization for MIMO is also required. Although some initial works have recently been presented (as [264], [63]), more measurement campaigns and evaluation of results would be desirable, for example measurements with multiple MIMO BS’s in outdoor locations and covering outdoor areas.

The topic of MIMO characterization in a specifically propagation sense (multipath and cluster analysis, direction of arrival and departure for multipath signals, etc…) is a very active one nowadays. Many new measurements have recently been presented obtained with very accurate and broadband MIMO channel sounders aiming at characterizing different channel parameters as the joint distribution of DoA and DoD with time delay, and also at evaluating how clusters of scatterers can be modelled. Many further research is still going on here.

Finally, it is worth mentioning that from an operator point of view it is also interesting to know how parameters of the BS may affect the MIMO system performance. Although some have been studied in this thesis (as location and polarization) others are still open issues, as numerical analysis of the effect of spacing among antennas as a function of the environment.
ANTENNA ASPECTS OF MIMO SYSTEMS:
Design, implementation of antenna array and evaluation methods for MIMO

6.1 Design of a compact antenna array for MIMO systems: a solution for realistic user terminals ........................................................................................................ 238
6.1.1 Design of the antenna element ....................................................... 238
   6.1.1.1 Basics on PIFA design.................................................................. 239
   6.1.1.2 Initial antenna element design..................................................... 241
   6.1.1.3 Final element characteristics ..................................................... 246
6.1.2 Antenna array: low mutual coupling and compact design .......... 249
6.1.3 Final array characteristics and prototype implementation........ 252
6.1.4 Reference antenna array ............................................................... 258
6.1.5 Conclusions .................................................................................. 262

6.2 Evaluation of antenna arrays for MIMO ....................................... 264
6.2.1 Evaluation using measured radiation patterns............................... 264
   6.2.1.1 MIMO capacity using the Kronecker model and the antenna covariance ................................................................. 264
   6.2.1.2 MIMO capacity using a standardized channel model and measured radiation patterns ....................................................... 272
6.2.2 Evaluation using measured radio channel responses.................... 281
6.2.3 Evaluation with a reverberation chamber .................................... 289
6.2.4 Comparison of methods for evaluation of MIMO antenna arrays and conclusions ........................................................................................................ 292

6.3 General conclusions, contributions and further research on antenna aspects for MIMO systems ................................................................. 299
The antenna module is one of the main aspects to be considered in the design of a multi-antenna system. Despite this, so far little attention has been given to the study of realistic antenna arrays and to the analysis and optimization of antenna parameters (such as configuration for compact designs or reduction of coupling effects). Moreover, the best way to evaluate the antenna performance for a MIMO system is an open issue.

In this chapter, a novel antenna array for PDA-type user terminal is proposed. After design and implementation, different methods are proposed and compared to evaluate its performance. In order to contrast the results, also a reference array is used to compare the performance with the proposed antenna array for the different used methods.
6.1 Design of a compact antenna array for MIMO systems: a solution for realistic user terminals.

Most previous works on MIMO systems neglect or pay little attention to the effect of the antenna and its radiation characteristics in the MIMO system. From an information theory and algorithmic point of view, the antenna has usually been simulated as an ideal sensor. However, in a real system the antenna plays an important role and it is interesting to optimally design the antenna so the best performance can be attained.

One of the problems to bring MIMO techniques into real systems is the difficulty to locate several antenna elements in the user terminal, which is usually constrained in size and weight. As a result, designs with small antennas are preferred, and few elements are usually considered. Up to now, two to four elements have been considered as a fair number of elements for the user terminal in MIMO next generation communications.

When the elements are very closely located, the coupling between them increases, thus deteriorating the matching and reducing the antenna efficiency. Moreover, the spatial diversity is also reduced. Hence, some differences between the elements in the array, such as radiation pattern diversity and polarization diversity may be of interest. Also, a design to minimize coupling is desirable.

A novel antenna array has been designed for MIMO applications. Although the main objective here was not to present a thorough study in designing antennas for MIMO, but instead to examine and to propose different methods to evaluate them, an effort has been made to design a realistic antenna module, in contrast with conventional wire or patch antennas that have been used in other measurement campaigns.

6.1.1 Design of the antenna element

Some previous works have addressed the design of small antennas for user terminals and handheld devices. Printed planar antennas that can be integrated into the handset are good candidates, due to their advantages by comparing with the conventional external antennas (monopoles or helix antennas), such as less easily broken off, less sensitivity to the geometry of the handset and possibly less interaction with the user.

A conventional type of printed antenna is the microstrip patch antenna, which has been widely studied for several years and many applications [265]. However, the relative large size of this type of antenna (typically a length close to \(\lambda/2\)) makes it not the best option for our application. In order to reduce the size of the resonator, the patch antenna can be varied to the so called planar inverted-F antenna (PIFA) [266][267].

Based on the concept of the image principle, the PIFA consists of a ground plane, a top patch, a feed wire and a shorting mechanism that short-circuits the top patch to the ground plane. This way the patch is converted to a quarter-wave resonator or even multiple resonant
modes, leading to a reduction of patch size of at least 50% by comparing to the conventional half-wave patch. Different shorting mechanisms can be used, being the most commonly used the shorting walls and the shorting pins.

Other examples of small antennas are fractal antennas [268], patch antennas with optimized geometry using slot and slit loading and more complex microstrip antennas, such as the slot loaded bow-tie microstrip patch, the stacked shorted patch antenna or the planar tapered slot antenna [269]. Due to its small size and simple structure, the PIFA was chosen as antenna element for the MIMO array to be studied in this work.

6.1.1.1 Basics on PIFA design.

Some basic concepts on PIFA design are given here, since they were used in the design of the antenna element. However, please note that a detailed theoretical analysis is out of the scope of this thesis.

The general structure of a PIFA is shown in Figure 6-1. The main parameters to be adjusted in the design are the patch dimensions $L_1$ and $L_2$, the structure height $h$, the shorting wall width $w_s$ and the distance from the probe to the shorting wall, $s$. When $w_s \rightarrow 0$, a shorting pin instead of a shorting wall is used. In the figure below, a dielectric substrate, such as air or another dielectric material, acts as substrate that separates the shorted patch from the ground plane.

![Figure 6-1. General structure of a PIFA.](image-url)
The patch dimensions $L_1$ and $L_2$, together with the height $h$ are chosen to obtain a structure that resonates at a certain frequency, $f_r$. The resonant frequency can be roughly estimated by simple equations, depending on the type of shorting technique. When $w=L_1$ (whole shorting wall), the resonant condition is:

$$L_2 + h \cong \frac{\lambda}{4}$$

(6-1)

where $\lambda$ is the wavelength for the resonant frequency, that is, $c = f_r \lambda$. Please note that here the fringing effect has not been considered, which would slightly reduce the actual size of the patch for obtaining a certain resonant frequency. If the wall is reduced to a shorting pin ($w\rightarrow0$) located close to the corner, the electromagnetic waves will travel a longer path, and then the resonant condition becomes:

$$L_1 + L_2 + h = \frac{\lambda}{4}$$

(6-2)

As in the case of conventional patches, the bandwidth of the PIFA increases with increasing height, $h$.

Regarding the shorting structure, its width has an impact in both the resonant frequency and the bandwidth. As the shorting wall is narrowed, the inductive loading of the antenna element increases, and hence the resonant frequency decreases. As a result, the size of the antenna can be reduced for a certain working frequency by reducing the shorting wall width. However, a drawback is that the relative bandwidth is reduced. Therefore, a trade-off between size and bandwidth should be achieved.

In order to simplify the implementation, the patch is usually etched on a dielectric material with good mechanical features (rigid, easy to be used for etching microstrip structures…). Also air may be used as substrate, but then the structure may be not robust enough and some construction difficulties may be solved. Including the extra substrate, the PIFA structure is modified and two dielectric layers should be considered in the design, as in Figure 6-2.

![Figure 6-2. Front view of PIFA with two substrates.](image)

Some studies [270] have shown that an equivalent relative dielectric constant $\varepsilon_r$ of a structure with a patch, two layers of substrate and a ground plane can be computed as,
\[ \varepsilon_{r,\text{eff}} = \frac{\varepsilon_{r_1} \cdot \varepsilon_{r_2} (h_1 + h_2)}{\varepsilon_{r_1} \cdot h_2 + \varepsilon_{r_2} \cdot h_1} \]  

(6-3)

Therefore, when computing the effective wavelength of the upper patch in the structure shown in Figure 6-2 the two dielectric materials should be considered, as in equation (6-4).

\[ \lambda_{\text{eff}} = \frac{\lambda_0}{\varepsilon_{r,\text{eff}}} \]  

(6-5)

Recently, it has been proposed to fold the patch in the PIFA in order to reduce even more the antenna size, as in the work [271]. The front view of the structure of a general PIFA with folded patch would be the one shown in Figure 6-3. In next section the proposed and implemented element, based on this type of structure, is depicted (see Figure 6-4).

![Figure 6-3. Front view of a PIFA with two dielectric substrates and folded top patch.](image-url)

Finally, it is worth mentioning that some novel designs have been devised in order to obtain dual-band PIFAs, as the use of meandered patches or with slits or slots. Other options are the use of an extra patch, coupled to the main one, or the use of LC resonators.

### 6.1.1.2 Initial antenna element design.

Since the objective here was to design an antenna array for a PDA-like terminal, it was desirable to use an element with a very small size. Therefore, a PIFA structure with folded patch was chosen as a good candidate.

Specifically, the PIFA structure presented by Chen et al. in [272] was selected, due to its good properties as a dual-band antenna with a very small footprint. The proposed antenna is a PIFA with a meandered patch folded on a supporting dielectric. A small rectangle was cut from the upper part of the patch in order to adjust the two resonant frequencies, which for Chen’s design were 900 and 1800 MHz. A shorting pin was used as shorting mechanism. The feeding probe was situated at one corner of the patch.

We have modified the initial design of Chen, with a two-fold objective: first of all, to obtain a different resonant frequency; secondly, to simplify as much as possible the patch footprint.
Regarding the operational frequency, the antenna element was designed in order to present a main resonant frequency at $f = 1766.6$ MHz. The reason to select this frequency is that a licence to transmit radio signals in this frequency was held by the university, and hence the radio channel measurements were realized at this frequency band. Moreover, this frequency is very close to the upper band of GSM/DCS applications, so the antenna can be envisioned for application in current mobile systems. Furthermore, a second resonant frequency band at $f = 2450$ MHz is desirable, being this one the central frequency for the lower band of WLAN systems. As a result, with these two bands the PDA-type user terminal could be used for both WLAN and mobile communication applications.

In order to reduce the antenna manufacture and design difficulties, the meanders that were introduced in the initial design by Chen were removed. By doing this, the resonant frequency is increased, which in fact is positive to achieve the required frequencies of our design. Furthermore, it was found that also the step of cutting a rectangle in the upper part of the patch could be removed, since the second resonant frequency for the structure can be tuned directly by adjusting the two dimensions $L_2$ and $L_{f2}$ (upper and lower parts of the folded patch) and also the shorting pin position. The final proposed structure for the antenna element is presented in Figure 6-4.

![Figure 6-4. Proposed structure of the element for a compact array for MIMO.](image-url)
It is interesting to note that in the proposed element design the feeding pin is kept at a
fixed position, namely one of the corners of the patch, while the shorting pin position is changed
in order to adjust the resonant frequencies and the impedance matching.

The design procedure was as follows: first of all, the total length of the top patch and
the folded bottom patch is chosen to be about one quarter wavelength of the lower resonant
frequency, 1766.6 MHz (\(\lambda = 169.8\) mm). After that, by adjusting the shorting pin position, the
higher resonant frequency can be tuned and the impedance matching improved. We must note,
however, that the design for both the element and the whole MIMO array was optimized for the
lower frequency, which was the one used for the study of different evaluation methods.
Nevertheless, as we will show later, the performance (regarding return losses) is still fair at the
upper frequency, taking into account commercial values for antennas of handheld devices and
mobile user equipments.

The simulation software CST Microwave Studio for 3D electromagnetic simulations
was used for optimizing the design parameters. As an advantage compared to other
electromagnetic simulators, CST allows a finite ground plane to be simulated, so the dimension
of the ground plane was chosen to fit the size of a generic PDA. After surveying the market,
dimensions of \(L_1 \times W_1 = 125 \times 75\) mm were chosen as a typical size for PDAs.

The finite ground plane, as well as the folded patch and the shorting and feeding pins
were simulated as perfect conductor material. Rohacell foam with \(\varepsilon_r = 1.05\) and \(h_1 = 10\) mm was
selected as the first substrate that is used as separation of the ground plane and the shorted and
folded patch. A substrate with \(\varepsilon_r = 2.33\) and \(h_2 = 0.75\) mm was selected as filling dielectric.

The upper patch initial dimensions were chosen as \(L_1 \times L_2 = 10 \times 20\) mm. The bottom
part of the patch was chosen to be \(L_{f2} = L_2 - 1\) (mm) long, thus folding the patch as much as
possible in order to minimize the element size. The bottom patch must be slightly smaller in
length than the upper patch, so short circuiting with the feeding and shorting pins is avoided.
The overall length of the patch is \(L = L_2 + L_{f2} = 39\) mm, which is close to \(\lambda/4\) as required. The
effect of the patch length, \(L_2\), and the position of the shorting pin with reference to the feeding
port, \(s\), were studied. Fixed values were chosen for the rest of the parameters, and \(L_2\) and \(s\) were
varied, in order to optimally choose them.

In Figure 6-4 the reflection coefficient, \(S_{11}\), is shown for different values of patch length,
\(L_2\). We may note that by varying the upper patch length \(L_2\) also the lower patch length is
changed, since we chose \(L_{f2} = L_2 - 1\) (mm). A fixed value of \(s = 5\) mm was selected. The rest of
the parameters were selected as expressed in previous paragraphs.

As expected, when the length of the patch is increased, both the first \(f_{r1}\) and the second
\(f_{r2}\) resonant frequencies are reduced. The ratio \(f_{r2}/f_{r1}\) is somewhat reduced for higher \(L_2\), that is,
for lower resonant frequencies. We also note that a better matching is obtained for higher \(L_2\) in
the upper resonant frequency band, but no significant differences in matching are observed for the lower band. For \( L_2 = 22 \) mm a resonant frequency very close to the desired working frequency (\( f' = 1766.6 \) MHz) is obtained, so we select it as upper patch length.

![Figure 6-5. S11 as a function of patch length, L2.](image)

Regarding the shorting pin position, we have fixed the patch length to \( L_2 = 22 \) mm, and we have simulated the antenna results for different pin positions. Figure 6-6 shows the \( S_{11} \) parameter for 4 pin positions, where the shorting pin has been moved from the corner of the feeding pin (\( s = 3 \) mm) to the opposite corner (\( s = 9 \) mm). For positions further to the feeding pin, we observed an increase in the resonant frequency, and a slightly improvement in impedance matching. The resonant frequencies ratio \( f_{r2}/f_{r1} \) is reduced for higher frequencies, since the upper resonant frequency is not significantly affected when varying the shorting pin position (as opposed to the case of varying \( L_2 \)). Therefore, by adjusting the shorting pin position and the patch length, both the upper and the lower resonant frequency can be tuned.

![Figure 6-6. S11 as a function of shorting pin position, L2.](image)
Also the effect of different locations of the element in the PDA mockup was studied. Since the final antenna array will consist of several PIFA elements, it is interesting to know if the position in the whole structure will affect the antenna performance. Figure 6-7 shows the return losses $S_{11}$ for two different position of the element on the selected area of the PDA (as depicted in Figure 6-8). The element dimensions were chosen as in previous cases, with $L_2=22$ mm and $s=5$ mm. We see that the matching gets slightly deteriorated for position 2, since the dimensions were optimized for position 1 (where the element was located for previous results). However, the resonant frequency does not suffer any change. We may conclude that for the array design, some small changes in the dimensions of each element may be required in order to optimize matching.

Figure 6-7. $S_{11}$ for two different locations of element in the PDA mockup.

The radiation pattern is also affected by the element position. Figure 6-9 shows the radiation patterns that were obtained when simulating with CST the two studied positions for the element. As observed, the radiation pattern suffers a slight variation due to the different characteristics of the surrounding structure when the element is located at different positions.

Figure 6-8. The two studied position for the element on the selected area.

Figure 6-9. Radiation patterns (azimuth and elevation planes) for the two studied element positions. The same dimensions and element characteristics were used for the two positions.
In the previous figure we observe that the element behaves not as similarly as an omnidirectional element when it is located at one side of the PDA, possibly due to fringing effect and non-ideal finite ground plane. Nevertheless, the diversity that may be obtained due to different radiation patterns obtained for each element location in the PDA structure may also be of interest, since it gives the possibility to include some extra diversity in the MIMO system.

Other design options and characteristics were also studied, as simulating the effect of using a dielectric that covers the whole selected area of 75 $\times$ 75 mm, or changing the ground plane thickness. Since their effect was negligible, the results are not shown here.

### 6.1.1.3 Final element characteristics

The final element for the MIMO array under study was designed taking into account the previous study. The dimensions of the final design, as well as other parameters, are summarized in Table 6-I. For a graphical description of each of the parameters, see Figure 6-4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Size (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Element characteristics</strong></td>
<td></td>
</tr>
<tr>
<td>$L_1$</td>
<td>Patch width</td>
</tr>
<tr>
<td>$L_2$</td>
<td>Upper patch length</td>
</tr>
<tr>
<td>$L_{f2}$</td>
<td>Lower patch length</td>
</tr>
<tr>
<td>$h_T$</td>
<td>Total structure height</td>
</tr>
<tr>
<td>$h_1$</td>
<td>Lower dielectric height</td>
</tr>
<tr>
<td>$h_2$</td>
<td>Upper dielectric height</td>
</tr>
<tr>
<td>$\varepsilon_{r1}$</td>
<td>Relative Dielectric constant, lower dielectric (foam)</td>
</tr>
<tr>
<td>$\varepsilon_{r2}$</td>
<td>Relative Dielectric constant, upper dielectric (substrate)</td>
</tr>
<tr>
<td>$s$</td>
<td>Distance between feeding pin and shorting pin</td>
</tr>
<tr>
<td><strong>PDA mock-up characteristics</strong></td>
<td></td>
</tr>
<tr>
<td>$L_{T1}$</td>
<td>PDA width</td>
</tr>
<tr>
<td>$L_{T2}$</td>
<td>PDA length</td>
</tr>
</tbody>
</table>

Table 6-I. Dimensions of final element design.

With the previous values, the return losses and radiation pattern that were obtained for one element located at the “side” position (position 1 in Figure 6-8) of the PDA are shown in Figure 6-10 and Figure 6-11, respectively (Figure 6-12 shows the cuts of radiation pattern for $\theta=90^\circ$, and $\phi=90^\circ$). These results were obtained from simulations by using CST Microwave Studio. Figure 6-13 shows the final design view in this simulation software.
Antenna aspects of MIMO systems

Figure 6-10 $S_{11}$ parameter and Smith Chart for the final element design.

Figure 6-11. Radiation pattern representation for the final PIFA element (simulation results).

Figure 6-12. Radiation pattern, cuts for $\theta=90^\circ$ (left) and $\phi=90^\circ$ (right).
As shown in Figure 6-10, the frequency bandwidth for the band of interest (lower band, centred at working frequency $f = 1766.6$ MHz) is $BW_l = 165$ MHz, that is, a bandwidth of 9.3%. Thus, the obtained bandwidth widely fulfils the requirements for MIMO extensions of several wireless communication systems, such as MIMO-WLAN communications, defined in the draft for 802.11n (which proposes a bandwidth of up to $BW = 40$ MHz [11]) and MIMO in 3G-UMTS (which proposes a bandwidth of $BW = 5$ MHz [10]). The obtained frequency bandwidth for the second frequency band (upper band) is smaller (75 MHz, or a 3\% relative bandwidth), but still fulfils well enough the recommendations. The bandwidth was defined for a return loss limit of 6 dB, being this value a commonly used limit when designing antennas for user terminals and handheld devices [274] as the one in this case (it represents that a 25\% of the power is reflected).

Regarding radiation pattern, we observe in Figure 6-11 and Figure 6-12 that the obtained radiation diagram is neither directive nor ideally omnidirectional. A wide beam with maximum radiation value is obtained for an elevation angle slightly higher than the horizontal plane, and mainly at the x-axis. This behaviour may be due to the position of the element, closely located at the side of the PIFA mock-up. The radiation level is lower for the bottom spatial hemisphere ($z<0$) which is also reasonable since there is a ground plane (perfect conductor) at the bottom of the structure. The maximum gain value that is obtained from simulation results is:

$$G_{\text{max}} = 2.88 \text{ dBi} \quad (\phi_{G_{\text{max}}} = 15^\circ, \theta_{G_{\text{max}}} = 50^\circ)$$

For comparison reasons, we can observed that the obtained gain is higher than the theoretical directivity for the $\lambda/2$ dipole ($D_{\text{dip}} = 2.15$ dBi) but lower than the theoretical directivity for the $\lambda/4$ monopole on a ground plane ($D_{\text{mon}} = 5.15$ dBi).

As a conclusion, we can say that the obtained element provides a very small footprint, with electrical dimensions of $0.13\lambda \times 0.06\lambda$ at 1766.6 MHz, making it suitable for a compact design. The radiation pattern is sufficiently omnidirectional, and different locations in the PDA
may provide different patterns, thus offering the option of achieving an extra diversity in a 
multi-element antenna. Double-band matching was obtained, with good results regarding return 
loss and bandwidth, considering that a patch-type structure was used. The bandwidth, larger 
than 150 MHz for the lower band, is good enough for a realistic communication system.

6.1.2 Antenna array: low mutual coupling and compact design

After the design of one element for the MIMO antenna, we must choose the number of 
elements, their relative and absolute position on the PDA mockup and the whole array size. Moreover, some adjustments may be required in the original element design in order to optimize 
the overall array from a MIMO system point of view.

The concept “optimal design” is not so clear from a MIMO perspective. From a 
classical point of view, an antenna array has usually some radiation pattern requirements for the 
array, which is considered as a single-port radiant device with a certain feeding network. In this 
case, mutual coupling between elements and differences in radiation pattern of elements are 
undesirable effects and they should be mitigated, since they cause changes in the array radiation 
pattern. As a result, the elements in a conventional array are not very closely located (a spacing 
of $\lambda/2$ is a typical spacing).

Unlike the classical perspective, in a MIMO system each element is considered as a 
single radiant device, and the received or transmit signals are not directly combined. Therefore, 
multiple solutions regarding radiation requirements may be adopted as adequate ones for a 
MIMO array. Regarding spacing between elements, the closer the elements are located the more 
compact the design is. However, close elements suffer from high mutual coupling and a certain 
variability in radiation pattern of elements. For a MIMO system, the first one may turn to be 
even a positive effect, since some differences between radiation pattern of elements may imply 
a higher decorrelation between received or transmitted signals, and thus higher diversity and 
system capacity. When properly designed, this may be used to achieve multiple parallel 
subchannels, with the sometimes called radiation pattern diversity.

Mutual coupling among elements, which usually increases when reducing elements 
spacing, introduces two effects: correlation between transmitted/received signals and a certain 
impedance mismatch. The first one is usually considered undesirable in MIMO systems, 
although some processing schemes may take advantage of it. Impedance mismatch, however, is 
always a negative effect, since it reduces the antenna efficiency and thus the system gain and 
capacity.

We have optimized the design of the MIMO array by trying to minimize the mutual 
coupling among elements, thus reducing the correlation between signals at the input ports. 
Simultaneously, a design as compact as possible was desirable.
The number of elements to be used was chosen as four, which is a reasonable number for user terminals, considering size and weight constraints. A subarea of $75 \times 75 \text{ mm}^2$ within the whole PDA area ($125 \times 75 \text{ mm}^2$) was selected for placing the radiant elements, and several configurations were studied. As in previous sections, CST Microwave Studio was used to simulate the array electromagnetic performance, and coupling coefficients were compared to select the best array configuration.

Multiple array configurations were simulated and analysed. For the sake of summarizing, only the most interesting ones are shown. Figure 6-14 depicts the 4 main array configurations that were studied. In the case of configurations 2 and 3, the 4 elements were placed with the same relative position (horizontal or vertical placement, respectively), while for configurations 1 and 4 horizontally- and vertically-placed elements were alternatively selected, in order to reduce correlation. Moreover, since the elements tend to have an elliptical polarization, in configurations 1, 3 and 4 they were placed so that the feed probes were located in a mirror fashion, i.e., at alternate corners, to provide as high polarization diversity as possible and thus low correlation of received/transmitted signals.

![Figure 6-14. Main array configurations that were studied. The grey zone represents the $75 \times 75$ selected area for the elements placement. Big and small dots at the corner and side of each element show the position of feeding and shorting pins, respectively.](image)

Figure 6-15 shows the S-parameters for each of the 4 studied configurations. First of all, we see that the $S_{ii}$ parameters (return loss) varies for each element, even if the dimensions are kept the same for all the elements. This may be due to the effect of different locations with respect to ground plane (and thus, different bound conditions) for each element. As a result, a small adjustment in patch length may be considered for the final array design. In Figure 6-15 we also observe that mutual coupling is quite similar for the four configurations, although configurations 1 and 3 seems to give lower coupling levels, while configuration 4 is the one with higher mutual coupling level.

In order to represent mutual coupling in detail for the working frequency, Figure 6-16 shows a zoom of the S parameters for the 4 studied configurations for the lower frequency band (only S parameters with higher values are shown, for clarity reasons).
Figure 6-15. S parameters for the 4 studied configurations.

Figure 6-16. Selection of higher coupling coefficients (zoom) for the working frequency \( f=1766.6 \) MHz, for the four studied configurations.

In the previous figure we observe that configuration 1 is the one with lower overall coupling coefficients, complying with a reasonable target for the maximum coupling coefficient
This target $S_{ij}$ value is exceeded by $S_{43}$ in configuration 2, $S_{32}$ in configuration 3 and $S_{21}$, $S_{41}$ and $S_{32}$ in configuration 4. As a result, configuration 1 was chosen as the best suited for the final array to be implemented.

### 6.1.3 Final array characteristics and prototype implementation

As stated above, configuration 1 with rotated and mirrored elements was chosen for the final compact array that would be studied as MIMO array, due to its good characteristics regarding compact design and low mutual coupling.

Figure 6-17 depicts the structure of the final array, as well as a top view highlighting the feeding pins. A whole layer of dielectric material was simulated, instead of simulating the substrate just as small pieces under the patches. This is done to accurately represent the structure that would be implemented, which would have a single substrate layer in order to simplify the fabrication. As presented in the study above, different locations for each element generate slight differences on matching frequency. Thus, the patch length for each element $L_1$ was optimized in order to obtain minimum return losses at working frequency $f = 1766.6$ MHz. The final dimensions for each element are specified in Table 6-II, as well as final dimensions for the whole structure. The radiation and impedance characteristics of the final array design, which were obtained with the simulation software (CST Microwave Studio™) are presented below.

In Figure 6-18 the obtained reflection coefficient for each element with the final array design is presented. We observe that a good matching is obtained for the four elements. The best matching is achieved for element 3 while a worse but still fair matching is obtained for element 4. The attained bandwidths (assuming minimum return losses $S_{ii}<-6$ dB) go from 120 MHz (element 4) to 180 MHz (element 3).

Figure 6-19 represents the antenna gain for each element, in 3-D view, while Figure 6-20 shows the same parameter in a 2-D fashion. We should state here that since the dielectric was simulated as loss free material, the obtained directivity and gain are very similar.
Antenna aspects of MIMO systems

Element in final design for MIMO array (enumeration as in Figure 6-17)

<table>
<thead>
<tr>
<th></th>
<th>Element 1</th>
<th>Element 2</th>
<th>Element 3</th>
<th>Element 4</th>
</tr>
</thead>
<tbody>
<tr>
<td>L₂ (Upper patch length), [mm]</td>
<td>22</td>
<td>22</td>
<td>21</td>
<td>21</td>
</tr>
<tr>
<td>L₂ (Lower patch length), [mm]</td>
<td>21</td>
<td>21</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>s (Distance between feeding pin and shorting pin) [mm]</td>
<td>6</td>
<td>6</td>
<td>6</td>
<td>5</td>
</tr>
<tr>
<td>L₁ (Patch width) [mm]</td>
<td>10</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>h₁ (Total structure height), [mm]</td>
<td>10.75</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>h₁ (Lower dielectric height) [mm]</td>
<td>10</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>h₂ (Upper dielectric height) [mm]</td>
<td>0.75</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>εᵣ₁ Relative Dielectric constant, lower dielectric (foam)</td>
<td>1.05</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>εᵣ₂ Relative Dielectric constant, upper dielectric (Rohacell)</td>
<td>2.33</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>L₇₁ (PDA width) [mm]</td>
<td>75</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>L₇₂ (PDA length) [mm]</td>
<td>125</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 6-II. Dimensions of final array design.

Figure 6-18. Simulated reflection coefficient for final array.
Figure 6-19. 3D representation of radiation patterns (far field) for each element in the final array design. The radiation patterns are depicted over each respective element, for intuitive representation purposes.

Figure 6-20. Simulated gain for final array design.
A remarkable characteristic to be observed in previous figures (Figure 6-19 and Figure 6-20) is that each element presents different radiation patterns. Indeed, element 1 shows a maximum gain in the positive x-axis direction \((\phi=0^\circ)\) for elevation angles close to the horizontal plane \((\theta \sim 0^\circ)\). Another main lobe, although with slightly lower gain, is observed for the same axis but for the opposite direction \((\phi=180^\circ)\), but the beam is pointing at higher elevation angles \((\theta \sim 30^\circ)\). For element 2, it is also the x-axis direction the one with higher gain; both for positive and negative directions the main lobe is slightly pointing at the bottom of the structure \((\theta \sim 140^\circ)\). Element 3 presents its main lobes at the x-axis direction again, with the lobe pointing at the positive part of the axis \((\phi=0^\circ)\) as the one with the higher gain, and focused again on a wide angle range over the horizontal plane \((\theta \sim 30^\circ)\), but with an important gain also for \(\phi=180^\circ\). Finally, element 4 shows the highest differences with the previous elements, since only one main lobe is observed, with a quite high gain \((g_{\text{max}} \sim 5 \text{ dBi})\) compared with the other elements (a difference of around 3 dB between maximum gain is observed). The maximum gain for this element is obtained for the negative x-axis direction, for an elevation angle of approximately \(\theta \sim 45^\circ\) (pointing over the horizontal plane).

This behaviour is reasonable, since each element is strongly influenced by the electrical environment of its own position, patch location and situation of feeding pin. Moreover, this fact implies that a good pattern diversity may be obtained, since each element will provide with a quite different radiation behaviour, should similar impinging signals arrive to all the elements.

**Prototype implementation**

A prototype of the designed MIMO array was implemented, following the dimensions of Table 6-II as precise as possible. A 10-mm thick Rohacell foam was used for the lower dielectric, while the copper patches were etched on a 75×75mm Teflon substrate of thickness 0.75mm. The ground plane was implemented with a 125×75mm brass plate, and both the upper and lower patches were realized with copper and linked with Cu-tape. Figure 6-21 shows a photograph of the implemented antenna array.

The antenna array was firstly characterized by measuring conventional antenna parameters: reflection coefficient, radiation pattern and radiation efficiency. Figure 6-22 a) shows the \(S_{ii}\) parameters of each antenna element, measured with an HP 8753 Network Analyzer. Figure 6-22 b) represents the coupling coefficients for the higher coupling pairs of elements. In both figures, the measured results are compared with the ones obtained from the simulated array. It is clear that the agreement between measured and simulated results is good, especially considering that the folded shape of the antenna patch was made with a non-precise technique (just soldering a piece of Cu-tape), which made it difficult to achieve the expected L length of the patch. This fact may explain why it is the upper resonance frequency which differ the most between simulation and measurements. Still, the results are very good. Regarding
achieved bandwidth for an $S_\text{dB}<6$ dB, we observe that the bandwidth varies for both the upper and the lower resonant frequencies, being for the best-case element of 9 % and 3.1 % at 1766 MHz and 2450 MHz, respectively. As for the coupling coefficients, we may note that also for the implemented prototype the coupling is quite large, although values of coupling coefficient which are lower than the targeted 10 dB are obtained.

![Implemented compact antenna array.](image)

Figure 6-21. Implemented compact antenna array.

![a) reflection coefficients](image)

![b) Coupling coefficients (worst cases)](image)

Figure 6-22. Reflection (left) and coupling (right) coefficients for the implemented prototype. Dashed lines represent the previous results (simulation) for the same design.

The full 3D radiation pattern was also measured in an anechoic chamber. 128 evenly distributed angles were measured in azimuth, and 64 positions in elevation, using the SATIMO chamber of AMC Centurion. The results are shown in a 2-D fashion in Figure 6-23.
It is remarkable the similarities we observe in measured and simulated radiation patterns (Figure 6-20 and Figure 6-23), being the main lobes pointing at the same spatial directions. However, we observe that the antenna efficiency is lower for the implemented array than for the simulated one. We see that the maximum measured gain for each array element is 1.77, 1.71, 2.05 and 3.31 dBi, respectively, when for the simulated array this same parameter is 4.34, 4.41, 4.42 and 4.28 dBi for each element (from 1 to 4). This may be explained by two reasons: firstly, the second substrate was simulated as a loss-free material, and although its thickness is very small (especially when compared with the Rohacell first substrate) it may affect the antenna efficiency; Secondly, some radiation losses such as radiation coupling may be difficult to be simulated and included with the simulation software.

The antenna efficiency was computed from the measured gain over all the measured angles. We must take into account that the obtained gain is the rate between radiated power $P_{\text{rad}}$ and the incident power $P_{\text{in}}$ on the antenna port, and thus the loss due to reflected power is also considered in this parameter,

$$\eta = \frac{P_{\text{rad}}}{P_{\text{in}}} \quad (6-6)$$

The efficiency is shown in Table 6-III and Table 6-IV, both the measured and the simulated values, respectively. The results were obtained for both the high and low frequency bands of
interest. We computed the efficiency for different high frequency for each element, in order to consider the different center frequency for the upper band (mainly due to small manufacture flaws as explained above). We observe that a high total antenna gain is obtained with the simulation tool, while a more realistic value for this type of antenna (close to 0.7) is obtained with the implemented prototype. The differences may be due to using a perfect electric conductor in the simulation. Moreover, for the simulation the obtained radiation efficiency (which does not take into account return losses) is very close to 1 (0.99 or even 1 for some ports) which is not very reasonable for a real antenna.

<table>
<thead>
<tr>
<th>Element</th>
<th>1766 MHz</th>
<th>2444 MHz</th>
<th>2500 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Element 1</td>
<td>0.89</td>
<td>0.91</td>
<td>-</td>
</tr>
<tr>
<td>Element 2</td>
<td>0.83</td>
<td>0.80</td>
<td>-</td>
</tr>
<tr>
<td>Element 3</td>
<td>0.80</td>
<td>-</td>
<td>0.82</td>
</tr>
<tr>
<td>Element 4</td>
<td>0.77</td>
<td>-</td>
<td>0.78</td>
</tr>
</tbody>
</table>

Table 6-III. Simulated total antenna efficiency for compact array

<table>
<thead>
<tr>
<th>Element</th>
<th>1766 MHz</th>
<th>2444 MHz</th>
<th>2500 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Element 1</td>
<td>0.734</td>
<td>0.761</td>
<td>-</td>
</tr>
<tr>
<td>Element 2</td>
<td>0.625</td>
<td>0.762</td>
<td>-</td>
</tr>
<tr>
<td>Element 3</td>
<td>0.694</td>
<td>-</td>
<td>0.659</td>
</tr>
<tr>
<td>Element 4</td>
<td>0.680</td>
<td>-</td>
<td>0.716</td>
</tr>
</tbody>
</table>

Table 6-IV. Measured total antenna efficiency for compact array

As a result, we may conclude that a prototype should always be preferred if accurate parameters are need, especially regarding gain and antenna efficiency.

6.1.4 Reference antenna array

In order to be able to validate the compact array performance, a not so realistic antenna array was also implemented. The main features should be good efficiency and omnidirectional radiation pattern, so no direction is given preference. Moreover, a simple implementation is appreciated, as well as low coupling between element pairs. Obviously, the same number of elements as in the case of the compact array should be chosen, in order to get similar characteristics for a MIMO system.

An array of four monopoles was chosen as reference antenna array, due to its good characteristics of efficiency, easy to manufacture and non directive radiation pattern. Since only one frequency can be chose as the matching frequency for these type of antennas, the
monopoles were designed in order to resonate at the main working frequency of the study, that is, \( f = 1766 \text{ MHz} \). The \( \lambda/4 \) monopoles were made of compact brass cylinders, fed by coaxial SMA connectors. They were situated at spacing \( \lambda/2 \), over a \( 170 \times 425 \text{ mm}^2 \) \((1\lambda \times 2.5\lambda)\) brass plane, which represented the ground plane. Since the full dimension of the array is quite big, it is not adequate to be used as a realistic MIMO array on user terminals, but it is a good reference array, thanks to its minimal losses (good impedance matching and loss-free material used to implement it) and low antenna coupling. The monopoles length was optimized in order to improve the matching for the working frequency. As a first value, it was selected as slightly smaller than the ideal \( d_{th} = \lambda/4 = 4.25 \text{ cm} \), in order to take into account the currents on the top of the rod (due to not having a zero-diameter element). The Tai equation was used to initially calculate the length to be used for the monopole arrays. Afterwards, simulations of the monopoles with the 3D electromagnetic simulation software CST Microwave Studio were performed, in order to optimize the length of the monopoles (\( L_m \)). Also the spacing between the monopole and the ground plane (\( L_s \)) was optimized by adjusting the length of the inner conductor of the coaxial feed (and thus the length of the outer dielectric of the SMA connector), since it was observed via simulations that the input impedance could be varied by changing this parameter. Figure 6-24 shows a snapshot of the simulated structure for a single monopole, showing the optimized parameters.

![Figure 6-24. Simulated element for the reference array.](image)

Finally, the implemented monopole arrays were characterized by measuring its main radiation and matching parameters. Figure 6-25 shows a photograph of the implemented monopole array.
Figure 6-25. Photograph of the implemented reference array (4-monopole array)

Figure 6-26 represents the measured reflection coefficient for each element of the reference array. A good matching is achieved, as expected. The obtained bandwidth for a reflection coefficient lower than -6 dB is larger than 1100 MHz, thus much higher than for the compact array.

Coupling coefficients are also depicted in Figure 6-26, for the worst-case values. Again, good results are obtained, with Sij parameters lower than -14 dB in all cases, thanks to the fair spacing between elements and the non existence of surface waves (although radiation coupling may exist).

Figure 6-26. Measured S parameters for the reference monopole array.
The radiation pattern of the reference array was measured in the same anechoic chamber and with the same measurement system as for the compact array, and the obtained gain patterns for each element are shown in Figure 6-27 (3-D representation) and Figure 6-28 (2-D representation).

Figure 6-27. Measured 3D radiation pattern for the reference monopole array.

It is clear that the obtained radiation patterns are more omnidirectional-like than for the case of the compact array. For the reference array, we observe that the maximum gain has a very small variation among elements, being 4.34, 4.41, 4.42 and 4.28 dBi for each element, from 1 to 4. This fact is due to very small differences between elements, thanks to the easy manufacturing of this type of antenna. The maximum gain occurs at an elevation angle $\theta_{G_{\text{max}}} = 50^\circ$, for all the elements. This separation from the theoretical maximum elevation angle ($\theta_{\text{theor}} = 90^\circ$) is caused by the finite ground plane. All elements show a null at $\theta = 0^\circ$, as expected. In the azimuth $\phi$ direction, the maxima show the symmetry of the array with respect to the xz-plane. Therefore, element 1 at one end has maxima at $\phi = 180^\circ \pm 60^\circ$, whereas element 4 at the other end has maxima at $\phi = \pm 60^\circ$. 
Finally, the efficiency was also measured for the reference array, and the results are shown in Table 6-V. As expected, the reference array has very good antenna efficiency, due to the used antenna type.

<table>
<thead>
<tr>
<th>1766 MHz</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Element 1</td>
<td>0.928</td>
</tr>
<tr>
<td>Element 2</td>
<td>0.910</td>
</tr>
<tr>
<td>Element 3</td>
<td>0.897</td>
</tr>
<tr>
<td>Element 4</td>
<td>0.916</td>
</tr>
</tbody>
</table>

Table 6-V. Measured antenna efficiency for reference array

6.1.5 Conclusions

The design, implementation and measurement results of a compact array for MIMO applications have been presented. A PIFA element has been selected as element type, due to its very small size, robustness, planar features and easy integration in handheld devices. The requirements for a realistic antenna array have been highlighted and considered during design, such as small size and simplicity in implementation.
Conventional antenna parameters have been measured, specifically the radiation pattern of each element and the S parameters of the array. The results show a good performance regarding radiation features, reflection coefficient and coupling between elements. It has been shown that the measurements confirm the good results previously obtained by means of simulation.

In order to better evaluate the performance of the designed antenna, a reference array was also implemented. A very simple monopole array was chosen, due to its well-known characteristics. The same conventional parameters were selected, showing the omnidirectional radiation patterns of the elements, the low coupling and the good matching of each element, as expected.

Although the results obtained by evaluating conventional parameters give a good overview of the antenna performance as a radiant system, it is difficult to evaluate how good it would perform in a MIMO system. There, not only gain and impedance matching are important, but also other features such as radiation diversity or decoupling between antennas. For example, we observe that the monopole elements in the reference array present very similar radiation patterns, while the PIFA elements in the compact array show much more different radiation patterns. This different behaviour must be considered when evaluating the antenna array for MIMO, since it involves different correlation levels and diversity among received (or transmitted) signals. Next chapter presents several methods that allow the evaluation of an antenna array from a MIMO system perspective.
6.2 Evaluation of antenna arrays for MIMO

Typically, conventional parameters such as impedance matching, reflection coefficient or gain have been used to evaluate an antenna performance. When the antenna consists of several elements, as in antenna array cases, other parameters such as coupling coefficients or issues regarding the feeding network are also considered. The natural next step, currently being a very active field of research, is to consider which parameters have to be measured in order to evaluate a multi-element antenna for MIMO systems. In these systems, the antenna itself is not a single-port module, but a multi-port one, and each received (or transmitted) signal is (generally speaking) independently processed.

Along this section, three different methods to evaluate an antenna array for MIMO systems are proposed and compared. Each method is firstly presented in a general way, so it can be used for a general antenna array. After that, results for the specific case of the designed compact array and the monopole reference array are shown.

Finally, the three methods are compared regarding complexity, information they provide and possible drawbacks of each one.

6.2.1 Evaluation using measured radiation patterns

Taking the conventional antenna parameters as a stating point, the next step in order to account for multiple input (or output) signals is to account for the antenna covariance, or more generally speaking, to calculate the correlation of the signals after being received (or transmitted) by the antenna array.

By using the covariance as an input parameter, the capacity of the MIMO system can be estimated by using different channel models. The theoretically achievable capacity is one of the key parameters to evaluate MIMO systems, together with the achievable throughput or bit rate (for a certain BER). Therefore, we propose below how to compute the capacity of the MIMO system taking into account the antenna covariance (and thus the radiation patterns). The specific results for the implemented compact array and reference array are also shown and commented.

6.2.1.1 MIMO capacity using the Kronecker model and the antenna covariance

We present here a method to compute the capacity of a MIMO system when the antennas are considered by including the radiation pattern effect. A simple channel model that assumes independence between transmit and receive distribution of signals (the so called Kronecker model) has been combined with the computation of the covariance at the antenna ports, including the effect of having a different radiation pattern for each element.
MIMO capacity for a narrowband channel

In order to present capacity results below, we summarize here the capacity computation for a MIMO system, focusing on the specific case under study. A comprehensive study on the capacity of MIMO systems and its theoretical computation has been presented in chapter 2.

It is well known that the input $x(t)$ and output $y(t)$ in a narrowband MIMO system are related by the channel matrix $H$:

$$y(t) = Hx(t) + n(t)$$  \quad (6-7)

where $n(t)$ represent additive noise. In the case of non-ideal antennas (a more realistic case than assuming ideal omnidirectional antennas), $H$ should include all the antenna effects that would change the original transmitted signal $x(t)$ or the received signal $y(t)$, such as the radiation pattern at both the transmit or receive side. The capacity for a system with $n_T$ transmit antennas and $n_R$ receive antennas with full channel knowledge (CSI) at the transmitter is given by the well-known expression (e.g.[275]):

$$C = \max_{Q} \log_2 \det \left[ I_{n_R} + QH^H H \right] \quad \text{bits/s/Hz} \quad (6-8)$$

where $Q$ is the covariance matrix of transmitted signals fulfilling the power constraint $Tr(Q)=\rho$. In the case of $n_T=1$, the meaning of $\rho$ is the average signal to noise ratio (SNR) at the receiver. The maximum capacity is achieved using water-filling so that the singular value decomposition (SVD) of $H$ is realized. In the case of no CSI at the transmitter, we simply use $Q=\rho/n_T I$ so that [62]:

$$C = \log_2 \det \left[ I_{n_R} + \left( \rho/n_T \right) HH^H \right] \quad \text{bits/s/Hz} \quad (6-9)$$

The “Kronecker Model”

As explained in chapter 2, the Kronecker model is a simplified MIMO channel model which assumes that the angular and power distribution of signals at the transmitter is independent of the distribution of signals at the receiver. In other words, it is assumed that, in their travelling along the environment, the signals “forget” the way they were sent to the channel. Obviously, this assumption is not true for real life; however, it can be considered accurate enough in a rich multipath and scatterers environment, and it can serve as a first approach to get the antenna performance in a MIMO system. This way, the effort is not put in simulating a very complex channel, but in studying the antenna effect.

In the Kronecker model, the normalized channel matrix $H$ is modelled as [104]:

$$H = 1/ g (R_{Rx})^{1/2} G (R_{Tx})^{1/2} \quad (6-10)$$

Here, $G$ is a stochastic $n_R \times n_T$ matrix of i.i.d. $\mathcal{CN}(0,1)$ elements and $g$ is a normalization constant chosen so that $\mathbb{E} \left[ \| H \|_F^2 \right] = n_T \cdot n_R$ for uncorrelated antennas. $R_{Rx}$ and $R_{Tx}$ are the covariance
matrices at the receiver and the transmitter side, respectively. The operation $(\cdot)^{1/2}$ makes reference to the Choleski decomposition of a matrix $A$, that is, $(A)^{1/2}=B$, so that $B^TB=A$.

**Covariance calculation**

The receive (or transmit) covariance between antenna elements $m$ and $n$, with field patterns $E_m$ and $E_n$ respectively, is calculated as:

$$R(m,n) = \int_\theta \int_\phi E_m^* S(\theta, \phi) \sin(\theta) d\theta d\phi$$

where $S(\theta,\phi)$ is the distribution of an incident (or transmitted) field of random polarization. Therefore, the covariance matrix $R$ depends largely on the radiation patterns of the antennas (by means of $E_m$ and $E_n$) but also on the channel (by means of $S(\theta,\phi)$). The distribution of $S(\theta,\phi)$ has been estimated in e.g. [275], but, in order to emphasize the effect of the antenna patterns, we propose to use a continuous incident field distribution (as in [276], [277]).

With the aim to highlight any dependence on $\theta$, we propose to firstly use a uniformly distribution function in $\theta$:

$$S_1(\theta, \phi) = S_1(\theta) = \frac{1}{4\pi \sin(\Delta\theta/2)}$$

$$|\theta - \pi/2| < \Delta\theta/2$$

and zero otherwise. Thus $\Delta\theta \to 0$ corresponds to incidence in the horizontal plane only. The variance is normalized so that its value for $\Delta\theta = 180^\circ$, i.e. isotropic incidence, is equal to the antenna efficiency $\eta$. Generally, for modelling outdoor scenarios the elevation power spectrum is neglected, since most of the power impinges in the horizontal plane. However, for mobile terminal located in indoor building the elevation spectrum must also be considered, hence the interest in studying the arrays covariance as a function of elevation incidence region $\Delta\theta$.

Next, let us study the effect of distribution of incident fields in azimuth $\phi$. For an indoor scenario, it is commonly assumed an isotropic azimuth distribution, uniform over $360^\circ$. However, for certain cases such as LOS situations and corridors, it is reasonable to assume that a significant amount of power impinges from a confined region $\Delta\phi$. Thus, we propose also to study the covariance when the distribution function $S(\theta,\phi)$ is modelled as a uniformly distributed function in $\phi$:

$$S_2(\theta, \phi) = S_2(\phi) = \frac{1}{2\Delta\phi}$$

$$|\phi| < \Delta\phi/2$$

and zero otherwise. Thus, $\Delta\phi \to 0$ corresponds to impinging fields in the broadside direction. Also in this case, the variance is normalized so that its value for $\Delta\phi = 360^\circ$, i.e. isotropic incidence, is equal to the antenna efficiency $\eta$.  

266
Results for the compact array and comparison with reference array

Using the previous methodology, the covariance matrix and MIMO capacity for the compact array presented in section 6.1 has been computed. For comparison reasons, the same parameters were also computed for the reference array.

Firstly, the covariance has been studied as a function of elevation, following expressions (6-11) and (6-12). Figure 6-29 shows the variance (that is, the diagonal of \( \mathbf{R} \)) for the compact array (PIFAs, also called AUT: array under test) and the reference array (monopoles), while in Figure 6-30 the covariance (that is, \( R_{n,m} \), for \( m \neq n \)) between antenna pairs have been depicted, for both antenna arrays. We may note that some covariance values are equal (\( R_{12} = R_{21} \), and so on), as deduced from the definition of covariance as given by eq. (6-11). The variance values for the isotropic case (at the far right of the figure, where \( \Delta \theta = 180^\circ \)) correspond to the measured efficiency \( \eta = \frac{P_{rad}}{P_{in}} \) of the antennas. Note that this efficiency includes the loss due to reflected power, i.e. \( P_{in} \) is the incident power on the antenna port. As expected, the reference array elements have a higher efficiency. Also note that variance is greater than unity for the monopoles of the REF array when the incident field is confined to \( \Delta \theta \approx 90^\circ \) due to the element pattern having a maximum slightly above the \( xy \)-plane.

It is clear that both arrays are quite insensitive to the distribution of the incident field in \( \theta \). For the monopoles which are separated by \( \lambda/2 \) we have almost zero correlation when \( \Delta \theta = 180^\circ \), but as \( \Delta \theta \to 0 \), the correlation increases slightly. The AUT elements show slightly higher correlation in the isotropic case, but on the other hand it is lower as \( \Delta \theta \to 0 \). Provided that the azimuth distribution may be expected to be uniform over \( 360^\circ \) in an indoor environment, we conclude from these results that in both cases the MIMO performance should be limited by the radiation efficiency which is about 1.3 dB lower for the compact array (AUT) than for the reference monopole array.

Secondly, the covariance as a function of the incident region in azimuth has been studied, following expressions (6-11) and (6-13). The results are shown in Figure 6-31 for the variance and Figure 6-32 for the covariance. Regarding variance and covariance as a function of incident azimuth region \( \Delta \phi \), we may note that the variance for the reference array is quite insensitive to \( \Delta \phi \), as expected for omnidirectional antennas. But conversely to the previous case, the covariance for both arrays is highly dependent on the incidence region in azimuth. For the reference array, the covariance values approximately follow a Bessel function, which is reasonable for omnidirectional elements: whereas highly correlated signals are obtained for broadside incident region, the correlation would be small for a uniform \( 360^\circ \) incidence region. On the other hand, the AUT exhibits a lower covariance for small incidence region, due to the significant differences in radiation pattern between antenna elements. This also explains the different variance values for the AUT, being higher for element 1, whose pattern has a maximum at \( \phi = 0 \), and lower for element 4, which exhibits a minimum close to \( \phi = 0 \), as shown in Figure 6-19, Figure 6-20 and Figure 6-23. When increasing \( \Delta \phi \), covariance is reduced in
general. For incidence region close to 360° the covariance for AUT is slightly higher than for the reference array, as also shown in Figure 6-32. Initially, we could think that the expected behaviour should be to obtain a quite high covariance for the reference array, since the radiation patterns are very similar for the four elements. However, when computing the covariance of received signals not only the pattern module should be considered, but also the absolute and relative phase differences due to different pattern phases and also different position of elements will make a difference. As a result, the reference array shows a very low covariance value between element pairs for certain incident regions in azimuth, corresponding to cases where the received signals would be added as opposite signals (destructive addition).

Figure 6-29. Variance as a function of incidence region in elevation, computed from the radiation patterns and eqs. (6-11) and (6-12).

Figure 6-30. Covariance as a function of incidence region in elevation, computed from the radiation patterns and eqs. (6-11) and (6-12).
Antenna aspects of MIMO systems

After computing and evaluating the covariance matrices for both the compact array and the antenna under test, it is of utmost interest to calculate the achievable capacity for each of the studied antenna arrays, since it will give us an idea of the performance of each array under MIMO channel characteristics.

Once the covariance is known, it is straightforward to compute the theoretical capacity with the simple Kronecker channel model, by using eqs. (6-9) and (6-10). Although the assumption of independent field distribution at transmitter and receiver side may be unrealistic, this method allows to highly simplify the channel model, and hence the capacity computation. 1000 independent trials were realised, in order to account for the random nature of the channel. In Figure 6-33 we see the Kronecker average capacity for a 4×4 system with average receive SNR = ρ = 10 dB for the REF antennas and an ideal uncorrelated TX array (R_{Tx} = I) for different incident field distribution, varying over elevation Δθ and azimuth Δφ. Also the theoretical capacity that would be obtained with ideal omnidirectional antennas and
uncorrelated i.i.d. channel is presented. As expected from the covariance analysis, we note that both arrays are quite insensitive to variation in elevation distribution, and the reference array offers a higher capacity. Since the covariance properties are quite similar for both arrays, we conclude that the higher capacity for the reference array is due to its higher antenna efficiency, which also explains higher variance values for the reference array. When analysing the capacity obtained when varying the azimuth incidence angle, we observe that in general the achievable capacity is higher for the reference array, specially for large incidence regions (that is, no specific angle of arrival being preferred), which agrees with the lower covariance values obtained for the reference array. However, for a small incidence region $\Delta \phi < 40^\circ$, as the case of hallway propagation, the AUT obtains a better capacity. This is due to radiation pattern diversity offered by the AUT and lacking in the reference array for this specific case. This also explains the higher covariance values for the reference array for small $\Delta \phi$. Nevertheless, we must notice that in general indoor scenarios it is unlikely to observe very small $\Delta \phi$, being realistic only for very specific situations as may be LOS or corridors.

Figure 6-33. Calculated MIMO average capacity using the Kronecker model with 4 uncorrelated antennas at the TX and the measured patterns of REF and AUT arrays at RX.

We can also represent the cumulative distributed function CDF of the MIMO capacity. Figure 6-34 shows it, for an SNR = 10 dB. For this representation, no specific direction is given preference, so we assume that impinging waves are uniformly distributed all along $\theta = [0..180^\circ]$ ($\Delta \theta = 180^\circ$) and $\phi = [0..360^\circ]$ ($\Delta \theta = 360^\circ$).
Figure 6-34. CDF of capacity for SNR=10 dB, for the AUT and the reference array. Kronecker model and measured radiation patterns are used.

Again, we observe that the reference array offers a better performance, for both the mean capacity and also for the outage capacity at 10%: The reference array shows a slightly lower capacity than the i.i.d. channel, due to its good radiation efficiency, very close to 1. The result here emphasizes that this method (use of radiation patterns and the Kronecker model) is very influenced by the efficiency of the antenna, and it is especially focused on the efficiency differences, while different behaviour of arrays for different situations regarding multipath distribution or other channel characteristics are neglected here. This explains that the reference array shows so similar performance to the i.i.d channel. Moreover, simulating an ideal omnidirectional antenna array would give equal results than the i.i.d. channel, since in the Kronecker model the antenna arrays are assumed to receive rays from all spatial directions.

Finally, and for comparison reasons, the average capacity for different SNR values has been computed, for both the AUT and the reference array. Figure 6-35 shows the result, where also the capacity for i.i.d. channel is plotted. The results agree with the previous ones, showing a better performance of the reference array compared with the compact array. Specifically, the reference array offers an improvement of approximately 1 dB for obtaining the same capacity (improvement which is a bit lower for low SNR values), or equivalently, an increase of c.a. 1.5 b/s/Hz in capacity for the same SNR value.
6.2.1.2 MIMO capacity using a standardized channel model and measured radiation patterns

The Kronecker model used above assumes independence between transmitter and receiver correlation. This assumption simplifies the channel modelling and allows independent optimization at transmitter and receiver sides, thus making it attractive for evaluation of e.g. algorithm and transceiver schemes via simulation. However, this approach neglects the correlation terms across the link, which may be relevant in certain scenarios [52].

A more realistic channel model is the 3GPP-3GPP2 Spatial Channel Model (SCM) [10]. It is a propagation-based model, which includes physical parameters in the simulation but also with a stochastic basis, since some of them are obtained as random variables. Thus, the advantages of both types of channel models (physically-based channels and stochastic models) are obtained. A drawback in our study is that this model is mainly intended for outdoor communications (macrocell and microcell scenarios are considered), so some parameters should be changed in order to account for indoor characteristics. Moreover, the channel model does not account for elevation spectrum, thus it is a 2-D channel model. We may note that there is a standardized MIMO channel model for WLAN indoor applications [139], but it is based on the Kronecker model and its assumptions, thus suffering from the same drawbacks. Therefore, the 3GPP model was preferred for comparison purposes.
The 3GPP-3GPP2 MIMO channel model defines the channel at two different layers: link-level and system-level. The former one defines parameters to be used when simulating a single link (one base station, one mobile station) such as the number of paths, the spatial characteristics per path, etc. It is an extension (to spatial domain) of the fixed tap-delay channel models specified in ITU-R Rec. M.1225. Since it does not consider multiple users, nor multiple snapshots or base stations, it is recommended to use the link-level model for calibration purposes only. Therefore, we have used the system-level model.

The principle of the SCM system-level channel follows some of the COST 259 recommendations. The model is a ray-based one where a subset of the parameters are stochastic. No specific antenna topologies are enforced, which gives freedom for selecting the antenna array configuration. The number of paths is a fixed parameter, being 6 paths, where each one consists of 20 sub-rays. The physical interpretation is that each path is the last interaction with a cluster of 20 scatterers, which has been found to be a proper number for realistic cases in mobile communications systems with not too broad band (such as 3G). In this model the simulation is carried out as a sequence of “drops”. At the beginning of a drop, for each mobile station, antenna orientation and gain parameters are fixed. Then composite rms delay spread, angle spread, and shadowing parameters are drawn from distributions functions with predefined parameters, thus being obtained as random variables with a certain distribution. Each path is characterized with a power and delay value, consisting on several sub-rays. Each sub-ray’s amplitude varies with a fast-fading characteristic, according to the speed of the MS. The path loss is simulated by using previous non-spatial channel models, such as COST231 Hata for macrocell, and so on. The parameters are regenerated randomly after each drop. Large-scale channel parameters, such as angle spread, delay spread, shadowing stay constant during a drop.

Figure 6-36 shows a geometrical description of the model for a BS-MS pair.

The meaning of the parameters depicted above is briefly summarized in the next list:
Chapter 6

\( \Omega_{BS} \): BS antenna array orientation, defined as the difference between the broadside of the BS array and the absolute North (N) reference direction.

\( \theta_{BS} \): LOS AoD direction between the BS and MS, with respect to the broadside of the BS array.

\( \delta_{n,AoD} \): AoD for the nth \((n = 1 \ldots N)\) path with respect to the LOS AoD \( \theta_0 \).

\( \Lambda_{n,m,AoD} \): Offset for the nth \((m = 1 \ldots M)\) subpath of the nth path with respect to \( \delta_{n,AoD} \).

\( \theta_{n,m,AoD} \): Absolute AoD for the nth \((m = 1 \ldots M)\) subpath of the nth path at the BS with respect to the BS broadside.

\( \Omega_{MS} \): MS antenna array orientation, defined as the difference between the broadside of the MS array and the absolute North reference direction.

\( \theta_{MS} \): Angle between the BS-MS LOS and the MS broadside.

\( \delta_{n,AoA} \): AoA for the nth \((n = 1 \ldots N)\) path with respect to the LOS AoA \( \theta_{0,MS} \).

\( \Lambda_{n,m,AoA} \): Offset for the nth \((m = 1 \ldots M)\) subpath of the nth path with respect to \( \delta_{n,AoA} \).

\( \theta_{n,m,AoA} \): Absolute AoA for the nth \((m = 1 \ldots M)\) subpath of the nth path at the MS with respect to the BS broadside.

\( v \): MS velocity vector.

\( \theta_v \): Angle of the velocity vector with respect to the MS broadside: \( \theta_v = \text{arg}(v) \).

Some of the parameters above are user-selected (as MS velocity, number of elements in the arrays…), while others are directly obtained from predefined distribution functions (as angle spread…) or indirectly obtained from randomly selected parameters (as position of MS, which defines LOS angle of departure, etc…). Explaining in detail how to generate and define each of the parameters above is out of the scope of this thesis, but the main ones regarding the specific studied case will be commented later. Interested readers may consult more details of the MIMO channel implementation in [10].

For an \( S \) element BS array and a \( U \) element MS array, the channel coefficients for one of \( N \) multipath components are given by an \( S \)-by-\( U \) matrix of complex amplitudes, \( H_n(t) \), \( n = 1 \ldots N \). The \((u,s)\)th component \((s = 1, \ldots, S; u = 1, \ldots, U)\) of \( H_n(t) \) is given by:

\[
 h_{u,s,n}(t) = \sqrt{\frac{P_n \sigma_{SF}}{M}} \sum_{m=1}^{M} \left( G_{BS} \left( \theta_{n,m,AoD} \right) \exp\left( jk \delta_{s} \sin\left( \theta_{n,m,AoD} \right) + \Phi_{n,m} \right) \times \right) \left( G_{MS} \left( \theta_{n,m,AoA} \right) \exp\left( jk \delta_{u} \sin\left( \theta_{n,m,AoA} \right) \right) \times \exp\left( jk \| v \| \cos\left( \theta_{n,m,AoA} - \theta_v \right) \right) \right)
\]  

(6-14)

where the main parameters are:

\( P_n \): power of the nth path.

\( \sigma_{SF} \): lognormal shadow fading, applied as a bulk parameter to the nth path for a given drop.
Some of the parameters are user-defined (as BS and MS antenna gains, \(d_s\) and \(d_u\)). The MS speed may be directly defined by user (as a fixed parameter) or drawn as a random variable with a predefined distribution function. The angular parameters (as \(\theta_{n,m,\text{AoD}}\), \(\theta_{n,m,\text{AoA}}\)) are obtained from predefined distribution functions, which likewise depends on several parameters as angular spread, angular offsets, etc…). They also depend on the considered environment and are specified in the technical report [10]. The average power of each path \(P_n\) is computed as an exponentially decaying function, fulfilling a certain power delay profile. The phases \(\Phi_{n,m}\) are drawn from a uniform 0 to 360 degree distribution. Again, a much more complete explanation of how all the parameters above are computed can be found in [10].

Three environments are considered in the SCM:

a) Suburban macrocell (approximately 3Km distance BS to BS)  
b) Urban macrocell (approximately 3Km distance BS to BS)  
c) Urban microcell (less than 1Km distance BS to BS)  

The type of selected environment varies the initial value of some parameters, and also the way some values are computed (for example, the path loss or the angular spread). Table 5.1 in [10] gives recommendations for how to compute each of the parameters above, for the three considered environments.

Other options in the system are: considering dual-polarized arrays, including far scatterer cluster, including line of sight, or simulating urban canyon. For the sake of
summarizing, they are not detailed here in detail. However, since the realized simulations used the dual-polarized option, we will briefly explain its main ideas.

When multi-polarized antennas are used, or when multiple polarization is to be considered, 2 channel coefficients are required for each Tx-Rx pair: co-polarized and cross-polarized one. The SCM describes the multi-polarized channel model by defining or including new parameters: the cross polar discrimination XPD and its inverse parameter, \( r = 1/XPD \). Also the antenna radiation pattern is divided into two response: the response for vertically polarized components and the response for horizontally polarized components. Although the model defines copolar and crosspolar polarizations as vertical and horizontal, other orthogonal bases could be used (right circular polarization and left circular polarization, +45º slanted polarization and -45º slanted polarization…). We have used the orthonormal basis \( \{\hat{\Theta}, \hat{\phi}\} \), which in the horizontal plane (which is the simulated plane in the 2-D 3GPP spatial channel model) is equivalent to vertical-horizontal basis.

The channel coefficients for the multi-polarized channel are computed as:

\[
h_{n,m}(t) = \frac{P}{M} \sum_{n=1}^{N} \left( \begin{array}{c} \chi_{n,m}^{(v)}(\theta_{n,m,AoD}) \exp\left(j\Phi_{n,m}^{(v,v)}\right) \\
\chi_{n,m}^{(h)}(\theta_{n,m,AoD}) \exp\left(j\Phi_{n,m}^{(h,h)}\right) 
\end{array} \right) \times \\
\left( \begin{array}{c} \sqrt{r_{n1}} \exp\left(j\Phi_{n,m}^{(v,h)}\right) \\
\sqrt{r_{n2}} \exp\left(j\Phi_{n,m}^{(h,v)}\right) 
\end{array} \right) \times \exp\left(jk\left|\cos(\theta_{n,m,AoD} - \theta_{h})\right| t\right)
\]

(6-15)

where the new parameters are defined as:

- \( \chi_{n,m}^{(v)}(\theta_{n,m,AoD}) \): BS antenna complex response for the vertically polarized component.
- \( \chi_{n,m}^{(h)}(\theta_{n,m, AoD}) \): BS antenna complex response for the horizontally polarized component.
- \( \chi_{n,m}^{(v)}(\theta_{n,m, AoD}) \): MS antenna complex response for the vertically polarized component.
- \( \chi_{n,m}^{(h)}(\theta_{n,m, AoD}) \): MS antenna complex response for the horizontally polarized component.
- \( r_{n1} \): random variable representing the power ratio of waves of the nth path leaving the BS in the vertical direction and arriving at the MS in the horizontal direction (v-h) to those leaving in the vertical direction and arriving in the vertical direction (v-v).
- \( r_{n2} \): random variable representing the power ratio of waves of the nth path leaving the BS in the horizontal direction and arriving at the MS in the vertical direction (h-v) to those leaving in the vertical direction and arriving in the vertical direction (v-v). The variables \( r_{n1} \) and \( r_{n2} \) are i.i.d.
- \( \Phi_{n,m}^{(v,x)} \): phase offset of the mth subpath of the nth path between the x component (either the horizontal h or vertical v) of the BS element and the y component (either the horizontal h or vertical v) of the MS element.

A MatLab implementation of the 3GPP SCM [278] was used to compute the MIMO channel matrix \( H_{n}(t) \) for the two implemented arrays (compact array and reference array) from the measured radiation patterns. In addition, results for an ideal array with isotropic elements
spaced \( d = \lambda / 2 \) were also computed. The wideband channel matrix \( H(t, \tau) \) for each antenna array was obtained as output, being \( H(t, \tau) \) the combination of the \( N \) multipath components \( (N = 6) \), where \( \tau \) is the time delay associated to each path. In order to compare results with other evaluation methods shown here, the equivalent narrowband channel matrix was computed for the center frequency, \( f_0 = 1766 \) MHz, as:

\[
H(t) = \sum_{n=1}^{N} H(t, \tau_n) \exp(-j2\pi f_0 \tau_n) \quad (6-16)
\]

In order to compute the channel coefficients including the antenna arrays effect, the measured radiation patterns were included in the channel simulations, by using expression (6-16) and the complex responses for the two orthogonal polarizations \( \hat{\theta}, \hat{\phi} \). The option for including dual polarization was switched on, so the cross-polarized channel was simulated. The arrays under test were evaluated in an indoor and outdoor-indoor environment (specifically, the scenario explained in Chapter 5). Thus, the urban microcell environment was selected for evaluating the compact array, since it is the closest one to the indoor case (small distances between BS and MS, higher number of scatterers, larger angle spread, LOS situations...). Therefore, regarding other parameters to be selected and computed, we have mainly used the recommended values and expressions for the urban microcell environment.

However, we have proposed an adjustment of the parameters regarding angle of departure, in order to account for the more dispersive and multipath-rich indoor environment where the system was tested. The standard channel model computes the per-path angle of departure from the base station as a random variable with different distribution function depending on the environment. Table 6-VI summarizes the recommended values for this parameter, as stated in Table 5.1 of [10]:

<table>
<thead>
<tr>
<th>Channel Scenario</th>
<th>Suburban Macro</th>
<th>Urban Macro</th>
<th>Urban Micro</th>
</tr>
</thead>
<tbody>
<tr>
<td>BS per-path AoD distribution function</td>
<td>( \eta(0, \sigma_{\text{AoD}}^2) )</td>
<td>( \eta(0, \sigma_{\text{AoD}}^2) )</td>
<td>( U(-40^\circ, 40^\circ) )</td>
</tr>
</tbody>
</table>

Table 6-VI. Distribution function for AoD calculation in SCM simulations, as defined in [10].

We note that these functions are well suited to outdoor BS locations, where the angles of departure of paths are expected to be confined in a narrow beam. For suburban and urban macro environments, the AoD is expected to follow a Gaussian distribution, with mean angle spread \( E[\sigma_{\text{AoD}}] \) of 5º for suburban (less multipath richness) and 8º or 15º for urban (a bit higher multipath rich environment). For urban micro scenario, a uniform distribution is assumed in the standard SCM, with maximum angle of departure \( AoD_{\text{max}} = 40^\circ \). This value is reasonable for dense building (scatterer) outdoor areas, where a dominant angular direction (streets) with wide angle spread (total maximum angle interval for AoD is 80º) can be assumed. However, for our case under study, the receiver is always assumed to be in indoor location, and the transmitters (BSs) are located either in indoor position or in outdoor but close to the building under study.
That means that the correlation between transmitters may be assumed lower than for the urban micro case. In order to account for this, the maximum angle of departure was selected as:

$$\text{AoD}_{\text{max}} = 180^\circ$$

This way, the total interval for obtaining the AoD parameter is 360°. Than means that omnidirectional antennas are considered for the transmitter, and equal probability of finding scatterers in all directions are assumed. Since we aimed at evaluating the antennas in the MS (receiver) side, no specific radiation pattern or radiation features were selected for the transmitter, so omnidirectional radiation was the most reasonable option here.

Once the $H$ narrowband channel matrix of (6-16) was computed, it was used to compute the achievable capacity by using eq. (6-9). An SNR = 10 dB was selected, and 1000 independent trials or “drops” were simulated. For each drop, 100 channel samples were simulated. It is important to note that, in order to fairly compare the antennas performance, the same random values were used for obtaining the H matrix including each of the radiation patterns during a drop, that is: the same correlation values, fast fading, path loss, MS position, etc, are seen by all the antenna arrays. A small cell radius was selected (100 m) in order to simulate small distances between Tx and Rx in indoor scenarios, and a slow pedestrian speed was chosen (1 m/s), simulating a person inside a building.

Table 6-VII summarizes the main parameters that were used in the simulations.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of drops</td>
<td>1000</td>
</tr>
<tr>
<td>Number of samples per drop</td>
<td>100</td>
</tr>
<tr>
<td>Scenario</td>
<td>Urban Micro</td>
</tr>
<tr>
<td>MS speed</td>
<td>1 [m/s]</td>
</tr>
<tr>
<td>Number of Tx</td>
<td>4</td>
</tr>
<tr>
<td>Number of Rx</td>
<td>4</td>
</tr>
<tr>
<td>Cell radius</td>
<td>100 [m]</td>
</tr>
<tr>
<td>SCM option</td>
<td>polarized</td>
</tr>
<tr>
<td>Normalized Pw polarization</td>
<td>Yes</td>
</tr>
<tr>
<td>$\text{AoD}_{\text{max}}$</td>
<td>180°</td>
</tr>
</tbody>
</table>

Table 6-VII. Summary of parameters used for simulation with 3GPP SCM.

Parameters that are not specifically given in this table were taken with no changes from the suburban micro environment parameters in Table 5.1 of [10].
Figure 6-37 shows the cumulative distributed function (CDF) of the simulated capacity. We observe that the reference array shows a higher capacity than the compact array (AUT), which agrees with the results obtained directly from the evaluation with the Kronecker model and the measured radiation patterns. As in that study, a better performance is obtained for the array with omnidirectional elements. This result is reasonable, taking into account that the AUT shows a lower radiation efficiency, while the ideal isotropic case assumes optimum radiation efficiency $\eta_{iso} = 1$. The ideal case of an i.i.d channel is also plotted, for comparison reasons. As expected, the achievable capacity is always lower for the 3GPP SCM than for the i.i.d. channel, since the latter gives uncorrelated transmit and receive signal, and thus optimum performance for no CSI at transmitter.

![Figure 6-37. CDF of capacity for the AUT and the reference array, obtained with 3GPP2 SCM and measured radiation patterns. $AoD_{max} = 180^\circ$.](image)

Also the average capacity as a function of SNR has been computed, and it is shown in Figure 6-38. When compared with results obtained with the first proposed method (Kronecker model and measured radiation patterns) we note that this method gives quite lower capacity values than the former one. This is a logical result: while the first assumption of Kronecker model is the use of an i.i.d. matrix, $G$, and only the antenna correlation reduces the performance from the ideal i.i.d. channel, the 3GPP spatial channel model is based on finite multiple paths addition, which involves higher correlation values at first place (even for ideal isotropic antennas) and thus lower capacity.
Figure 6-38. Average capacity for different SNR values, for the AUT and the reference array. The 3GPP spatial channel model and measured radiation patterns are used.

The improvement in SNR for a fixed capacity of 12 b/s/Hz when using the reference array instead of the antenna under test is now lower $\Delta \text{SNR} = 0.95$ dB, that is, 0.4 dB lower than the results obtained with the Kronecker model. For a fixed SNR of 15 dB, the improvement in capacity when comparing both studied arrays is $\Delta C = 0.8$ b/s/Hz. Consequently, we conclude that, the first method slightly exaggerates the performance differences between antennas. When analysing absolute values, we observe that the curve is moved approximately 2.4 dB and 2.8 dB to the right for the AUT and the reference array, respectively, if results from the 3GPP SCM and Kronecker model are compared. Clearly, the first method overestimates the arrays performance, when compared with results drawn from a more realistic channel model.

If the $A_{oD_{\text{max}}}$ recommended by the standard ($A_{oD} = 40^\circ$) was used instead of the proposed one, the obtained capacity would be lower, since that scenario would assume a lower level of diversity and thus higher correlation between transmit antennas. Figure 6-39 shows the capacity results with $A_{oD} = 40^\circ$, for comparison reasons. We observe that the achieved capacity for a fixed SNR value is much lower than in the case of of $A_{oD} = 180^\circ$, for the two arrays (approx. 1.5 b/s/Hz lower). Therefore, for estimating the performance of MIMO arrays in an indoor environment, using a high $A_{oD}$ is preferred.
6.2.2 Evaluation using measured radio channel responses

Evidently, the ideal way to evaluate an antenna array for MIMO application is by measuring its performance with a MIMO prototype system. Since having a full prototype working available may be too ambitious, a simpler version such as a MIMO test-bed or channel sounder may be good enough, if appropriate channel measurements in different environments relevant to a real system are accomplished. Although time consuming and inflexible, this method can be considered as the baseline to which comparing other methods of evaluation.

A channel measurement campaign was undertaken with a MIMO test-bed including the two antenna arrays under evaluation. The measurement campaign was conducted in the S3 building and surroundings, in the KTH campus, Stockholm. A narrow-band MIMO DSP-based test-bed developed in the Department of Signals, Sensors and Systems, KTH, was used for data acquisition. The details about the test-bed, the measurement setup, the signal processing and the measured scenario were already described in chapter 5, as well as the detailed results from a propagation perspective. Here we give a brief overview on the measurement setup, and details about the antenna arrays that were skipped in previous chapter. The results presented here are referring to comparison and evaluation of the two antenna arrays, and were not presented in chapter 5.
In order to evaluate their performance, the compact array and the reference array were used as receive antennas of the system. They were closely located on the trolley performing as MS (as shown in Figure 6-40), and the impinging radio waves were simultaneously received by means of 8 independent RF chains and an 8-port digital acquisition system. This way, the radio channel observed by the two arrays was very similar, so their performance could be compared in a fair way. For all the measurements, the transmitter module was placed at a fixed location, while the receiver module was carried along several indoor routes, as explained in chapter 5. The antennas were mounted so that the $xy$-plane coincided with the horizontal plane. Two locations were considered for the transmitter: an indoor and an outdoor one. A total of 22 routes were measured and repeated for each transmitter location. The receiver routes covered 3 different floors with similar plans, at pedestrian speed (ca. 1 m/s) recording 100 $H$ matrices per second. The measurements included situations of line of sight (LOS) and non-line of sight (NLOS), as well as routes inside the offices. More details of the measurement setup can be consulted in chapter 5.

![Figure 6-40. The two arrays under study, mounted on a trolley (MS).](image)

From the measured $H$ matrices, the channel capacity was computed for the two opposite cases: no CSI at transmitter (eq. (6-9) is used) and total CSI at transmitter (eq. (6-8) and SVD method is used). Two levels of SNR were considered: low SNR (SNR = 10 dB) and high SNR (SNR = 30 dB). In the capacity calculations, the channel matrix is assumed to be normalized, so it does not account for path losses (they are included in the $\rho$ parameter). However, different routes or scenarios will involve different path losses and thus different received power. For example, in a LOS case higher power is received than in an NLOS case with equivalent BS-MS distance. From a propagation perspective, where power control is not assumed, it can be more interesting to assume a certain average SNR over the whole measured routes in order to account also for shadow fading, slow fading and path losses. In this case, the $H$ matrix is not instantaneously normalized, as in some of the results presented in chapter 5. However, from a
system point of view, assuming power control is reasonable, and then $H$ matrix should be normalized to consider a somehow fixed, local SNR value. In order to obtain a fair comparison between antenna arrays, we have chosen to normalize the measured channel matrices for both the AUT and the reference array, $H_{AUT}(t)$ and $H_{REF}(t)$, to the same instantaneous value, so that the local SNR value at the receivers is fixed. The average value of the Frobenius norm of $H$ over a sliding window of 100 samples (corresponding to roughly 1 m of distance) was used as normalization value, similar to what was proposed in [63], so the fast fading effect is not removed. The normalized channel matrix for time instant $t$ is then:

$$H_{Norm,A}(t) = \frac{H_A(t)}{\left( \frac{1}{2} E \left( \|H_{REF}(t)\|_{Frob}^2 + \|H_{AUT}(t)\|_{Frob}^2 \right) \right)^{1/2}} = \frac{2N_{Tx}N_{Rx}}{E} \left( \sum_{i=1}^{N_{Tx}} \sum_{j=1}^{N_{Rx}} (h_{j,i,REF}(t)h_{j,i,REF}^*) + (h_{j,i,AUT}(t)h_{j,i,AUT}^*) \right)$$

(6-17)

where $\| \cdot \|_{Frob}$ denotes Frobenius norm, $E\{\cdot\}$ is the expectation operator over the sliding window, $(\cdot)^*$ denotes conjugate and $H_A$ is the channel matrix for the reference array ($H_{REF}$) or the array under test ($H_{AUT}$). Since the channel was measured simultaneously for both arrays, we expected similar slow fading. However, due to their slightly different position on the trolley, some differences in shadow fading and slow fading were observed, as we will see below.

The CDF of the capacity was calculated from all the measured routes, including indoor and outdoor transmitter location, in order to get a result as much general as possible. The results are shown in Figure 6-41 a) and b) for the AUT and the reference array, respectively. We observe that the AUT exhibits a 1 to 2 bits/s/Hz lower capacity than the reference array for both SNR scenarios and CSI cases. As a comparison, the mean capacity from for an ideal 1×1 (SISO) and 4×4 system under i.i.d. Rayleigh fading, no CSI, and 10 dB average SNR is 2.9 and 10.9 bits/s/Hz, respectively. Thus, the AUT and REF arrays in the LOS and NLOS office environment, with a mean capacity between 8.7 and 9.7 bits/s/Hz, achieve roughly three times the SISO capacity. We may also notice that the capacity differences between no CSI and CSI at transmitter cases are smaller for higher SNR scenarios, as expected.

Also the capacity as a function of SNR was computed, Figure 6-42, in order to compare results with the previous methods. We see that the array differences (1.25 dB) are similar to the ones obtained with the Kronecker model and measured patterns (1.35 dB for the same fixed average capacity). However, regarding absolute values, both curves are moved to the right with respect to the first method’s result. According to the results from measured channel response, roughly 1 dB more is required for getting an average capacity of 12 b/s/Hz. Thus, the first method overestimates the achievable capacity with both arrays. The opposite is true for the second method (3GPP SCM) and measured radiation pattern: it underestimates the achievable
capacity, for both arrays. Nevertheless, the differences between arrays (ΔSNR and ΔC) are kept approximately constant for the first and the third method.

![Graph](image1)

a) CDF of capacity, SNR = 10 dB

![Graph](image2)

b) CDF of capacity, SNR = 30 dB

Figure 6-41. Comparison of CDF of capacity for the two arrays with and without TX CSI using all the measurement routes.

![Graph](image3)

Figure 6-42. Average capacity for different SNR values, for the AUT and the reference array. measured channel response is used.

As a measure of diversity, the eigenvalues of the correlation matrix \( R = H H^H \) have been calculated for all the routes. The eigenvalue decomposition of the channel matrix and its interpretation in MIMO communications can be consulted in chapter 2, section 2.2.2. Figure
6-43 shows the eigenvalue analysis and; we see that only two of the eigenvalues are significant, as expected in conventional scenarios (not especially multipath reach). We also present the CDF of their dispersion rate, $\Lambda$, i.e. the ratio of their geometric and arithmetic means, which is computed as:

$$\Lambda(t) = \frac{m_{\lambda,\mathrm{d}}(t)}{m_{\lambda,\mathrm{a}}(t)} = \frac{1}{K} \left( \prod_{k=1}^{K} \lambda_k(t) \right)^{1/K}$$  \hspace{1cm} (6-18)

where $\lambda_k$ is the $k$-th eigenvalue and $K$ is the number of non-null eigenvalues. The dispersion rate provides a power-invariant measure of the dispersion of the channel eigenvalues [279]. From the MIMO point of view, a higher dispersion of channel eigenvalues implies that a higher diversity is available.

![Figure 6-43. CDF of eigenvalues (top) of the channel covariance matrix, $\mathbf{H}_H^H$, and eigenvalue dispersion (bottom), for all the measured routes.](image)

In Figure 6-43 we see that the eigenvalues distribution (CDF) is very similar for both arrays, except for the first (most relevant) eigenvalue, which is higher for the reference array. This may be due to the higher antenna efficiency of reference array, which implies higher received power and hence higher values of $\mathbf{H}$ and eigenvalues. However, it is interesting to note that the second eigenvalue $\lambda_2$ is very similar for both arrays (even slightly higher for the AUT), which is translated in a slightly higher dispersion rate, in spite of the lower capacity offered by
the AUT. Since the differences in dispersion rate are small, we conclude that for a general indoor scenario, the main difference in performance between both arrays is the lower antenna efficiency of the AUT.

The results above show the overall performance using all the measured routes. Therefore, they show the overall performance of the antenna arrays. However, one advantage of using channel measurements to evaluate MIMO antennas performance is the possibility of studying their behaviour in different environments and situations. Moreover, since multiple channel measurements were taken along the routes, the achieved capacity can be studied as a function of RX location. Only the more interesting results are shown here; a detailed study of all the routes and comparison of instantaneous capacity for the two arrays can be found in [280].

In Figure 6-44 the CDF of capacity for the route along the north corridor of fourth floor is presented, for both indoor and outdoor transmitter location and the two antenna arrays. An SNR = 10 dB is assumed and no CSI at transmitter. In Figure 6-45 the floormap of the forth floor is shown, as well as the two studied TX locations and the covered route for this example. We see that for indoor TX location the route implies a LOS propagation, while for the outdoor TX location it is mainly an NLOS route (except for the small amount of power entering through the windows).

![Figure 6-44. CDF of capacity obtained in the route on 4th floor, north corridor, for indoor (dashed line) and outdoor (solid line) TX location, for the two antenna arrays. The REF is shown in red and the AUT in blue.](image-url)
The difference in average capacity between the reference array and the AUT is highlighted in the figure, for the two TX locations. For the outdoor TX location, the reference array again shows a better performance than the AUT, as expected from previous results. However, for the indoor case the improvement is not so clear, and the AUT exhibits a slightly better mean capacity than the reference array. This may be explained by the higher radiation pattern diversity offered by the AUT in this scenario, which is noticeable for small incidence regions in azimuth, as in the corridor-LOS scenario. Hence, this result agrees with the one shown in Figure 6-33, obtained from covariance analysis.

Figure 6-46 and Figure 6-47 show the capacity (3-D curves) averaged over 0.5 s as a function of location with indoor and outdoor transmit antennas, respectively, for the two arrays. Furthermore, the color map plotted on the floor represents the ratio of received power $\Delta P$ for the reference array to the AUT, that is:

$$\Delta P(t) = \frac{P_{\text{REF}}(t)}{P_{\text{AUT}}(t)} = \frac{E \left\| H_{\text{REF}}(t) \right\|_{Frob}^2}{E \left\| H_{\text{AUT}}(t) \right\|_{Frob}^2}$$  \hspace{1cm} (6-19)
Figure 6-46. Capacity as a function of local RX position, for AUT and reference array. Indoor TX location (LOS case) and SNR = 10 dB. 2D color map represents power ratio REF/AUT (in dB).

Figure 6-47. Capacity as a function of local RX position, for AUT and reference array. Outdoor TX location (NLOS case) and SNR = 10 dB. 2D color map represents power ratio REF/AUT (in dB).
It seems clear from Figure 6-46 and Figure 6-47 that the statistics of the achievable capacity are quite dependent on the scenario, especially when comparing LOS and NLOS situations. For the NLOS case, we observe small variations of capacity with location for both the AUT and the reference array, while for LOS cases the capacity variations with position are higher and quite different for the two arrays, as could be expected after results shown in Figure 6-33. The higher variance may be explained by the specific scatterers and elements along the route, partially blocking the signal in the LOS case, which is then more sensitive to the specific RX location.

We may note that the power ratio varies strongly for the LOS case. It is also interesting to notice that for the NLOS scenario the received power is slightly higher for the reference antenna than for the AUT, which agrees with the radiation efficiency difference between both arrays, shown in previous results with other evaluation methods.

As a conclusion, the analysis of the capacity shows that the reference array offers in general a slightly better result than the compact array, due to its higher antenna efficiency. However, under certain specific scenarios such as LOS the improvement of reference array over AUT is not so clear. The analysis of maps representing capacity as a function of location and power ratio has shown that the variance is higher in the LOS case for both arrays. Therefore, it has been proved that this type of representation is of interest to study specific scenarios.

### 6.2.3 Evaluation with a reverberation chamber

As an alternative to time-consuming channel measurements or 3-D radiation pattern measurements, a uniform multi-path environment can be generated artificially in a reverberation chamber and it can be used to evaluate antenna arrays for MIMO.

A reverberation chamber is a controlled environment to optimally create a rich multipath environment. Multiple radio waves are transmitted, and by using several metal stirrers the electromagnetic waves are reflected and diffracted several times. Thus, the main principle is to create a multi-wave environment, as opposed to the operation of an anechoic chamber, based on receiving a single wave coming in LOS from the transmitter and avoiding any reflection.

A reverberation chamber provides a statistically repeatable environment for characterizing MIMO antennas, by measuring $H$ matrices to compute the theoretical capacity. Since the antennas are measured in a controlled environment, it is straightforward to compare the performance of different antennas with this method. The measurement technique is briefly described below. A detailed explanation can be found in [167].

The reverberation chamber consists of a small chamber made of metal walls, a platform to locate the antenna under test, some fixed antennas mounted on the walls and several metal stirrers mechanically controlled. Firstly, the MIMO array under test is located inside the
reverberation chamber, so that it is at least 0.5λ far from metal walls and stirrers, in order to avoid significant disturbance of the array radiation characteristics. Also a reference antenna (with known radiation efficiency) is located inside the reverberation chamber. We connect one of the array ports to the network analyzer output port, and terminate all the other ports and the reference antenna in 50Ω. With this setup, the S-parameters between the port and the three wall-mounted antennas are collected for all positions of the platform and mechanical stirrers and for all frequency points. The measurement procedure is then repeated for every antenna port for exactly the same stirrer positions and position of the array inside the chamber. The complex transmission coefficients $S_{ij}$ between the connected port and each of the three fixed wall antennas, as well as the reflection coefficients $S_{ii}$ of each of the wall antennas and $S_{jj}$ of the array port, are stored for every stirrer position and frequency point. Finally, the reference antenna is connected to the network analyzer and the same measurements as for the array are performed, in order to obtain a reference measurement. These results, together with the known radiation efficiency of the reference antenna are used to normalize the measured $S_{ij}$ parameters. The normalized $S_{ij}$ values represent the channel matrix coefficients of a rich-multipath channel, which can be used to compute the channel capacity from a MIMO perspective.

Figure 6-48. Drawing of a reverberation chamber and measurement scheme (from [167]).

By using this measurement technique, the AUT and the REF arrays were characterized. Since the available reverberation chamber has 3 wall-mounted antennas, a 3×4 MIMO channel was created in the chamber, and the arrays were evaluated for this case by computing the capacity using eq. (6-9) (no CSI is assumed at TX). The result is seen in Figure 6-49, where we have also represented the theoretical capacity from an ideal 3×4 i.i.d. channel. Figure 6-50 shows photographs of the reference array and the AUT mounted in the reverberation chamber.
Figure 6-49. Calculated capacity for a 3×4 MIMO system using the channel data from a reverberation chamber, and comparison with ideal i.i.d. 3×4 channel (not CSI at TX).

As expected from the low correlation calculated in previous sections, when comparing the two antenna arrays the capacity curves basically differ as the signal strength varies, due to different radiation efficiencies. The difference in capacity between both arrays is 1 bit/s/Hz. Since the mean capacity is approximately linear to the SNR in dB (for medium and high SNR values), we may equivalently express this as 1.2 dB difference at 10 b/s/Hz of capacity.

Figure 6-50. Studied arrays mounted in the reverberation chamber where measurements were gathered.
As compared with the ideal i.i.d. channel, we see that the reverberation chamber measurement slightly overestimates the capacity results, so the capacity is even a little higher for the reference array than for an ideal i.i.d. channel where. For a rich multi-path environment with no preferred direction of arrival or departure of waves, the i.i.d. channel results should represent an upper bound of theoretical capacity, since in this scenario the antennas are assumed to be omnidirectional with 100 % antenna efficiency. The better performance of the reference array with respect to i.i.d. channel may be due to small errors in the normalization step when measuring in a reverberation chamber. Also, the reference array consists of $\lambda/4$ monopoles, which radiate omnidirectionally in the horizontal plane but following a sine distribution in the vertical plane. As shown in Figure 6-28, they present a maximum in gain at an angle slightly higher than the horizontal plane. Therefore, if there is a slightly higher concentration of waves in the upper semi sphere of space (which could be the case, since transmit antennas are located on top of the chamber), the reference array may give better results than ideal isotropic antennas, especially considering that the radiation efficiency of the reference array is very high (close to 95%).

### 6.2.4 Comparison of methods for evaluation of MIMO antenna arrays and conclusions

We have presented 4 methods to evaluate antenna arrays for MIMO systems and their performance regarding MIMO capacity, namely:

- **Method 1**: use of measured radiation patterns and the Kronecker model.
- **Method 2**: use of measured radiation patterns and polarized 2-D spatial channel model from 3GPP-3GPP2, with some modifications to account for indoor scenario.
- **Method 3**: performing radio channel measurements with a MIMO test-bed.
- **Method 4**: measurements taken with a reverberation chamber.

Each method presents some advantages and disadvantages when compared with the other ones, which we highlight below.

As a general idea, each method represents a way to compute, estimate or measure the radio channel coefficients $H$ including the effect of the antenna array, and the capacity is obtained from the $H$ matrix by using eq. (6-9) for no CSI at TX, or the more general expression (6-8) for cases with CSI at TX. Thus, the differences in performance between the two arrays should be due only to the differences between the two antenna arrays (in radiation pattern, antenna efficiency, coupling between elements and so on...). A good evaluation method should be general enough to represent general scenarios (such as “office-like scenarios”, “corridor scenarios”, “indoor-outdoor scenarios”...), but it should also provide results for specific cases (as in “LOS situation in a corridor” or “NLOS situation and BS outside of..."
building”) where an antenna array could provide a better performance than in other situations. Of course, required equipments and resources are aspects to be considered, as well as the complexity and time requirements of each method.

In order to easily compare the four methods, the mean capacity as a function of local SNR at the receiver has been represented in Figure 6-51 for methods 1, 2 and 3, for a 4×4 MIMO channel. Method 4 is not included, since the measurements in the reverberation chamber were restricted to the use of up to 3 transmitters, due to hardware constraints.

![Average capacity as a function of SNR, 4×4 MIMO](image)

Figure 6-51. Comparison of methods to evaluate antenna arrays for MIMO systems. Capacity as a function of SNR, 4×4 MIMO. Blue has been used for the reference array, and red for the AUT.

We observe that the method 1 (Kronecker model and measured radiation patterns) is the most optimistic method. Compared to the results obtained from channel measures, method 1 overestimates the capacity results. This shows that assuming independence between transmit and receive sides yields to optimistic results regarding capacity, as also stated in [52]. Conversely, the 3GPP channel model underestimates the capacity results when compared to indoor channel measurements, being method 3 the one which obtains lower capacity values, even if an AoD$_{max} = 180^\circ$ was used. This is reasonable, since the 3GPP-3GPP2 SCM aims at simulating outdoor scenarios for macro or micro cell environments, where transmit correlation is usually higher than for indoor scenarios.
With the objective of also comparing results from method 4, the capacity as a function of SNR for a 3×4 MIMO channel has been computed for all the other 3 methods. For method 3, the forth TX antenna was not considered in the computation of capacity, that is, the last column of the measured $H$ was removed. For methods 1 and 2, only three TX ideal antennas were included in the simulation of the MIMO channel. Figure 6-52 shows the results for the 3×4 MIMO case.

![Graph showing average capacity as a function of SNR, 3×4 MIMO](image)

**Figure 6-52. Comparison of methods to evaluate antenna arrays for MIMO systems. Capacity as a function of SNR, 3×4 MIMO. Blue has been used for the reference array, and red for the AUT.**

For the 3×4 MIMO case, the results obtained with the 3GPP channel model are very similar to the ones obtained from direct channel measurements. However, we have to emphasize that the maximum angle of departure $AoD_{\text{max}}$ was modified from the standardized value ($AoD_{\text{max}} = 40^\circ$) to $180^\circ$, to account for realistic conditions in indoor environments. For the reverberation chamber, the obtained capacity is the highest one when compared with the other 3 methods. It gives similar results as method 1 (Kronecker and radiation patterns), thus overestimating the
capacity when compared with results obtained from channel measurements. This result is reasonable, since the reverberation chamber emulates an optimal multipath environment with lower correlation. The direct channel measurement may offer a more realistic result than the reverberation chamber method, but with the drawback of being a more complex method for evaluating the antennas. Despite this, the differences between both antenna arrays are quite similar for both methods. Thus, if we are only interested in comparing antenna arrays both methods may give similar results.

Regarding complexity, each method presents very different requirements with respect to required equipment and processing complexity (and thus time to process data and get results). Method 1 requires the knowledge of the 3-D radiation pattern of each antenna element, thus it is necessary to have an anechoic chamber available and to realise quite time consuming measurements to get the 3-D patterns. However, once those are known, computing the $H$ matrices and from them the covariance and capacity results is straightforward and requires very few computational effort. Method 2 only needs information from 2-D patterns, which reduces the requirements regarding the time to measure them. However, the channel simulations are more complex and require more time to compute $H$ and the capacity.

Method 3 is by far the most complex and time consuming of the presented methods. A MIMO testbed is required, with several transmitters and receivers working simultaneously. Furthermore, a quite extensive measurement campaign has to be realized in order to gather enough information and be able to get results that can be generalized and not only for a very specific location. For the specific case of the measurement campaign that was realized for this study, 2 full-days were required for setting-up and performing the measurements that have been presented in this chapter (the total measurement campaign was 4-day long, other results not presented here are presented in chapter 5). The big amount of collected information must be processed in order to estimate the $H$ matrices for each of the covered routes. Moreover, since the amount of estimated $H$ matrices is quite large, also the computation of the capacity takes much more time than in the case of using methods 1 or 2. The positive side of it is that the results represent a real scenario and give more information about different possible situations.

Finally, method 4 does not require complex measurement campaigns to be carried out, nor knowing the radiation pattern of the antenna elements. Nevertheless, it requires the use of a reverberation chamber, which is not so common and thus not available for everybody. The time to measure the created MIMO channel matrices in the chamber (including the array effect) is quite low, and the correction to take into account the normalization is quite fast and can be automatic. However, this method is the one that gives less information about the array performance, and its results overestimates the achievable capacity, as seen before.

In order to have a clear idea on the time requirements and computational complexity for each method, we show in Table 6-VIII the time that was required to evaluate the two studied
antenna arrays for the four presented methods. The same PC was used for all the computations (Pentium III @ 2.4 GHz, 512 MHz RAM), so the required time can be compared. The overall time was split into the different required tasks for each method, so a clearer overview can be obtained.

<table>
<thead>
<tr>
<th>Method 1: 3-D Radiation patterns and Kronecker model</th>
</tr>
</thead>
<tbody>
<tr>
<td>· Time to measure 3-D radiation pattern of all the elements (4 elements for each array, AUT and REF).</td>
</tr>
<tr>
<td>· Time to simulate $H$ matrices (1000 independent trials), and to compute capacity from $H$, as function of SNR, 25 SNR values.</td>
</tr>
<tr>
<td>$8 \times 30 \text{ min} = 240 \text{ min}$</td>
</tr>
<tr>
<td>$5.9 \text{ seconds (negligible)}$</td>
</tr>
<tr>
<td>TOTAL REQUIRED TIME (METHOD 1) $\sim 240 \text{ min}$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Method 2: 2-D Radiation patterns and 3GPP2 SCM</th>
</tr>
</thead>
<tbody>
<tr>
<td>· Time to measure 2-D radiation pattern of all the elements (4 elements for each array, AUT and REF).</td>
</tr>
<tr>
<td>· Time to simulate $H$ matrices (1000 “drops”, 100 samples per drop), and to compute capacity from $H$, as function of SNR, 25 SNR values.</td>
</tr>
<tr>
<td>$8 \times 10 \text{ min} = 80 \text{ min}$</td>
</tr>
<tr>
<td>$\sim 2 \text{ min}$</td>
</tr>
<tr>
<td>TOTAL REQUIRED TIME (METHOD 2) $\sim 82 \text{ min}$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Method 3: radio channel measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>· Time to perform channel measurement campaign</td>
</tr>
<tr>
<td>· Time to estimate $H$ matrices from measurements</td>
</tr>
<tr>
<td>· Time to compute capacity from all estimated $H$ matrices, as function of SNR, 25 SNR values.</td>
</tr>
<tr>
<td>$2 \text{ days} = 16 \text{ hours}$</td>
</tr>
<tr>
<td>$\sim 8 \text{ hours}$</td>
</tr>
<tr>
<td>$11.2 \text{ min}$</td>
</tr>
<tr>
<td>TOTAL REQUIRED TIME (METHOD 3) $\sim 24 \text{ hours 11 min}$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Method 4: measurements in a reverberation chamber</th>
</tr>
</thead>
<tbody>
<tr>
<td>· Time to measure $H$ matrices for the two antenna arrays and 3 transmitters ($3 \times 4 \text{ MIMO}$)</td>
</tr>
<tr>
<td>· Time to compute capacity from estimated $H$ matrices ($2550 \text{ H values}$), as function of SNR, 25 SNR values.</td>
</tr>
<tr>
<td>$1 \text{ hour}$</td>
</tr>
<tr>
<td>$11.2 \text{ sec (negligible)}$</td>
</tr>
<tr>
<td>TOTAL REQUIRED TIME (METHOD 4) $1 \text{ hour}$</td>
</tr>
</tbody>
</table>

Table 6-VIII. Comparison of required time to evaluate 2 4-element arrays for MIMO application, with different methods.
Some of the times are approximate (as time to measure the radiation patterns, since it depends on the type of measurement method and available anechoic chamber used…). We have not included the time to develop the required programs and software to realize the data processing, assuming that this is only once realized and can be used any time an antenna array has to be evaluated.

The required time for arrays evaluation is also presented in Figure 6-53, in the form of bar scheme, for the sake of showing a straightforward visualization.

![Time to evaluate antenna arrays](image)

**Figure 6-53. Comparison of required time to evaluate the two antenna arrays with the four proposed methods.**

As observed, the method based on radio channel measurements is the most time-consuming of all, with a great difference with the other ones. Method 4 is in principle the one that requires less time and processing effort, but it is necessary to use a reverberation chamber, which is not always available. Finally, methods 1 and 2 are quite similar in time requirements, and also in required equipment, since both of them are based in having access to anechoic chamber measurements to know the radiation pattern of the antennas.

To sum up the comparison of methods, Table 6-IX explicitly states advantages and disadvantages of each method. None of the methods can be selected as “the best” or the “worst” one, since this clearly depends on the needs regarding the information of interest, the features we want to know about the antenna array and also the available equipment and time needs.
### Method 1: 3-D Radiation patterns and Kronecker model

**Advantages**
- Easy to analyse different types of scenarios by changing field distribution (small or large angle spread, different distribution functions for angle spread…).
- It gives information focused more on the antenna and less on the channel.
- Fast and easy to compute (once 3-D radiation patterns are known).
- Low computational requirements.

**Disadvantages**
- It requires 3-D costly measurements.
- Less realistic (assumes TX-RX independent distribution).
- Generally, it overestimates the capacity results.

### Method 2: 2-D Radiation patterns and 3GPP2 SCM

**Advantages**
- This channel model is more realistic than the Kronecker model, especially for outdoor scenarios.
- It can be used for system-level simulation (multiple users and BSs).
- Less requirements on knowledge of radiation pattern.

**Disadvantages**
- Higher computation requirements.
- It is difficult to control the effect of the channel (many parameters to be selected…).
- It is less oriented to give information about the antenna effects, and more oriented to the channel effects.
- Generally, it underestimates capacity results.

### Method 3: radio channel measurements

**Advantages**
- It allows to obtain more information about the system (specific routes, LOS vs NLOS…).
- It allows capacity maps representation, study of coverage zones, etc…
- It is the most realistic method.

**Disadvantages**
- Very time and effort consuming.
- High computational requirements.
- It requires a very specific hardware (MIMO test-bed or channel sounder).
- It is not clear where to measure to obtain general results.

### Method 4: measurements in a reverberation chamber

**Advantages**
- It does not require knowledge of radiation pattern.
- It is a fast method.

**Disadvantages**
- It overestimates capacity results.
- It requires a reverberation chamber available.
- It does not give extra information (only H matrices, no knowledge about radiation patterns).

Table 6-IX. Main characteristics of the 4 proposed methods to evaluate antenna arrays for MIMO.
6.3 General conclusions, contributions and further research on antenna aspects for MIMO systems

One of the current open issues to be solved before seeing MIMO systems deployed in real networks, is how to implement multi-antenna terminals in the user terminal, which has some constraints as size, weight and batteries. To contribute to this issue, we have presented the design and implementation of a MIMO array suitable for realistic hand-held devices. Based on microstrip technology, a PIFA element was designed for the GSM high frequency (1800 MHz), where good impedance matching and bandwidth is obtained. Although it was not the main aim of the design, also a fair matching is obtained for the second resonant frequency, at the lower WLAN frequency (2450 MHz). The implemented antenna array consists of 4 of these PIFA elements, covering a small area which is typical for PDA terminals. The designed array was optimized to get low coupling between elements, as well as polarization diversity, provided that only a certain small area was available.

Firstly, the implemented compact array has been evaluated by means of conventional parameters. Regarding S parameters, a good impedance matching and bandwidth has been achieved, obtaining up to 9% of bandwidth for a reflection coefficient lower than 10 dB. The antenna efficiency has also been measured, being approximately of 70%. Also the radiation patterns were measured for each element of the compact array, showing an interesting pattern diversity, which could be useful in a MIMO scenario with low spatial diversity.

Despite its undeniable interest, conventional antenna parameters are not enough to fully characterize an antenna array from a MIMO perspective. To begin with, MIMO systems require several radiant elements to be closely placed, and hence the effect of coupling and differences in radiation pattern between elements is of interest. Moreover, one important parameter in a MIMO system is the achievable capacity, which can also be of mayor importance when evaluating the array performance. Therefore, throughout this chapter we have proposed several methods to account for these aspects.

We have presented results of four different methods of MIMO array evaluation: two based on measured antenna patterns and channel models (Kronecker model and 3GPP spatial channel model), one using a reverberation chamber, and a complete radio channel measurement. The investigated methods give very similar relative results for the two arrays tested. There is some difference in the absolute values since the real radio environment does not in general provide all possible degrees of freedom. The same applies for the 3GPP SCM simulations, which include some degree of correlation in order to account for outdoor scenarios. Conversely, the Kronecker model and the reverberation chamber overestimate the achievable capacity, since they assume ideal rich multipath channels, which is not the case in many real environments.

We have also compared the four proposed methods in terms of time and equipments requirements. While the radio channel measurement involves a very complex and time
consuming measurement campaign, the reverberation chamber is the fastest method provided that one of those chambers is available. If only radiation patterns are known, the methods based on simulation of radio channels are the most suitable ones, and can also provide good results.

We have also seen that the difference in performance between the arrays is rather small, in the order of 1 bits/s/Hz at 10 dB SNR, or 1.2 dB in power. This may be attributed to the measured difference in radiation efficiency, but we may also conclude that the differences in pattern shape have a minor effect in a general indoor scenario. Overall, we found that the use of the two antenna arrays in an indoor office environment achieve 80% and 89%, respectively, of the ideal i.i.d. capacity.

Contributions

The contributions of this chapter of the thesis to the current state of the art can be summarized as follows:

✓ A novel compact array has been designed and implemented, aiming at making MIMO systems available to hand-held devices. The designed antenna array presents a size suitable for conventional PDA-like user terminals, and consists of 4 PIFA with a very small footprint. The array has been designed in order to get a low coupling coefficient between elements, and good impedance matching was obtained. Thanks to the used microstrip technology, a low profile and low-cost implementation is obtained.

✓ A reference array for MIMO systems has been defined and implemented, in order to get reference values regarding array performance in a MIMO system. A linear array consisting of monopoles has been chosen, due to its good features regarding efficiency (close to 90%), good impedance matching and low coupling between element pairs. Also the omnidirectional radiation pattern of the elements was of interest, thus not giving more importance to any spatial direction.

✓ We have shown that conventional antenna parameters are not sufficient to fully evaluate an antenna array from a MIMO point of view, despite their interest. To do so, the novel compact array has firstly been evaluated by means of conventional antenna parameters, as radiation patterns, S parameters and efficiency. The designed novel array has shown good features regarding reflection coefficient, bandwidth and radiation patterns. The efficiency has been also computed, being c.a. 70%, lower than the efficiency of other antennas as dipoles or monopoles, mainly due to the used microstrip technology.

✓ In order to contribute to a better knowledge of “MIMO features” of antenna arrays, we have proposed 4 different methods to evaluate arrays from a MIMO point of view. Each method has been presented, its mathematical basis explained and finally each one has been used to compare the performance of the compact array and the reference array. As
a result, we have shown the advantages and disadvantages of each one, showing that there is not a definite one being the best method, but different methods for different cases or required results. Also, we have compared the time and equipment requirements for each one. As a conclusion, the method based on radio channel measurements has been highlighted as the most complex and time-consuming one, but also the one that offers more information, especially regarding site-specific data. Conversely, the method based on knowing the radiation pattern and using the Kronecker model and also the method based on measurements on a reverberation chamber are the less dependent on environment and less time-consuming, but they also give a too optimistic result regarding capacity when compared with more realistic results.

✔ By evaluating the compact array and the reference array, we have shown that the parameter that contributes the most in the achievable capacity for a general scenario is the radiation efficiency. The radiation patterns seem to have a minor effect in general, although it can be relevant in some specific situations such as those of low spatial diversity (e.g. indoor LOS cases in corridors). Finally, coupling coefficient were only considered in two of the analysed methods (radio channel measurements and reverberation chamber measurements), but they seem to have a negligible effect when they are kept low enough (below -10 dB).

Further research

The study and evaluation of antenna arrays for MIMO systems is currently a hot topic, and thus there are still many research aspects to work on. We have presented here a novel antenna array for realistic MIMO terminals, based on PIFA elements. However, other elements and technologies can also be of great interest, such as multimodal antennas or antennas arrays based on parasitic elements. Also reconfigurable antennas have experience an increasing interest within the MIMO community. Furthermore, antennas capable of working in two or even three frequency bands are currently being studied, in order to cover GSM, UMTS and WLAN systems simultaneously. Also dual- or even three-polarized antennas have been studied, in order to include polarization diversity in lower space. More research efforts are being carried out in this line to improve features of these antenna arrays and to adapt them to realistic terminals.

Also the evaluation of antennas for MIMO can be improved. We have presented and analysed four interesting methods, which mainly evaluates the achievable capacity including the antenna effect in the system. By using the same implemented antenna array, other authors have studied the influence of the body and the hand in the antenna performance from a MIMO point of view [281], which is also a very interesting and active topic to be studied and improved for achieving realistic antennas and terminals for MIMO. Also new parameters to evaluate the array performance from a MIMO-system point of view have recently been proposed, as in [167].
may be of interest to define novel terms or parameters to quantitative measure other MIMO characteristics, such as pattern or spatial diversity, the “spatial efficiency” or how good the space among antenna elements is used, etc. Finally, it may be interesting not only to consider no CSI at transmitter, but also to study how to optimally design the antenna array from the point of view of full CSI or partial CSI at transmitter.
GENERAL CONCLUSIONS, FUTURE WORK AND PUBLICATIONS

7.1 General conclusions and contributions .................................................... 306

7.2 Future work ................................................................................................ 309

7.3 Publications ................................................................................................. 310
The work carried out along this thesis covers a multidisciplinary theme, from theoretical analysis and comparison of algorithms to prototype implementation and evaluation. Each one of the chapters has been devoted to one specific topic, but all of them are guided by the same objectives: showing the possibilities of multi-antenna systems in mobile and wireless communications, casting light on novel aspects of these systems in order to boost their adoption and proposing new solutions to overcome some implementation issues. Although specific conclusions have been presented at the end of each chapter, we summarize here the main conclusions of the overall work, and also the research lines that are proposed for the future.

The research work that has been done throughout this thesis has been accepted for publication in several technical journals. Moreover, the work has been presented at the main international conferences in the field of antennas and signal processing for communications. These contributions are mentioned here.
7.1 General conclusions and contributions

The work in this thesis has been divided into four main blocks or areas, corresponding to the results presented in chapters 3 to 6, and according to the main research topics and open issues in multi-antenna systems that were identified in chapter 2. For each area, several conclusions have been presented at the end of the chapter, but we summarize here the main idea drawn from the realized work. Figure 7-1 depicts the four main blocks of work and the overall contributions of the thesis to the state of the art, skipping the details (which can be consulted in the corresponding chapters).

![Figure 7-1. Topics investigated in this thesis and synopsis of main contributions.](image)
Most of the results presented in this thesis have been obtained from measurements and evaluation of real systems or prototypes. Also new measurement campaigns have been carried out in order to analyse propagation characteristics. The use of real-world data obtained from measurements instead of just simulations means a further step in the obtained results, since aspects as non-ideal effects are also included.

As a general conclusion of the overall thesis, multi-antenna systems have been shown to offer significant improvements in wireless systems, especially regarding interference reduction and capacity or throughput increase. Furthermore, throughout the thesis it has been shown that some real-implementation problems may arise when trying to bring these systems into real prototypes and realistic user equipments, but we have presented analysis of some of these problems and also some possible solutions to mitigate the undesired effects have been presented, showing that even with non-ideal effect adaptive antennas and MIMO system bring very interesting advantages.

In this thesis we have demonstrated that a real-time prototype of adaptive antenna is feasible. It has been designed and implemented, and along this document we have mentioned some impairments due to non-ideal effects in the real prototype. Despite these effects, we have shown that the implemented four-element adaptive antenna can offer an increase in gain of up to 6 dB for optimal operation, and a reduction in interferences of approximately 15 dB was achieved. All these results were obtained from measurements in controlled and also realistic scenarios. Moreover, we have thoroughly studied the possibilities to seamlessly integrate this adaptive antenna in a real UMTS network and have proposed and developed a set-up module, testing its operation in the first version of a real 3G network.

Adaptive antennas will provide nice results in environments where multipath richness is small, such as rural areas, or where interferences are strong and spatially coloured, as in hot spots. However, when wireless environments are rich in multipath and scatterers (as in indoor or urban areas), MIMO processing is preferred to take advantage of spatial diversity. In order to evaluate several aspects of MIMO systems we have developed a new MIMO testbed, based on rapid prototyping approach, and we have presented its possibilities to perform MIMO channel measurements, to evaluate different array configurations and to test multi-antenna algorithms. Several conclusions have been drawn from a measurement campaign carried out with this MIMO testbed, such as a comparison of performance in terms of capacity and path loss for single-polarized and dual-polarized MIMO systems in indoor environments.

Other propagation results have been presented from another measurement campaign, focusing on the comparison of indoor and outdoor-to-indoor scenarios. While the latter ones have shown to offer a higher diversity and they may simplify the deployment of base stations to cover buildings and other indoor scenarios, they require higher transmitted power (up to 30 dB in our measured scenario) even if directive (in elevation) antennas are used at base station.
Other interesting conclusions have been presented regarding base station location and signal combination for both no CSI at transmitter and perfect CSI at transmitter, which may help as hints for indoor network deployment for MIMO systems.

Finally, in order to cover the three main parts in the communication multi-antenna system study (base station, channel and user equipment), we have also investigated the terminal for the user. With the current trend towards increasing the size of mobile phones (with, for example, bigger screens) and adding new features that require higher data rate (WiFi connection, high quality video-conferences…), designing and evaluating new compact antenna arrays for handheld devices is needed. In this thesis we have presented a novel design for a multi-element array with four antennas that can be implemented on a realistic PDA, obtaining good results in terms of conventional antenna parameters. Moreover, we have shown that conventional antenna parameters as radiation patterns can be used to compute other parameters of great interest to evaluate the antenna performance in MIMO systems, as the available capacity under certain assumptions. Other methods that are not directly based on just conventional parameters have also been presented and compared. We have concluded that, contrary to what it is sometimes assumed, indoor scenarios cannot be considered as rich in multipath as ideal i.i.d channels, and therefore capacity measured in reverberant chambers in order to evaluate antennas for MIMO will be an optimistic approach in general. On the other hand, the standardized MIMO channel from 3GPP only considers a 2-D channel, thus neglecting contributions from diversity in 3D radiation patterns and channel. Therefore, results from the channel model including 2D radiation pattern will be a pessimistic approach in general. Other interesting conclusions are obtained from comparing the studied methods to evaluate antennas for MIMO.
7.2 Future work

Multi-antenna system for wireless communications is a very wide topic with many research areas to be covered and many open issues. Despite some of them have been tackled in this thesis, there is still a lot of work to be done. We mention here some of the lines to be investigated in the future that are more related to the work presented here.

Regarding adaptive antennas based on beamforming methods, we have presented here a prototype and we have evaluated its performance in some scenarios. However, validations tests in multiple situations and with many users would be desirables to evaluate the performance of adaptive antennas from a system-level point of view. Moreover, our solution does not consider ciphering of higher levels, which in current UMTS networks is the usual operational mode. Thus, commercial adaptive antennas should address this issue, by defining a higher level protocol to communicate the required cipher information to the adaptive antenna. Besides this, some previous works have already evaluated adaptive antennas at system-level in cellular communications, but measurements in real networks with multiple adaptive antennas are scarce but still very interesting.

More channel measurements and detailed analysis are needed in order to better understand the MIMO channel and model its features. Some examples of open issues in this area are modelling of scatterers clustering, investigation of three-polarized channels or efficient methods to model wideband channels, but much more can be found in the literature (and also in chapter 2 of this thesis).

We have analysed available capacity for two opposite assumptions: no CSI at transmitter and perfect CSI at transmitter. However, in a real application it is likely that only partial CSI will be available at the transmitter. Although some previous work has been done in this topic, new algorithms for partial CSI are very interesting for improving system performance in real world. We have not addressed this issue along the thesis, but it may be a future research line.

Novel antennas specially designed for MIMO systems and the evaluation of antennas in MIMO systems have been addressed in this thesis. However, due to the wide area of this topic, there is still a lot of work to be done regarding antenna aspects. For example, other type of antennas for handheld devices and other type of user terminals (such as laptops, smart phones, etc…) are desirable. For example, antenna arrays with parasitic elements or multimodal antennas are good candidates for user equipments. Working together with equipment manufacturers will be required in order to optimize the antenna array including the effect of the terminal and in order to see multi-antenna terminals in real life.


7.3 Publications

The work presented in this thesis has given rise to several publications in international technical journals. The work has also been presented at some of the most important international and national conferences in communications and antenna areas. They are listed below, classified in four groups corresponding to the four blocks (chapters) which we have divided the work into.

Adaptive antennas for mobile systems (chapter 3)

Articles published in technical journals


Articles presented in international conferences:


Articles presented in national conferences:


Rapid prototyping for MIMO (chapter 4)

Articles presented in international conferences:


**Articles presented in national conferences:**


General conclusions, future work and publications


Propagation aspects of MIMO systems (chapter 5)

Articles published in technical journals:


Articles presented in international conferences:


Articles presented in national conferences:

Antenna aspects of MIMO systems (chapter 6)

Articles published in technical journals:


Articles presented in international conferences:


Finally, the work presented in this PhD thesis has given rise to one patent, two contributions to technical books and two invited talks at conferences. As part of this PhD thesis, two Master Thesis were directed. They are detailed below:

* Patent:*
· Book contributions:


· Invited talks:


· Supervised Master Theses:


APPENDIX I: Antennas for measurements

The main features of the antennas that were used to carry out the measurement campaign of Chapter 5 are included here.

TX ANTENNA FOR INDOOR SCENARIOS

Two dual-polarized planar antennas were used as TX antennas for indoor BS locations. The model Huber-Suhner SPA 1800/85/8/0/DS (see Figure A-1) was selected, as a good representative of typical antennas used for indoor applications. It is a quite cheap antenna with a wide beam in the horizontal plane, thus suitable for applications that do not require high gains or directive beams. Figure A-2 shows the radiation pattern of this antenna, for both vertical and horizontal planes. The main features of the antenna are summarized in Table A-I. For more details on the antenna, the manufacturer data sheet [282] can be consulted.

Figure A-1. Planar element used at TX for indoor measurements

Figure A-2. Radiation pattern of antenna used at TX for indoor locations, for both horizontal (left) and vertical (right) planes.
**Table A-I. Main characteristics of planar antenna used for indoor TX scenarios.**

<table>
<thead>
<tr>
<th><strong>Huber-Suhner SPA 1800 planar antenna</strong></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
<td>1770-1880 MHz</td>
</tr>
<tr>
<td>Polarization</td>
<td>dual linear, ± 45° slant</td>
</tr>
<tr>
<td>Gain</td>
<td>8.0 dBi</td>
</tr>
<tr>
<td>3 dB beamwidth horizontal</td>
<td>85°</td>
</tr>
<tr>
<td>3 dB beamwidth vertical</td>
<td>75°</td>
</tr>
<tr>
<td>Max. power</td>
<td>75 W (CW) at 25°C</td>
</tr>
<tr>
<td>Connectors</td>
<td>2, SMA</td>
</tr>
</tbody>
</table>

**TX ANTENNA FOR OUTDOOR-TO-INDOOR SCENARIOS**

Two dual-polarized planar antenna arrays were used as TX antennas for outdoor BS locations. The model UXM 1710-2170-65-15i-0-D (7700.00) from Powerwave was selected (see Figure A-3 for a photograph of one antenna). It is a broadband antenna with a wide beam in the horizontal plane, but a directive beam with a mechanically selectable tilt in the vertical plane. Thus, its main applications are radio links and base stations for mobile communications. Figure A-4 shows the radiation pattern of this antenna, for both vertical and horizontal plane. The main features of the antenna are summarized in Figure A-4. For more details on the antenna, the manufacturer data sheet [283] can be consulted.

![Figure A-3. Antenna used for outdoor TX locations.](image)
Appendix I

Figure A-4. Radiation pattern of antenna used at TX for outdoor locations, for both horizontal (left) and vertical (right) planes.

<table>
<thead>
<tr>
<th><strong>Powerwave 7700.00 antenna array</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
</tr>
<tr>
<td>Polarization</td>
</tr>
<tr>
<td>Gain</td>
</tr>
<tr>
<td>3 dB beamwidth horizontal</td>
</tr>
<tr>
<td>3 dB beamwidth vertical</td>
</tr>
<tr>
<td>Max. power</td>
</tr>
<tr>
<td>Connectors</td>
</tr>
</tbody>
</table>

Table A-II. Main characteristics of planar antenna used for outdoor TX scenarios.
BIBLIOGRAPHY


A. Wolfgang, N. N. Ahmad, S. Chen, L. Hanzo, “Genetic algorithm assisted error probability optimisation for beamforming”.


[136] [http://www.ftw.at/cost273](http://www.ftw.at/cost273)


[141] [https://www.ist-winner.org/](https://www.ist-winner.org/)


Bibliography


[210] Ramón Martínez Rodríguez-Osorio, Estudio sobre las Prestaciones de Antenas Inteligentes en Sistemas de Comunicaciones Móviles de Tercera Generación (UMTS), PhD Thesis (only in Spanish), Universidad Politecnica de Madrid, 2004


[233] ITU-T X.691 Abstract Syntax Notation One (ASN.1).


