

OFDM Wireless Multimedia Communications

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Richard van Nee
Ramjee Prasad



Artech House
Boston • London

TK5103
.2
.N44
2000
2000.1

Library of Congress Cataloging-in-Publication Data

Nee, Richard van.

OFDM for wireless multimedia communications / Richard van Nee, Ramjee Prasad
p. cm. — (Artech House universal personal communications library)

Includes bibliographical references and index.

ISBN 0-89006-530-6 (alk. paper)

1. Wireless communications systems. 2. Multimedia systems. 3. Multiplexing. I.
Prasad, Ramjee. II. Title III. Series

TK5103.2.N44 2000

621/3845—dc21

99-052312

CIP

British Library Cataloguing in Publication Data

Nee, Richard van

OFDM wireless multimedia communications. — (Artech House
universal personal communications library)

1. Wireless communications systems 2. Multimedia systems

I. Title II. Prasad, Ramjee

621.3'82

ISBN 0-89006-530-6

Cover design by Igor Vladman

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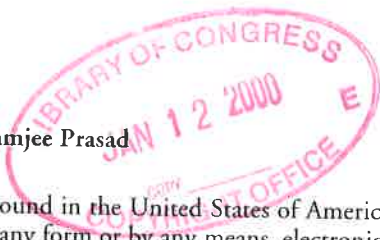
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International Standard Book Number: 0-89006-530-6

Library of Congress Catalog Card Number: 99-052312

10 9 8 7 6 5 4 3 2 1



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Preface

सर्वद्वारेषु देहेऽस्मिन् प्रकाश उपजायते ।
ज्ञानं यदा तदा विद्याद् विवृद्धं सत्त्वमित्युत

*sarva-dvāreṣu dehe 'smin
prakāśa upajāyate
jñānaṁ yadā tadā vidyād
vivṛddhaṁ sattvaṁ ity uta*

**The manifestations of the mode of goodness can be experienced when all
the gates of the body are illuminated by knowledge**

The Bhagavad Gita (14.11)

During the joint supervision of a Master's thesis "The Peak-to-Average Power Ratio of OFDM," of Arnout de Wild from Delft University of Technology, The Netherlands, we realized that there was a shortage of technical information on orthogonal frequency division multiplexing (OFDM) in a single reference. Therefore, we decided to write a comprehensive introduction to OFDM. This is the first book to give a broad treatment to OFDM for mobile multimedia communications. Until now, no such book was available in the market. We have attempted to fill this gap in the literature.

Currently, OFDM is of great interest by the researchers in the Universities and research laboratories all over the world. OFDM has already been accepted for the new wireless local area network standards from IEEE 802.11, High Performance Local Area Network type 2 (HIPERLAN/2) and Mobile Multimedia Access Communication (MMAC) Systems. Also, it is expected to be used for the wireless broadband multimedia communications.

OFDM for Wireless Multimedia Communications is the first book to take a comprehensive look at OFDM, providing the design guidelines one needs to maximize benefits from this important new technology. The book gives engineers a solid base for assessing the performance of wireless OFDM systems. It describes the new OFDM-based wireless LAN standards; examines the basics of direct-sequence and frequency-hopping CDMA, helpful in understanding combinations of OFDM and CDMA. It also looks at applications of OFDM, including digital audio and video broadcasting, and wireless ATM. Loaded with essential figures and equations, it is a must-have for practicing communications engineers, researchers, academics, and students of communications technology.

Chapter 1 presents a general introduction to wireless broadband multimedia communication systems (WBMCS), multipath propagation, and the history of OFDM. A part of this chapter is based on the contributions of Luis Correia from the Technical University of Lisbon, Portugal, Anand Raghawa Prasad from Lucent Technologies, and Hiroshi Harada from the Communications Research Laboratory, Ministry of Posts and Telecommunications, Yokosuka, Japan.

Chapters 2 to 5 deal with the basic knowledge of OFDM including modulation and coding, synchronization, and channel estimation, that every post-graduate student as well as practicing engineers must learn. Chapter 2 contains contributions of Rob Kopmeiners from Lucent Technologies on the FFT design.

Chapter 6 describes the peak-to-average power problem, as well as several solutions to it. It is partly based on the contribution of Arnout de Wild.

Basic principles of CDMA are discussed in Chapter 7 to understand multi carrier CDMA and frequency-hopping OFDMA, which are described in Chapters 8 and 9. Chapter 8 is based on the research contributions from Shinsuke Hara from the University of Osaka, Japan, a postdoctoral student at Delft University of Technology during 1995–96. Chapter 9 is based on a UMTS proposal, with main contributions of Ralf Böhnke from Sony, Germany, David Bhatoolaul and Magnus Sandell from Lucent Technologies, Matthias Wahlquist from Telia Research, Sweden, and Jan-Jaap van de Beek from Lulea University, Sweden.

Chapter 10 was written from the viewpoint of top technocrats from industries, government departments, and policy-making bodies. It describes several applications of OFDM, with the main focus on wireless ATM in the Magic WAND project, and the new wireless LAN standards for the 5 GHz band from IEEE 802.11, HIPERLAN/2 and MMAC. It is partly based on contributions from Geert Awater from Lucent Technologies, and Masahiro Morikura and Hitoshi Takanashi from NTT in Japan and California, respectively.

We have tried our best to make each chapter quite complete in itself. This book will help generate many new research problems and solutions for future mobile multimedia communications. We cannot claim that this book is errorless. Any remarks to improve the text and correct any errors would be highly appreciated.

Acknowledgments

The material in this book originates from several projects at Lucent Technologies and research activities at Delft University of Technology, The Netherlands. A great deal of OFDM-knowledge was acquired during the Magic WAND project, in which several companies jointly managed to build a wireless ATM network demonstrator based on OFDM. We wish to thank all Magic WAND members who were involved in the design of the OFDM modem, in particular Geert Awater, James Hopper, Rob Kopmeiners, Erik Busking, Han Schmitz, Theo Kleijne, Martin Janssen, Jan Kruys, Urs Bernhard, Urs Lott, Alex Grant, James Aldis, Thomas Mark, Rodolfo Mannpelz, and Ari Vaisanen. Another fruitful source of information was the OFDM-based proposal for UMTS, with main contributions coming from Ralf Böhnke, David Bhatoolaul, Magnus Sandell, and Jan-Jaap van de Beek.

Richard wishes to thank Masahiro Morikura and Hitoshi Takanashi for all their contributions and the pleasant cooperation on the joint OFDM proposal for the IEEE 802.11 wireless LAN standard. All members of IEEE 802.11 and HiperLAN/2 are thanked for numerous contributions, that greatly improved the quality of the final standards and helped to gain more insight in various OFDM techniques.

Geert Awater and Rob Kopmeiners made contributions and corrected numerous mistakes. Neeli Rashmi Prasad helped to prepare the complete manuscript, freeing us from the enormous burden of editorial requirements. Shinsuke Hara from the University of Osaka, Japan, Hiroshi Harada from the Communications Research Laboratory, Ministry of Posts and Telecommunications, Yokosuka, Japan, Luis Correia from the Technical University of Lisbon, Portugal, Anand Raghava Prasad from Lucent Technologies, and Arnout de Wild from Siemens, The Netherlands, are deeply acknowledged for their valuable contributions.

After the Magic WAND project, we studied several OFDM options for new high-rate wireless LAN products. A circle was completed by the selection of OFDM for the high rate extension of the IEEE 802.11 wireless LAN standard in July 1998. We hope this book will help to gain insight in the principles and design of OFDM-based systems, in particular the new OFDM-based wireless LAN standards.

*Richard van Nee
Ramjee Prasad
October 1999*

Chapter 1

Introduction

The spectacular growth of video, voice, and data communication over the Internet, and the equally rapid pervasion of mobile telephony, justify great expectations for mobile multimedia. Research and development are taking place all over the world to define the next generation of wireless broadband multimedia communications systems (WBMCS) that may create the “global information village.” Figure 1.1 illustrates the basic concept of the global information village, which consists of various components at different scales ranging from global to picocellular size. As we know, the demand for wireless (mobile) communications and Internet/multimedia communications is growing exponentially. Therefore, it is imperative that both wireless and Internet/multimedia should be brought together. Thus, in the near future, wireless Internet Protocol (IP) and wireless asynchronous transfer mode (ATM) will play an important role in the development of WBMCS.

While present communications systems are primarily designed for one specific application, such as speech on a mobile telephone or high-rate data in a wireless local area network (LAN), the next generation of WBMCS will integrate various functions and applications. WBMCS is expected to provide its users with customer premises services that have information rates exceeding 2 Mbps. Supporting such large data rates with sufficient robustness to radio channel impairments, requires careful choosing of modulation technique. The most suitable modulation choice seems to be orthogonal frequency division multiplexing (OFDM). Before going into the details of OFDM, however, first we give some background information on the systems that will be using it.

The theme of WBMCS is to provide its users a means of radio access to broadband services supported on customer premises networks or offered directly by public fixed networks. WBMCS will provide a mobile/movable wireless extension to

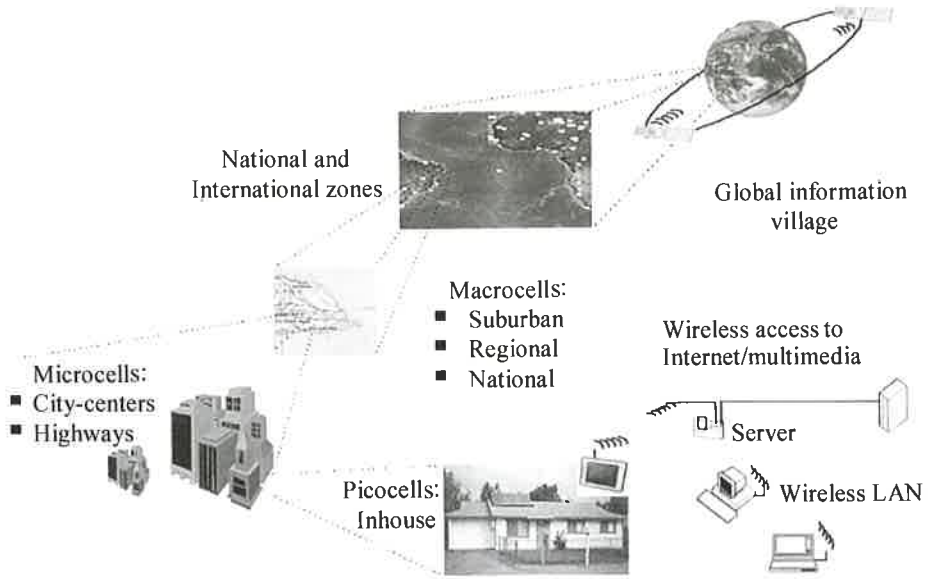


Figure 1.1 Global information village.

WBMCS is under investigation in North America, Europe, and Japan in the microwave and millimeter-wave bands to accommodate the necessary bandwidth. The research in the field of WBMCS has drawn much attention because of the increasing role of multimedia and computer applications in communications. There is a major thrust in three research areas: (1) microwave and millimeter-wave bands for fixed access in outdoor, public commercial networks, (2) evolution of WLAN for inbuilding systems, and (3) use of LAN technology outdoors rather than indoors. In short, WBMCS will provide novel multimedia and video mobile communications services, also related to wireless customer premises network (WCPN) and wireless local loop (WLL).

To implement the wireless broadband communication systems, the following challenges must be considered:

- Frequency allocation and selection;
- Channel characterization;
- Application and environment recognition, including health hazard issues;
- Technology development;
- Air interface multiple access techniques;
- Protocols and networks; and
- Systems development with efficient modulation, coding, and smart antenna techniques.

- Protocols and networks; and
- Systems development with efficient modulation, coding, and smart antenna techniques.

A significant number of research and development (R&D) projects are set up in the area of WBMCS. Within the European Advanced Communication Technologies and Services (ACTS) program are four European Union-funded R&D projects, namely Magic Wand (Wireless ATM Network Demonstrator), ATM Wireless Access Communication System (AWACS), System for Advanced Mobile Broadband Applications (SAMBA), and wireless broadband CPN/LAN for professional and residential multimedia applications (MEDIAN) [1–17]. Table 1.1 summarizes the European projects [11].

In the United States, seamless wireless network (SWAN) and broadband adaptive homing ATM architecture (BAHAMA), as well as two major projects in Bell Laboratories and a wireless ATM network (WATMnet), are being developed in the computer and communications (C&C) research laboratories of Nippon Electric Company (NEC) in the United States [2–6].

In Japan, Communication Research Laboratory (CRL) is working on several R&D projects, such as a broadband mobile communication system in the super high frequency (SHF) band (from 3 to 10 GHz) with a channel bit rate up to 10 Mbps and an indoor high speed WLAN in SHF band with a target bit rate of up to 155 Mbps [12].

In the Netherlands, Delft University of Technology has been busy with a multi-disciplinary research project, “Mobile Multimedia Communication (MMC),” since April 1996. The team consists of experts from the telecommunications and traffic control and information theory groups of the department of Electrical Engineering, the Product Ergonomics group of the department of Industrial Design Engineering, and the Organizational Psychology group of the department of Technology and Society.

The MMC has the following objectives to achieve at 60 GHz:

- Wireless access of 155 Mbps using OFDM;
- Both indoor and outdoor use;
- Less complex, inexpensive mobile stations by moving most functionality to the access points;
- Modified OFDM; and
- Constant bit rate (CBR), variable bit rate (VBR), and available bit rate (ABR) services.

Table 1.1
Summary of European ACTS Projects.

ACTS project	WAND	AWACS	SAMBA	MEDIAN
Parameter				
Frequency	5 GHz	19 GHz	40 GHz	61.2 GHz
Data rate	20 Mbps	70 Mbps	34 Mbps	155 Mbps
Modulation	OFDM, 16 subcarriers with 8-PSK (phase shift keying)	offset quadrature PSK (OQPSK)	OQPSK	OFDM, 512 carriers, differential QPSK (DQPSK)
Cell radius	20–50m (omnidirectional antennas)	50–100m (directional antennas, line-of-sight only)	10-50m (directional antennas at access point)	10m (directional antennas)
Radio access	time division multiple access/ time division duplex (TDMA/TDD)	TDMA/TDD	time division multiple access /frequency division duplex (TDMA/FDD)	TDMA/TDD

1.1 STANDARDIZATION AND FREQUENCY BANDS

There are three main forums for the standardization of wireless broadband communication systems; namely, IEEE 802.11 [18], European Telecommunication Standards Institute Broadband Radio Access Networks (ETSI BRAN) [19], and Multimedia Mobile Access Communications (MMAC) [20]. IEEE 802.11 made the first WLAN standard for the 2.4-GHz Industrial, Scientific, and Medical band (ISM). It specifies the medium access control and three different physical layers—direct-sequence spread spectrum, frequency hopping, and infrared—which give a data rate of 2 Mbps. Products based on this standard became available in 1998. Figure 1.2 shows an example of an IEEE 802.11 modem in a PCMCIA card. Following the initial 1- and 2-Mbps standard, IEEE 802.11 developed two new physical layer standards. One delivers data rates of up to 11 Mbps in the 2.4-GHz band, using complementary code keying [21,22]. Products based on this standard—with the old 1 and 2 Mbps as fallback rates—are available since mid 1999. An industry alliance called the Wireless Ethernet Compatibility Alliance (WECA) has been established to promote the high rate IEEE

available since mid 1999. An industry alliance called the Wireless Ethernet Compatibility Alliance (WECA) has been established to promote the high rate IEEE 802.11 technology and to certify interoperability of products from different vendors [23]. The second IEEE 802.11 standard extension targets a range of data rates from 6 up to 54 Mbps using OFDM in the 5-GHz band [24]. The OFDM standard was developed jointly with ETSI BRAN and MMAC, making OFDM effectively a worldwide standard for the 5-GHz band.

Table 1.2 lists the main characteristics of the IEEE 802.11 and the ETSI High Performance Local Area Network type 2 (HIPERLAN/2) standards. More details about these standards can be found in Chapter 10.



Figure 1.2 IEEE 802.11 modem for the 2.4-GHz band (WaveLAN™ from Lucent Technologies [25]).

Figure 1.3 shows that 2-, 5- and 60-GHz are the commercially important frequency bands because of geographically wide spectrum allocations in Europe, the United States (U.S.), and Japan for the wireless broadband multimedia communications networks. The 2.4-GHz band is an ISM band, which can be used for many types of transmission systems as long as they obey certain power, spectral density, and spreading gain requirements. The 5-GHz band is designated specifically for WBMCS. In Europe, only HIPERLAN devices are currently allowed in this band. HIPERLAN actually consists of a family of standards, one of which is an OFDM-based standard that is very similar to the IEEE 802.11 5-GHz standard. In Japan, MMAC supports both the IEEE 802.11 and the HIPERLAN standards. Notice that Japan only has 100 MHz available in the 5-GHz band, while the United States and Europe provide 300 and 455 MHz, respectively. In Europe, extra spectrum for HIPERLAN is available in the 17-GHz band, while Japan has allocated spectrum from 10- to 16-GHz to mobile broadband systems (MBS). An analysis of the propagation aspects at the bands foreseen for WBMCS microwaves, millimeterwaves, and infrared is presented in [26–31].

Table 1.2
Comparison of IEEE and HIPERLAN standards.

Parameter	IEEE 802.11 2 GHz	IEEE 802.11 5 GHz	HIPERLAN/2
Configurations	Centralized system with access points connected to wired network, or peer-to-peer networking	Centralized system with access points connected to wired network, or peer-to-peer networking	Centralized system with access points connected to wired network
Range	Up to 60m at 11 Mbps and up to 100m at 2 Mbps with omnidirectional antennas	Up to 30m at 24 Mbps and up to 60m at 6 Mbps with omnidirectional antennas	Up to 30m at 24 Mbps and up to 60m at 6 Mbps with omnidirectional antennas
Channel access	CSMA/CA, variable size data packets (up to 8192 bytes)	CSMA/CA, variable size data packets (up to 8192 bytes)	Reservation based access, scheduled by access point. Contention slots for making slot reservations
Frequency bands	2.4-2.4835 GHz	5.150-5.350 GHz; 5.725-5.825 GHz	5.150-5.350 GHz; 5.470-5.725 GHz;
Duplexing	TDD	TDD	TDD
Data rate	1, 2 Mbps (BPSK/QPSK) 5.5, 11 Mbps (CCK)	6, 9 Mbps (BPSK) 12, 18 Mbps (QPSK) 24, 36 Mbps (16-QAM) 54 Mbps (64-QAM)	6, 9 Mbps (BPSK) 12, 18 Mbps (QPSK) 24, 36 Mbps (16-QAM) 54 Mbps (64-QAM)

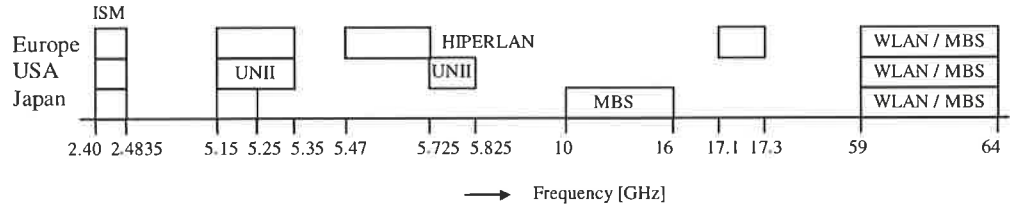


Figure 1.3 Frequency band for wireless broadband communications.

1.2 MULTIMEDIA COMMUNICATIONS

Multimedia and computer communications are playing an increasing role in today's society, creating new challenges to those working in the development of telecommunications systems. Besides that, telecommunications is increasingly relying upon wireless links. Thus, the pressure for wireless systems to cope with increasing data rates is enormous, and WBMCSs with data rates higher than 2 Mbps are emerging rapidly, even if at this moment applications for very high transmission rates do not exist.

Several WBMCSs are being considered for different users with different needs. They may accommodate data rates ranging between 2 and 155 Mbps; terminals can be mobile (moving while communicating) or portable (static while communicating); moving speeds can be as high as that of a fast train; users may or may not be allowed to use more than one channel if their application requires so; the system bandwidth may be fixed, or dynamically allocated according to the user's needs; communication between terminals may be direct or must go through a base station; possible ATM technology use; and so on. Many other cases can be listed as making the difference between various perspectives of a WBMCS, but two major approaches are emerging: WLANs directed to communication among computers, from which IEEE 802.11 [16, 18] and HIPERLAN [19, 20] are examples, with MBS [17] intended as a cellular system providing full mobility to B-ISDN users.

The different requirements imposed by the various approaches to WBMCSs have consequences on system design and development. The tradeoffs between maximum flexibility on one hand and complexity and cost on the other are always difficult to decide, as they have an impact not only on the deployment of a system, but also on its future evolution and market acceptance. GSM is a good example of a system foreseen to accommodate additional services and capacities to those initially offered, and the fact that operators are already implementing phase 2+ is proof of that.

This means that many decisions must be made on the several WBMCSs that will appear on the market. For IEEE 802.11, for example, those decisions have already been made, as the system will be commercialized in the very near future, but for other systems there are still many undecided aspects. Of course this depends on what are the applications intended to be supported by the systems, and whether these applications are targeted to the mass market or only to some niches. The former (from which mobile telephones are a good example) will certainly include WLANs, because the expansion of personal computers will dictate this application as a great success in WBMCSs; the latter will possibly have television broadcasters among their users (to establish links between HDTV cameras and the central control room).

Not only are market aspects at stake in the development and deployment of WBMCSs, but many technical challenges are posed as well. The transmission of such high data rates over radio in a mobile environment creates additional difficulties,

compared with that of existing systems; these difficulties are augmented by the fact that frequencies higher than UHF are needed to support the corresponding bandwidths, thus pushing mobile technology challenges (size and weight among other things) to frequencies where these aspects were not much considered until now. However, additional challenges are posed to those involved in WBMCSs development: in today's world, where consumers are in the habit of using a communications system that is available in different places (e.g., GSM roaming capability, because users can make and receive telephone calls in an increasing number of countries worldwide), or being able to exchange information among different systems (e.g., the exchange of files between different computer applications and systems), for future use it does not make sense to consider systems that offer a high data rate but do not support these capabilities to some extent.

1.2.1 Need for High Data Rates

Data rate is really what broadband is about. For example, the new IEEE and HIPERLAN standards specify bit rates of up to 54 Mbps, although 24 Mbps will be the typical rate used in most applications. Such high data rates impose large bandwidths, thus pushing carrier frequencies for values higher than the UHF band: HIPERLAN has frequencies allocated in the 5- and 17-GHz bands; MBS will occupy the 40- and 60-GHz bands; and even the infrared band is being considered for broadband WLANs. Many people argue whether there is a need for such high-capacity systems, however, bearing in mind all the compression algorithms developed and the type of applications that do require tens of megabits per second. We can examine this issue from another perspective.

The need for high-capacity systems is recognized by the "Visionary Group" [32], put together by the European Commission, to give a perspective of what should be the hot topics in the telecommunications for research in the next European programs (following R&D in Advanced Communications Technologies for Europe (RACE) and ACTS). In this visionary perspective, to meet the needs of society in the years to come as far as communications is concerned, capacity is one of the major issues to be developed because of the foreseen increase in demand for new services (especially those based on multimedia). Along with this, mobility will impose new challenges to the development of new personal and mobile communications systems.

We can conclude the following: even if at a certain point it may look academic to develop a system for a capacity much higher than what seems reasonable (in the sense that there are no applications requiring such high capacity), it is worthwhile to do so, as almost certainly in the future (which may be not very far off) applications will need those capacities and even more. The story of fiber optics is an example.

1.2.2 Services and Applications

The system concept of a WLAN such as IEEE 802.11 and of a mobile broadband cellular system such as MBS is totally different: each is directed to services and applications that differ in many aspects. A comparison of several systems, based on two of the key features (mobility and data rate), is shown in Figure 1.4 [33], where it is clear that no competition exists between the different approaches.

The applications and services of the various systems are also different. IEEE 802.11 is mainly intended for communications between computers (thus being an extension of wired LANS); nevertheless, it can support real-time voice and image signals, and users are allowed some mobility and can have access to public networks.

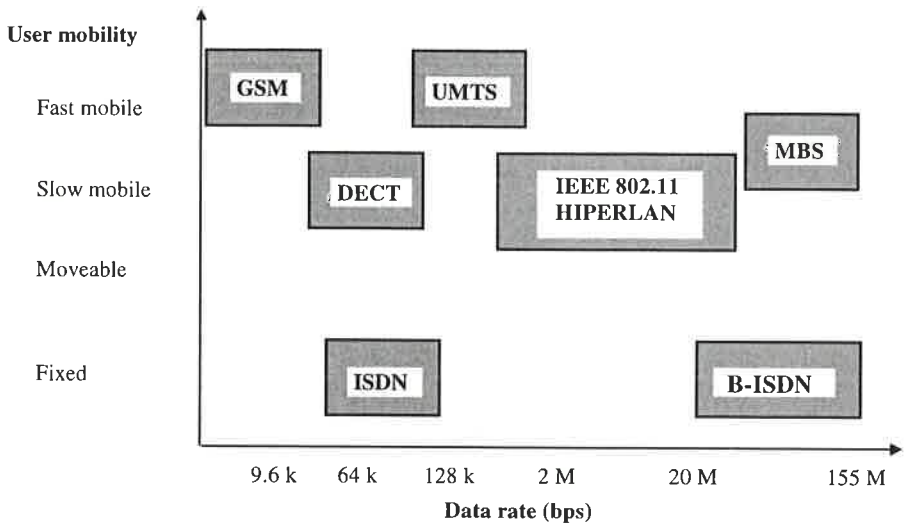


Figure 1.4 Comparison of mobility and data rates for several systems.

1.2.3 Antennas and Batteries

Antennas and batteries play a key role in wireless systems. With the advent of microelectronics and signal processing, antennas and batteries tend to impose the size and weight of mobile terminals. Of course, the higher one goes in frequency, the less developed the technology, and many problems are still found in size and weight at the millimeter-wave band. Power consumption is one example. Though these are likely to be solved in the near future. The number of hours battery-powered equipment can work or operate on stand by, and the percentage of its weight corresponding to the battery, is

not a specific problem of WBMCS. Laptop computers and cellular telephones are the most common terminals relying on batteries these days. This technology continues to demand huge R&D, to extend working time and reduce weight. Although a 100g mobile telephone battery corresponding to several hours of continuous work and a few days on standby is already on the market, users still want more. Mobile multimedia terminals, certainly those to be used in some applications of WBMCSs, will be an extension of the current cellular telephones. Therefore, we foresee that the current problems associated with batteries will be transposed on WBMCS terminals; the same applies to laptops.

Antennas (size, type, technology, etc.) are not a specific problem of WBMCSs as well, but again they are very much related to the type of systems that will be made available for users. It does not make sense to impose restrictions (e.g., pointing in a certain direction or avoiding someone to pass in between) on the type of mobility a mobile terminal can have. Even for portable terminals (e.g., computers), these restrictions make no sense. Hence, there are only two options as far as antennas are concerned: either an omnidirectional antenna (dipole type) or an adaptive array antenna is used. Either way, patch antennas seem a very promising solution for general use in WBMCS. For the frequency bands we consider, isolated patches or adaptive arrays (with many elements) can be made with a small size (e.g., a credit card), thus enabling the terminal not to be limited in size by the antenna system.

The role of antenna radiation patterns is not negligible when discussing system performance, nor is their influence on parameters associated with wave propagation. Although the tradeoff between an omnidirectional and a narrowbeam antenna is not particular of WBMCS, it assumes particular importance at microwave and millimeter waves because of the characteristics of wave propagation at these bands. Using an omnidirectional antenna means a lower gain, but also the possibility of receiving signals from various directions, without the requirement for knowing where the base station is, and allowing the received rays coming from reflections on the propagation scenario. On the other hand, the use of a very directive antenna provides a higher gain, but it must be pointed at the base station and does not receive reflected waves coming from directions very different from the one to which it is pointed. The need for pointing it can be very discouraging, if not a drawback, when a line of sight does not exist. On the other hand, omnidirectional antennas lead to high values of delay spread, but they may ensure that the link still exists, relying on reflections if the line of sight is lost.

1.2.4 Safety Considerations

Until a few years ago, the analysis of possible harmful effects of electromagnetic radiation on people was devoted mainly to power lines and radars, because of the huge power levels involved in those systems. Even when mobile telephone systems appeared, there was no major concern, as the antennas were installed on the roofs of cars. With

the development of personal communication systems, in which users carry mobile telephones inside their coat pockets, and the antenna radiates a few centimeters from the head, safety issues gained great importance and a new perspective. Much research in the literature focuses not only on the absorption of power inside the head, but also on the influence of the head on the antenna's radiation pattern and input impedance. However, these works have addressed only the frequency bands used in today's systems; that is, up to 2-GHz (mainly on the 900- and 1,800-MHz bands), and only very few references are made to systems working at higher frequencies, as it is in the case of WBMCSs.

The problems associated with infrared are different from those posed by microwaves and millimeter waves. Eye safety, rather than power absorption inside the head, is the issue here, because the eye acts as a filter to the electromagnetic radiation, allowing only light and near-frequency radiation to enter into it, and the amount of power absorption inside the human body is negligible. Exposure of the eye to high levels of infrared radiation may cause cataractlike diseases, and the maximum allowed transmitter power seems to limit the range to a few meters [31]. If this is the case, safety restrictions will pose severe limitations to the use of infrared in WBMCSs, as far as general applications are concerned. The question in this case is not that there are always problems during system operation (e.g., mobile telephones), but the damage that may be caused if someone looks at the transmitter during operation.

Microwaves and millimeter waves have no special effect on eyes, other than power absorption. In WLANs, antennas do not radiate very near (1 or 2 cm) to the user as in the mobile telephone case, thus enabling power limitations to be less restrictive; also if mobile multimedia terminals are used as they are in PDAs. But if terminals are used in the same form as mobile telephones, then maximum transmitter powers have to be established, similar to those for the current personal communication systems. The standards for safety levels have already been set in the United States and Europe, as the ones used for UHF extend up to 300 MHz (IEEE/ANSI and CENELEC recommendations are the references). Thus, it is left to researchers in this area to extend their work to higher frequencies, by evaluating SAR (the amount of power dissipated per unit of mass) levels inside the head (or other parts of the human body very near the radiating system), from which maximum transmitter powers will be established. This may not be as straightforward as it seems, however, because the calculation of SAR is usually done by solving integral or differential equations using numerical methods (method of moments or finite difference), which require models of the head made of small elements (e.g., cubes) with dimensions of the order of a tenth of the wavelength. This already requires powerful computer resources (in memory and CPU time) for frequencies in the high UHF band, and may limit the possibility of analyzing frequencies much higher than UHF. On the other hand, the higher the frequency, the smaller the penetration of radio waves into the human body, hence making it possible to have models of only some centimeters deep. This is an area for further research.

1.2.5 ATM-Based Wireless (Mobile) Broadband Multimedia Systems

Recent study of MBS, ATM, and ATM-oriented MBS has drawn the attention of several researchers [34–44]. With enormous complexity of managing and operating the many different types of networks now in use, the door is open for finding a common platform—a network on which all established services can be supported and which will allow new services to be introduced without needing new networks on which to run them. The answer seems to be ATM. This is the technology being defined and standardized for B-ISDN. Thus, ATM, when adequately modified, is also an answer for the future mobile wireless broadband multimedia systems.

ATM is a packet-oriented transmission scheme. The transmission path of the packets of constant length, the so-called ATM cells, is established during connection setup between the two endpoints by assigning a virtual channel. At this time, the necessary resources are provided and the logical channels are assigned. All packets of a virtual channel are carried over the same path. The transmission capacity of the virtual channel is characterized by the parameters, mean bit rate and peak bit rate during connection setup. ATM cells are generated according to the need of the data source. Thus, ATM is a very good method to meet the dynamic requirements of connections with variable data rates.

MBS is the interface between the fixed ATM net at the base station side and mobile ATM net at the mobile station side. Normally, the ATM net at the mobile station side only consists of one end system. For every end station, it is possible to operate several virtual channels with different data rates at the same time.

A conceptual view of the ATM-type broadband communications network is shown in Figure 1.5. The most important benefit of ATM is its flexibility; it is used for the new high bit rate services, which are either VBR or burst traffic. Several factors that tend to favor the use of ATM cell transport in MBS are as follows:

- Flexible bandwidth allocation and service-type selection for a range of applications;
- Efficient multiplexing of traffic from bursty data and multimedia sources;
- End-to-end provisioning of broadband services over wired and wireless networks;
- Suitability of available ATM switching for intercell switching;
- Improved service reliability with packet-switching techniques; and
- Ease of interfacing with wired B-ISDN systems.

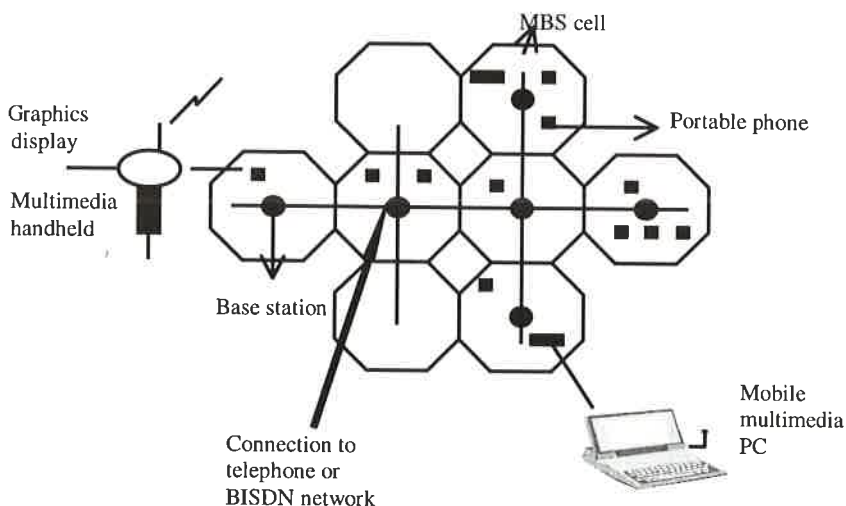


Figure 1.5 Conceptual view of MBS.

Taking the above points into consideration, adoption of ATM-compatible, fixed-length, cell-relay format for MBS is recommended. A possible ATM-compatible MBS approach is shown in Figure 1.6. With this approach, the 48-byte ATM cell payload becomes the basic unit of data within the MBS network. Within MBS, specific protocol layers (e.g., data link and medium-access control layer) are added to the ATM payload as required, and replaced by ATM headers before entering the fixed network [41]. The use of ATM switching for intercell traffic also avoids the crucial problem of developing a new backbone network with sufficient throughput to support intercommunication among large numbers of small cells. ATM multiplexers are used to combine traffic from several base stations into a single ATM port.

For a seamless internetworking mechanism with the wired broadband network, it is vital to have the MBS protocol layering harmonized with the ATM stack. Figure 1.7 shows a protocol reference model. In this approach, new wireless channel-specific physical, medium-access control, and data link layers are added below the ATM network layer. This means that regular network layer and control services such as call setup, virtual channel identifier/virtual path identifier (VCI/VPI) addressing, cell prioritization, and flow-control indication will continue to be used by mobile services. The baseline ATM network and signaling protocol are specified to particular mobility-related functions, such as address registration (roaming), broadcasting, handoff, and so forth.

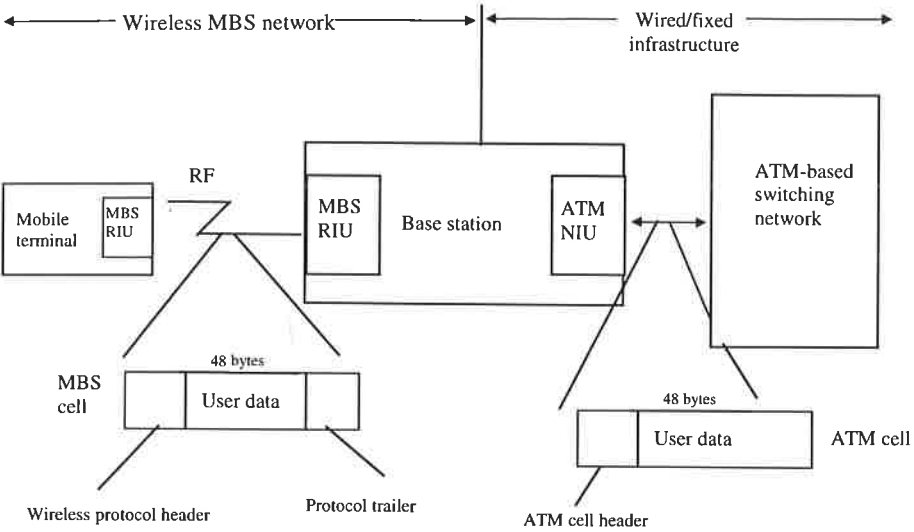


Figure 1.6 ATM-compatible MBS approach. Terms: radio interface unit (RIU); network interface unit (NIU).

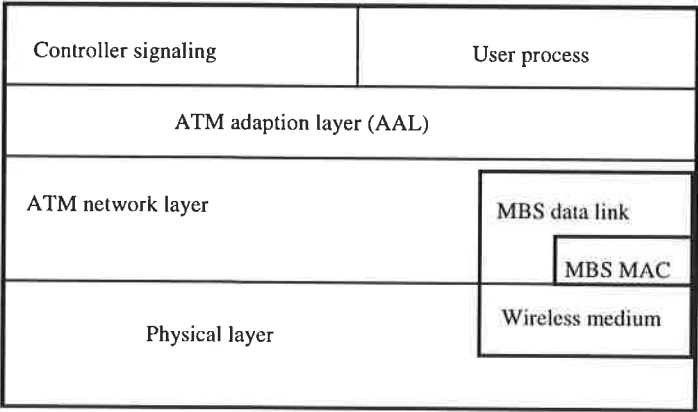


Figure 1.7 Relation of wireless network protocol layers.

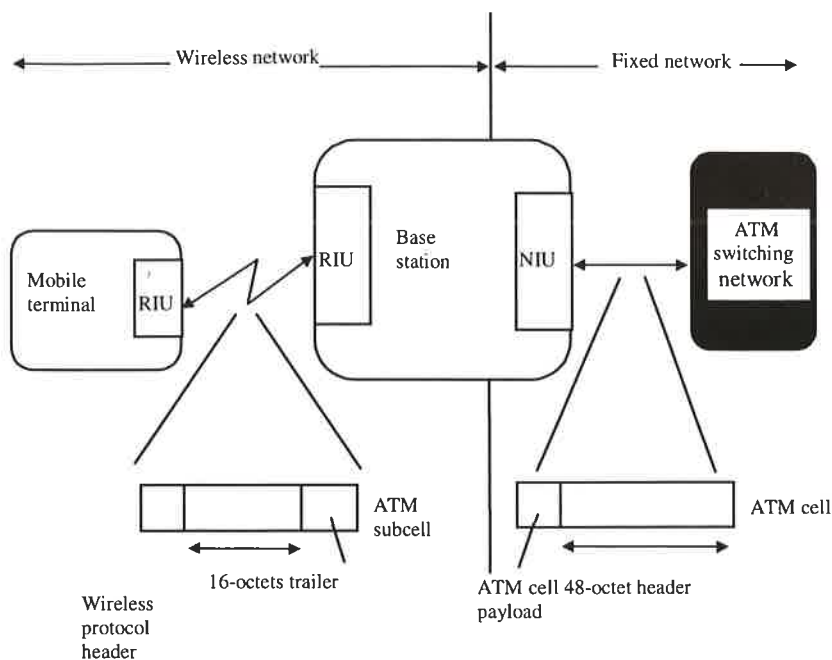


Figure 1.8 Conceptual view of an ATM-compatible air interface.

The payload of an ATM cell consists of 48 octets, meaning that without any segmentation, packet lengths of 384 bits would have to be transmitted over the mobile radio channel. Higher throughput levels over a fading mobile radio channel are achieved when using a packet of smaller lengths. So a suitable integer submultiple of a cell is chosen as the basic unit of data over the wireless radio medium. Based on the system parameters, a submultiple cell size of 16 octets or 128 bits is used as an appropriate value.

The air interface is terminated directly at the base station, as illustrated in Figure 1.8. Also shown is the segmentation of an ATM cell into its three subcells and the addition of an extra wireless ATM header.

1.3 MULTIPATH PROPAGATION

One of the basic reasons to use OFDM is the efficient way it can handle multipath propagation. To understand the effects of multipath fading, this section contains an overview of the most relevant parameters and models.

1.3.1 Multipath Channel Models

Fundamental work on modeling of the indoor radio channel is published in [45, 46]. One of the results of this work, which is also supported by many other measurements reported in the literature, is that the average received multipath power is an exponentially decaying function of the excess delay. Further, the amplitudes of individual multipath components are Rayleigh distributed. This observation has led to simplified channel models as used in [47, 48]. These models assume a fixed number of paths with equidistant delays. The path amplitudes are independent Rayleigh variables, while the path phases are uniformly distributed. Figure 1.9 shows an example of an average and an instantaneous power delay profile that were generated using this approach.

Compared to the more extensive models in [45, 46], the simplified model of [47, 48] may give somewhat optimistic results because the number of multipath components is fixed to the maximum possible amount. In the models of [45, 46], the number of paths is random. Paths arrive in clusters with Poisson distributed arrival times. Within a cluster, the path amplitudes are independent Rayleigh variables. The average power delay profile, averaged over a large number of channels, is an exponentially decaying function, just as for the models in [47, 48]. Thus, the only difference between the models is that the instantaneous power delay profiles have a slightly different shape, and channels generated by the method of [47, 48] generally show more multipath components than channels generated according to [45, 46]. This may give a somewhat optimistic diversity effect, because diversity is proportional to the number of paths. It probably does not make a difference when the models are used to determine the delay spread tolerance of a particular transmission system, however, because that is not depending on the number of paths, but rather on the amount of power in paths exceeding a certain excess delay. So, the channel models of [47, 48] seem a good basis to determine the delay spread tolerance of different modulation schemes. An additional advantage is that the simulation complexity is much lower than that of the models of [45, 46].

One of the key parameters in the design of a transmission system is the maximum delay spread value that it has to tolerate. For an idea of what typical delay spread figures are, the next section presents some measurement results obtained from the literature [49- 63].

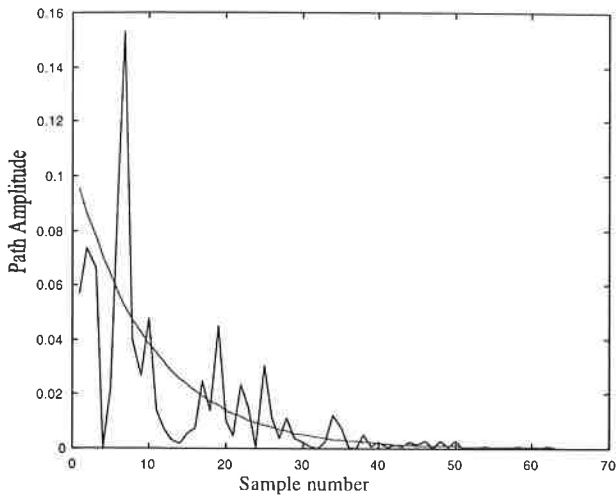


Figure 1.9 Example of generated average and instantaneous power delay profiles for a delay spread of 10 sampling intervals.

1.3.2 Delay Spread Values

Tables 1.3 to 1.5 summarize some delay spread results obtained from literature for frequencies from 800 MHz to 6 GHz. Two delay spread values are given; the *median* delay spread is the 50% value, meaning that 50% of all channels has a delay spread that is lower than the median value. Clearly, the median value is not so interesting for designing a wireless link, because there you want to guarantee that the link works for at least 90% or 99% of all channels. Therefore, the second column gives the measured *maximum* delay spread values. The reason to use maximum delay spread instead of a 90% or 99% value is that many papers only mention the maximum value. From the papers that do present cumulative distribution functions of their measured delay spreads, we can deduce that the 99% value is only a few percent smaller than the maximum measured delay spread.

Table 1.3
Measured delay spreads in frequency range of 800 MHz to 1.5 GHz.

Median delay spread [ns]	Maximum delay spread [ns]	Reference	Remarks
25	50	[46]	Office building
30	56	[58]	Office building
27	43	[59]	Office building
11	58	[60]	Office building
35	80	[61]	Office building
40	90		Shopping center
80	120		Airport
120	180		Factory
50	129	[62]	Warehouse
120	300		Factory

Table 1.4
Measured delay spreads in frequency range of 1.8 to 2.4 GHz

Median delay spread [ns]	Maximum delay spread [ns]	Reference	Remarks
40	120	[49]	Large building (New York stock exchange)
40	95	[50]	Office building
40	150	[51]	Office building
60	200	[54]	Shopping center
106	270		Laboratory
19	30	[55]	Office building: single room only
20	65	[56]	Office building
30	75		Cafeteria
105	170		Shopping center
30	56	[58]	Office building
25	30	[63]	Office building: single room only

Table 1.5
Measured delay spreads in frequency range of 4 to 6 GHz

Median delay spread [ns]	Maximum delay spread [ns]	Reference	Remarks
40	120	[49]	Large building (e.g., Stock Exchange)
50 35 10	60 55 35	[52]	Office building Meeting room (5m x 5m) with metal walls Single room with stone walls
40	130	[51]	Office building
40 65 25	120 125 65	[53]	Indoor sports arena Factory Office building
20	30	[63]	Office building: single room only

Interesting results that can be derived from the reported measurements are:

- Measurements done simultaneously on different frequencies show that there is no significant difference in delay spread in the frequency range of 800 MHz to 6 GHz.
- The delay spread is related to the building size; largest delay spreads (up to 270 ns) were measured in large buildings like shopping centers and factories. The reason for this phenomenon is that reflections can have larger delays if the distances between transmitter and reflectors (walls or other objects) are larger.
- For most office buildings, the maximum delay spread is in the range of 40 to 70 ns. Smaller delay spreads around 30 ns occur when both transmitter and receiver are within the same room. Delay spreads of 100 ns or more are seen only in office buildings with large rooms that are some tens of meters in diameter.
- Even small rooms (5m by 5m) can give significant delay spreads around 50 ns when there are metal walls [52]. The reason for this is that reflections from a metal wall only experience a small attenuation, so multiple reflections with significant delays still have considerable power, which increases the delay spread.

1.4 TIME VARIATION OF THE CHANNEL

To classify the time characteristics of the channel [1], the coherence time and Doppler spread are important parameters. The coherence time is the duration over which the channel characteristics do not change significantly. The time variations of the channel are evidenced as a Doppler spread in the frequency domain, which is determined as the width of the spectrum when a single sinusoid (constant envelope) is transmitted. Both

the time correlation function $\varphi_c(\Delta t)$ and the Doppler power spectrum $S_c(f)$ can be related to each other by applying the Fourier transform.

The range of values of the frequency f over which $S_c(f)$ is essentially nonzero is called the Doppler spread B_d of the channel. Because $S_c(f)$ is related to $\varphi_c(\Delta t)$ by the Fourier transform, the reciprocal of B_d is a measure of the coherence time $(\Delta t)_c$ of the channel; that is, $1/B_d$.

The coherence time $(\Delta t)_c$ is a measure of the width of the time correlation function. Clearly, a slow-changing channel has a large coherence time, or equivalently, a small Doppler spread. The rapidity of the fading can now be determined either from the correlation function $\varphi_c(\Delta t)$ or from the Doppler power spectrum $S_c(f)$. This implies that either the channel parameters $(\Delta t)_c$ or B_d can be used to characterize the rapidity of the fading. If the bit time T_b is large compared with the coherence time, then the channel is subject to fast fading. When selecting a bit duration that is smaller than the coherence time of the channel, the channel attenuation and phase shift are essentially fixed for the duration of at least one signaling interval. In this case, the channel is slowly fading or quasistatic. Like coherence bandwidth, there is no exact relationship between coherence time and Doppler spread. If the coherence time $(\Delta t)_c$ is defined as the time over which the time correlation function is above 0.5, $(\Delta t)_c$ is approximately given by $9/(16\pi B_d)$ [79, 80].

Doppler spread is caused by the differences in Doppler shifts of different components of the received signal, if either the transmitter or receiver is in motion. The frequency shift is related to the spatial angle between the direction of arrival of that component and the direction of vehicular motion. If a vehicle is moving at a constant speed V along the X axis, the Doppler shift f_m of the m th plane-wave component is given by $f_m = (V/\lambda)\cos(\alpha_m)$ [81], where α_m is the arrival angle of the m th plane-wave component relative to the direction of movement of the vehicle. It is worth noting here that waves arriving from ahead of the vehicle experience a positive Doppler shift, while those arriving from behind the vehicle have a negative shift. The maximum Doppler shift occurs at $\alpha_m = 0$, assuming $T_b \gg (\Delta t)_c$ for fast-fading channel and $T_b \ll (\Delta t)_c$ for slow-fading channel.

Indoor measurements [82] show that in any fixed location, temporal variations in the received signal envelope caused by movement of personnel and machinery are slow, having a maximum Doppler spread of about 6 Hz.

1.5 HISTORY OF OFDM

OFDM is a special case of multicarrier transmission, where a single datastream is transmitted over a number of lower rate subcarriers. It is worth mentioning here that OFDM can be seen as either a modulation technique or a multiplexing technique. One

of the main reasons to use OFDM is to increase the robustness against frequency selective fading or narrowband interference. In a single carrier system, a single fade or interferer can cause the entire link to fail, but in a multicarrier system, only a small percentage of the subcarriers will be affected. Error correction coding can then be used to correct for the few erroneous subcarriers. The concept of using parallel data transmission and frequency division multiplexing was published in the mid-1960s [64, 65]. Some early development is traced back to the 1950s [66]. A U.S. patent was filed and issued in January, 1970 [67].

In a classical parallel data system, the total signal frequency band is divided into N nonoverlapping frequency subchannels. Each subchannel is modulated with a separate symbol and then the N subchannels are frequency-multiplexed. It seems good to avoid spectral overlap of channels to eliminate interchannel interference. However, this leads to inefficient use of the available spectrum. To cope with the inefficiency, the ideas proposed from the mid-1960s were to use parallel data and FDM with overlapping subchannels, in which each carrying a signaling rate b is spaced b apart in frequency to avoid the use of high-speed equalization and to combat impulsive noise and multipath distortion, as well as to fully use the available bandwidth.

Figure 1.10 illustrates the difference between the conventional nonoverlapping multicarrier technique and the overlapping multicarrier modulation technique. As shown in Figure 1.10, by using the overlapping multicarrier modulation technique, we save almost 50% of bandwidth. To realize the overlapping multicarrier technique, however we need to reduce crosstalk between subcarriers, which means that we want orthogonality between the different modulated carriers.

The word orthogonal indicates that there is a precise mathematical relationship between the frequencies of the carriers in the system. In a normal frequency-division multiplex system, many carriers are spaced apart in such a way that the signals can be received using conventional filters and demodulators. In such receivers, guard bands are introduced between the different carriers and in the frequency domain which results in a lowering of spectrum efficiency.

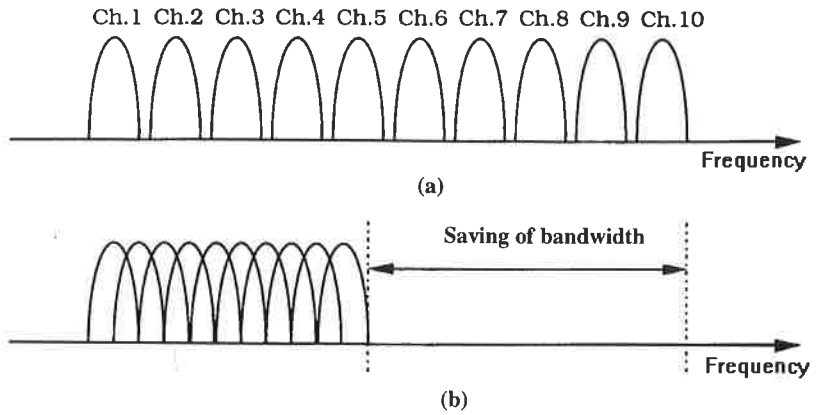


Figure 1.10 Concept of OFDM signal: (a) Conventional multicarrier technique, and (b) orthogonal multicarrier modulation technique.

It is possible, however, to arrange the carriers in an OFDM signal so that the sidebands of the individual carriers overlap and the signals are still received without adjacent carrier interference. To do this the carriers must be mathematically orthogonal. The receiver acts as a bank of demodulators, translating each carrier down to DC, with the resulting signal integrated over a symbol period to recover the raw data. If the other carriers all beat down the frequencies that, in the time domain, have a whole number of cycles in the symbol period T , then the integration process results in zero contribution from all these other carriers. Thus, the carriers are linearly independent (i.e., orthogonal) if the carrier spacing is a multiple of $1/T$. Chapter 2 presents in detail the basic principle of OFDM.

Much of the research focuses on the high efficient multicarrier transmission scheme based on “orthogonal frequency” carriers. In 1971, Weinstein and Ebert [68] applied the discrete Fourier transform (DFT) to parallel data transmission systems as part of the modulation and demodulation process. Figure 1.11 (a) shows the spectrum of the individual data of the subchannel. The OFDM signal, multiplexed in the individual spectra with a frequency spacing b equal to the transmission speed of each subcarrier, is shown in Figure 1.11(b). Figure 1.11 shows that at the center frequency of each subcarrier, there is no crosstalks from other channels. Therefore, if we use DFT at the receiver and calculate correlation values with the center of frequency of each subcarrier, we recover the transmitted data with no crosstalk. In addition, using the DFT-based multicarrier technique, frequency-division multiplex is achieved not by bandpass filtering but by baseband processing.

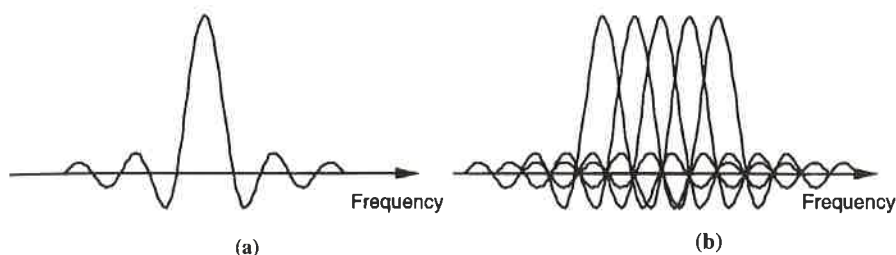


Figure 1.11 Spectra of (a) an OFDM subchannel and (b) an OFDM signal.

Moreover, to eliminate the banks of subcarrier oscillators and coherent demodulators required by frequency-division multiplex, completely digital implementations could be built around special-purpose hardware performing the fast Fourier transform (FFT), which is an efficient implementation of the DFT. Recent advances in very-large-scale integration (VLSI) technology make high-speed, large-size FFT chips commercially affordable. Using this method, both transmitter and receiver are implemented using efficient FFT techniques that reduce the number of operations from N^2 in DFT down to $M\log N$ [69].

In the 1960s, the OFDM technique was used in several high-frequency military systems such as KINEPLEX [66], ANDEFT [70], and KATHRYN [71]. For example, the variable-rate data modem in KATHRYN was built for the high-frequency band. It used up to 34 parallel low-rate phase-modulated channels with a spacing of 82 Hz.

In the 1980s, OFDM was studied for high-speed modems, digital mobile communications, and high-density recording. One of the systems realized the OFDM techniques for multiplexed QAM using DFT [72], and by using pilot tone, stabilizing carrier and clock frequency control and implementing trellis coding are also implemented [73]. Moreover, various-speed modems were developed for telephone networks [74].

In the 1990s, OFDM was exploited for wideband data communications over mobile radio FM channels, high-bit-rate digital subscriber lines (HDSL; 1.6 Mbps), asymmetric digital subscriber lines (ADSL; up to 6 Mbps), very-high-speed digital subscriber lines (VDSL; 100 Mbps), digital audio broadcasting (DAB), and high-definition television (HDTV) terrestrial broadcasting [75–80].

The OFDM transmission scheme has the following key advantages:

- OFDM is an efficient way to deal with multipath; for a given delay spread, the implementation complexity is significantly lower than that of a single carrier system with an equalizer.
- In relatively slow time-varying channels, it is possible to significantly enhance the capacity by adapting the data rate per subcarrier according to the signal-to-noise ratio of that particular subcarrier.

- OFDM is robust against narrowband interference, because such interference affects only a small percentage of the subcarriers.
- OFDM makes single-frequency networks possible, which is especially attractive for broadcasting applications.

On the other hand, OFDM also has some drawbacks compared with single-carrier modulation:

- OFDM is more sensitive to frequency offset and phase noise.
- OFDM has a relatively large peak-to-average power ratio, which tends to reduce the power efficiency of the RF amplifier.

1.6 PREVIEW OF THE BOOK

This book consists of 10 chapters. It covers all the necessary elements to achieve the OFDM-based WBMCS. In Chapter 2, the basics of OFDM are presented. It is explained how an OFDM signal is formed using the inverse fast Fourier transform, how the cyclic extension helps to mitigate the effects of multipath, and how windowing can limit the out-of-band radiation. Basic design rules are given how to choose the OFDM parameters, given a required bandwidth, multipath delay spread, and maximum Doppler spread.

In Chapter 3, we explain how coding and interleaving can be used to mitigate the effects of frequency-selective fading channels. *Quadrature amplitude modulation* is introduced as an appropriate modulation technique for the OFDM subcarriers.

Synchronization of the symbol clock and carrier frequency is the subject of Chapter 4. First, we discuss the sensitivity of OFDM to synchronization errors. Then, we describe different synchronization techniques. Special attention is given to packet transmission, which requires rapid synchronization at the beginning of each packet with a minimum of training overhead.

Chapter 5 describes channel estimation; that is, estimating the reference phases and amplitudes of all subcarriers. Various coherent and differential techniques are explained in conjunction with their relative merits on signal-to-noise ratio performance, overhead, and buffering delay.

The relatively large peak-to-average power (PAP) ratio of OFDM is the subject of Chapter 6. The distribution of the PAP ratio is shown, and various methods to reduce the PAP ratio are explained. This chapter demonstrates that for an arbitrary number of subcarriers, the PAP ratio can be reduced to about 5 dB without a significant loss in performance. This means that the required backoff—and hence the efficiency—of an RF power amplifier for an OFDM system is not much different from that for a single-carrier QPSK system.

In Chapter 7, we explain the basics of direct-sequence and frequency-hopping CDMA, which is especially helpful in understanding the following chapters on combinations of OFDM and CDMA.

Chapter 8 describes multicarrier CDMA. Different techniques with their transmitter and receiver architectures are introduced. Advantages and disadvantages compared with other CDMA techniques are discussed.

Orthogonal FDMA and frequency-hopping CDMA are the subjects of Chapter 9. It shows how OFDM and frequency hopping can be combined to get a multiple-access system with similar advantages as direct-sequence CDMA.

Finally, some applications of OFDM systems are described in Chapter 10: digital audio broadcasting, digital video broadcasting, wireless ATM in the Magic WAND project, and the new IEEE 802.11 and ETSI BRAN OFDM standards.

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Chapter 2

OFDM Basics

2.1 INTRODUCTION

The basic principle of OFDM is to split a high-rate datastream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. Because the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Intersymbol interference is eliminated almost completely by introducing a guard time in every OFDM symbol. In the guard time, the OFDM symbol is cyclically extended to avoid intercarrier interference. This whole process of generating an OFDM signal and the reasoning behind it are described in detail in sections 2.2 to 2.4.

In OFDM system design, a number of parameters are up for consideration, such as the number of subcarriers, guard time, symbol duration, subcarrier spacing, modulation type per subcarrier, and the type of forward error correction coding. The choice of parameters is influenced by system requirements such as available bandwidth, required bit rate, tolerable delay spread, and Doppler values. Some requirements are conflicting. For instance, to get a good delay spread tolerance, a large number of subcarriers with a small subcarrier spacing is desirable, but the opposite is true for a good tolerance against Doppler spread and phase noise. These design issues are discussed in Section 2.5. Section 2.6 gives an overview of OFDM signal processing functions, while Section 2. ends this chapter with a complexity comparison of OFDM versus single-carrier systems.

2.2 GENERATION OF SUBCARRIERS USING THE IFFT

An OFDM signal consists of a sum of subcarriers that are modulated by using *phase shift keying (PSK)* or *quadrature amplitude modulation (QAM)*. If d_i are the complex

QAM symbols, N_s is the number of subcarriers, T the symbol duration, and f_c the carrier frequency, then one OFDM symbol starting at $t = t_s$ can be written as

$$s(t) = \text{Re} \left\{ \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \exp(j2\pi(f_c - \frac{i+0.5}{T})(t-t_s)) \right\}, \quad t_s \leq t \leq t_s + T \quad (2.1)$$

$$s(t) = 0, \quad t < t_s \wedge t > t_s + T$$

In the literature, often the equivalent complex baseband notation is used, which is given by (2.2). In this representation, the real and imaginary parts correspond to the in-phase and quadrature parts of the OFDM signal, which have to be multiplied by a cosine and sine of the desired carrier frequency to produce the final OFDM signal. Figure 2.1 shows the operation of the OFDM modulator in a block diagram.

$$s(t) = \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \exp(j2\pi \frac{i}{T}(t-t_s)) \quad , \quad t_s \leq t \leq t_s + T \quad (2.2)$$

$$s(t) = 0, \quad t < t_s \wedge t > t_s + T$$

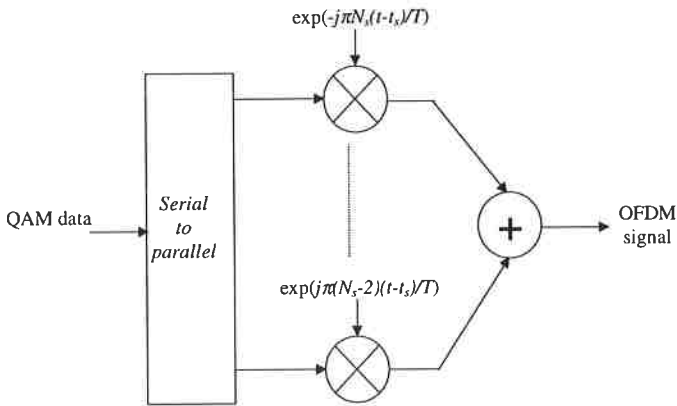


Figure 2.1 OFDM modulator.

As an example, Figure 2.2 shows four subcarriers from one OFDM signal. In this example, all subcarriers have the same phase and amplitude, but in practice the amplitudes and phases may be modulated differently for each subcarrier. Note that each subcarrier has exactly an integer number of cycles in the interval T , and the number of cycles between adjacent subcarriers differs by exactly one. This property accounts for

the orthogonality between the subcarriers. For instance, if the j th subcarrier from (2.2) is demodulated by downconverting the signal with a frequency of j/T and then integrating the signal over T seconds, the result is as written in (2.3). By looking at the intermediate result, it can be seen that a complex carrier is integrated over T seconds. For the demodulated subcarrier j , this integration gives the desired output $d_{j+N/2}$ (multiplied by a constant factor T), which is the QAM value for that particular subcarrier. For all other subcarriers, the integration is zero, because the frequency difference $(i-j)/T$ produces an integer number of cycles within the integration interval T , such that the integration result is always zero.

$$\begin{aligned}
 & \int_{t_s}^{t_s+T} \exp(-j2\pi \frac{j}{T}(t-t_s)) \sum_{i=-\frac{N_t}{2}}^{\frac{N_t}{2}-1} d_{i+N_t/2} \exp(j2\pi \frac{i}{T}(t-t_s)) dt \\
 &= \sum_{i=-\frac{N_t}{2}}^{\frac{N_t}{2}-1} d_{i+N_t/2} \int_{t_s}^{t_s+T} \exp(j2\pi \frac{i-j}{T}(t-t_s)) dt = d_{j+N_t/2} T
 \end{aligned} \tag{2.3}$$

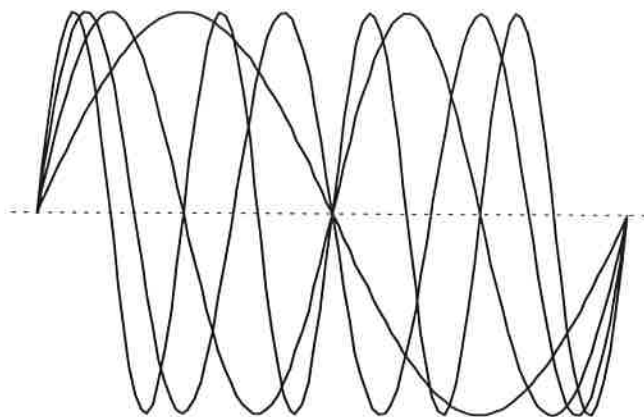


Figure 2.2 Example of four subcarriers within one OFDM symbol.

The orthogonality of the different OFDM subcarriers can also be demonstrated in another way. According to (2.1), each OFDM symbol contains subcarriers that are nonzero over a T -second interval. Hence, the spectrum of a single symbol is a convolution of a group of Dirac pulses located at the subcarrier frequencies with the spectrum of a square pulse that is one for a T -second period and zero otherwise. The amplitude spectrum of the square pulse is equal to $\text{sinc}(\pi f T)$, which has zeros for all frequencies f that are an integer multiple of $1/T$. This effect is shown in Figure 2.2, which shows the overlapping sinc spectra of individual subcarriers. At the maximum of each subcarrier spectrum, all other subcarrier spectra are zero. Because an OFDM

receiver essentially calculates the spectrum values at those points that correspond to the maxima of individual subcarriers, it can demodulate each subcarrier free from any interference from the other subcarriers. Basically, Figure 2.3 shows that the OFDM spectrum fulfills Nyquist’s criterium for an intersymbol interference free pulse shape. Notice that the pulse shape is present in the frequency domain and not in the time domain, for which the Nyquist criterium usually is applied. Therefore, instead of intersymbol interference (ISI), it is intercarrier interference (ICI) that is avoided by having the maximum of one subcarrier spectrum correspond to zero crossings of all the others.

The complex baseband OFDM signal as defined by (2.2) is in fact nothing more than the inverse Fourier transform of N_s QAM input symbols. The time discrete equivalent is the inverse discrete Fourier transform (IDFT), which is given by (2.4), where the time t is replaced by a sample number n . In practice, this transform can be implemented very efficiently by the inverse fast Fourier transform (IFFT). An N point IDFT requires a total of N^2 complex multiplications—which are actually only phase rotations. Of course, there are also additions necessary to do an IDFT, but since the hardware complexity of an adder is significantly lower than that of a multiplier or phase rotator, only the multiplications are used here for comparison. The IFFT drastically reduces the amount of calculations by exploiting the regularity of the operations in the IDFT. Using the radix-2 algorithm, an N -point IFFT requires only $(N/2) \cdot \log_2(N)$ complex multiplications [1]. For a 16-point transform, for instance, the difference is 256 multiplications for the IDFT versus 32 for the IFFT—a reduction by a factor of 8! This difference grows for larger numbers of subcarriers, as the IDFT complexity grows quadratically with N , while the IFFT complexity only grows slightly faster than linear.

$$s(n) = \sum_{i=0}^{N_s-1} d_i \exp(j2\pi \frac{in}{N}) \tag{2.4}$$

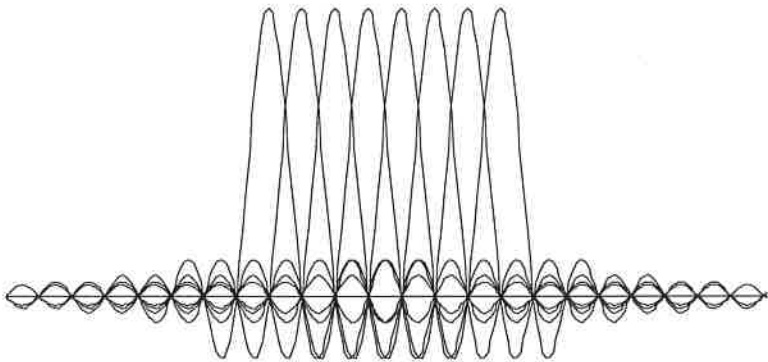


Figure 2.3 Spectra of individual subcarriers.

The number of multiplications in the IFFT can be reduced even further by using a radix-4 algorithm. This technique makes use of the fact that in a four-point IFFT, there are only multiplications by $\{1, -1, j, -j\}$, which actually do not need to be implemented by a full multiplier, but rather by a simple add or subtract and a switch of real and imaginary parts in the case of multiplications by j or $-j$. In the radix-4 algorithm, the transform is split into a number of these trivial four-point transforms, and non-trivial multiplications only have to be performed between stages of these four-point transforms. In this way, an N -point FFT using the radix-4 algorithm requires only $(3/8)N(\log_2 N - 2)$ complex multiplications or phase rotations and $N \log_2 N$ complex additions [1]. For a 64-point FFT, for example, this means 96 rotations and 384 additions, or 1.5 and 6 rotations and additions per sample, respectively.

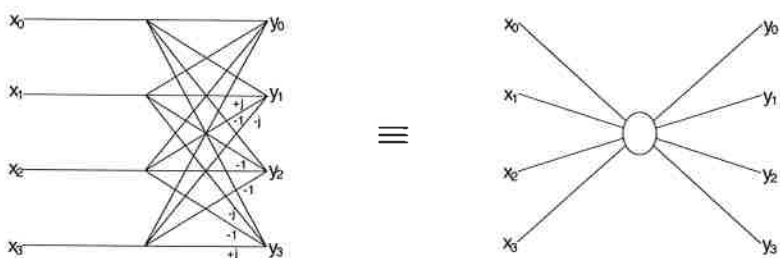


Figure 2.4 The radix-4 butterfly.

Figure 2.4 shows the four-point IFFT, which is known as the *radix-4 butterfly* that forms the basis for constructing larger IFFT sizes [1]. Four input values x_0 to x_3 are transformed into output values y_0 to y_3 by simple additions and trivial phase rotations. For instance, y_1 is given by $x_0 + jx_1 - x_2 - jx_3$, which can be calculated by doing four additions plus a few additional I/Q swappings and inversions to account for the multiplications by j and -1 .

The radix-4 butterfly can be used to efficiently build an IFFT with a larger size. For instance, a 16-point IFFT is depicted in Figure 2.5. The 16-point IFFT contains two stages with four radix-4 butterflies, separated by an intermediate stage where the 16 intermediate results are phase rotated by the twiddle factor ω^j , which is defined as $\exp(j2\pi i/N)$. Notice that for $N=16$, rotation by the twiddle factor ω^j reduces to a trivial operation for $i=0, 4, 8$ and 12 , where ω^j is $1, j, -1$ and $-j$, respectively. Taking this into account, the 16-point IFFT actually contains only eight non-trivial phase rotations, which is a factor of 32 smaller than the amount of phase rotations for the IDFT. These non-trivial phase rotations largely determine the implementation complexity, because the complexity of a phase rotation or complex multiplication is an order of magnitude larger than the complexity of an addition.

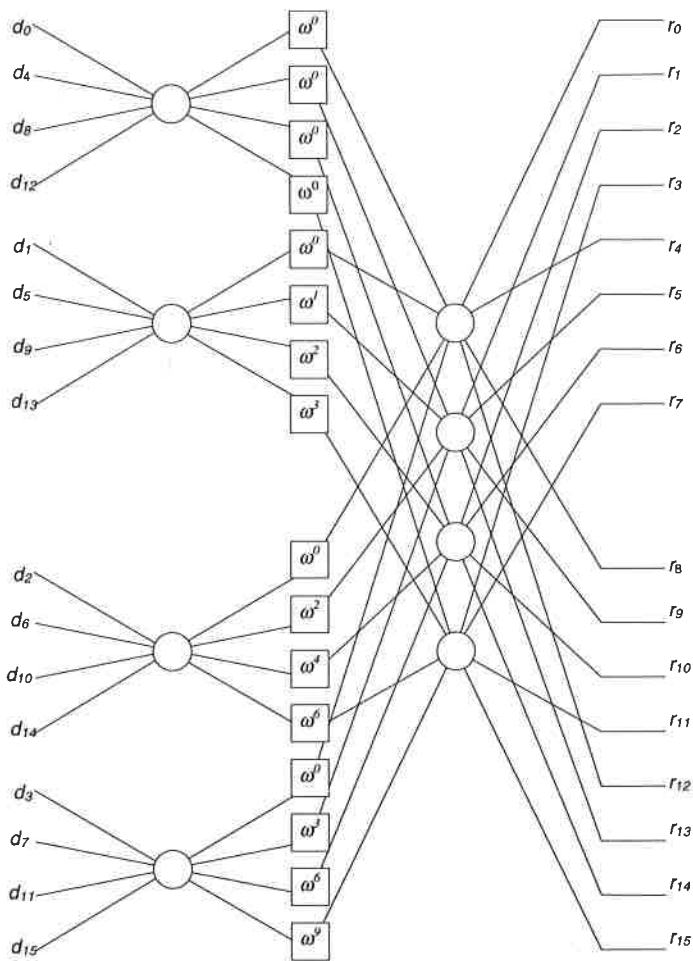


Figure 2.5 16-point IFFT using the radix-4 algorithm.

As an example of how to generate an OFDM symbol, let us assume that we want to transmit eight binary values { 1 1 1 -1 1 1 -1 1 } on eight subcarriers. The IDFT or IFFT then has to calculate:

$$\frac{1}{8} \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & \frac{1}{2}\sqrt{2}(1+j) & j & \frac{1}{2}\sqrt{2}(-1+j) & -1 & \frac{1}{2}\sqrt{2}(-1-j) & -j & \frac{1}{2}\sqrt{2}(1-j) \\ 1 & j & -1 & -j & 1 & j & -1 & -j \\ 1 & \frac{1}{2}\sqrt{2}(-1+j) & -j & \frac{1}{2}\sqrt{2}(1+j) & -1 & \frac{1}{2}\sqrt{2}(1-j) & j & \frac{1}{2}\sqrt{2}(-1-j) \\ 1 & -1 & 1 & -1 & 1 & -1 & 1 & -1 \\ 1 & \frac{1}{2}\sqrt{2}(-1-j) & j & \frac{1}{2}\sqrt{2}(1-j) & -1 & \frac{1}{2}\sqrt{2}(1+j) & -j & \frac{1}{2}\sqrt{2}(-1+j) \\ 1 & -j & -1 & j & 1 & -j & -1 & j \\ 1 & \frac{1}{2}\sqrt{2}(1-j) & -j & \frac{1}{2}\sqrt{2}(-1-j) & -1 & \frac{1}{2}\sqrt{2}(-1+j) & j & \frac{1}{2}\sqrt{2}(1+j) \end{bmatrix} \begin{bmatrix} 1 \\ 1 \\ -1 \\ 1 \\ 1 \\ -1 \\ -1 \\ 1 \end{bmatrix} = \frac{1}{8} \begin{bmatrix} 4 \\ \sqrt{2}(1+j(\sqrt{2}-1)) \\ 2+2j \\ -\sqrt{2}(1+j(\sqrt{2}+1)) \\ 0 \\ -\sqrt{2}(1-j(\sqrt{2}+1)) \\ 2-2j \\ \sqrt{2}(1-j(\sqrt{2}-1)) \end{bmatrix} \quad (2.5)$$

The left-hand side of (2.5) contains the IDFT matrix, where every column corresponds to a complex subcarrier with a normalized frequency ranging from -4 to +3. The right-hand side of (2.5) gives the eight IFFT output samples that form one OFDM symbol. In practice, however, these samples are not enough to make a real OFDM signal. The reason is that there is no oversampling present, which would introduce intolerable aliasing if one would pass these samples through a digital-to-analog converter. To introduce oversampling, a number of zeros can be added to the input data. For instance, eight zeros could be added to the eight input samples of the previous example, after which a 16-point IFFT can be performed to get 16 output samples of a twice-oversampled OFDM signal. Notice that in the complex IFFT as in (2.5), the first half of the rows correspond to positive frequencies while the last half correspond to negative frequencies. Hence, if oversampling is used, the zeros should be added in the middle of the data vector rather than appending them at the end. This ensures the zero data values are mapped onto frequencies close to plus and minus half the sampling rate, while the nonzero data values are mapped onto the subcarriers around 0 Hz. For the data of the previous example, the oversampled input vector would become $\{1 \ 1 \ 1 \ -1 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1 \ 1 \ -1 \ 1\}$.

2.3 GUARD TIME AND CYCLIC EXTENSION

One of the most important reasons to do OFDM is the efficient way it deals with multipath delay spread. By dividing the input datastream in N_s subcarriers, the symbol duration is made N_s times smaller, which also reduces the relative multipath delay spread, relative to the symbol time, by the same factor. To eliminate intersymbol interference almost completely, a guard time is introduced for each OFDM symbol. The guard time is chosen larger than the expected delay spread, such that multipath components from one symbol cannot interfere with the next symbol. The guard time could consist of no signal at all. In that case, however, the problem of intercarrier interference (ICI) would arise. ICI is crosstalk between different subcarriers, which means they are no longer orthogonal. This effect is illustrated in Figure 2.6. In this example, a subcarrier 1 and a delayed subcarrier 2 are shown. When an OFDM receiver

tries to demodulate the first subcarrier, it will encounter some interference from the second subcarrier, because within the FFT interval, there is no integer number of cycles difference between subcarrier 1 and 2. At the same time, there will be crosstalk from the first to the second subcarrier for the same reason.

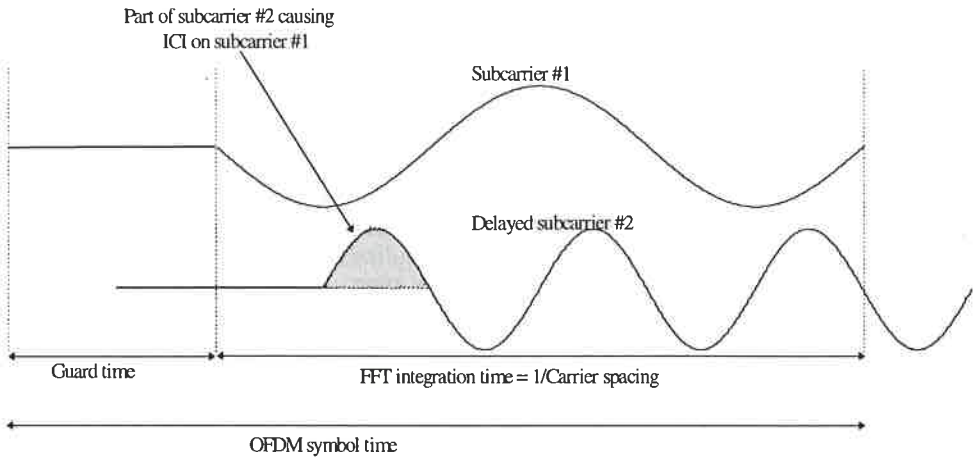


Figure 2.6 Effect of multipath with zero signal in the guard time; the delayed subcarrier 2 causes ICI on subcarrier 1 and vice versa.

To eliminate ICI, the OFDM symbol is cyclically extended in the guard time, as shown in Figure 2.7. This ensures that delayed replicas of the OFDM symbol always have an integer number of cycles within the FFT interval, as long as the delay is smaller than the guard time. As a result, multipath signals with delays smaller than the guard time cannot cause ICI.

As an example of how multipath affects OFDM, Figure 2.8 shows received signals for a two-ray channel, where the dotted curve is a delayed replica of the solid curve. Three separate subcarriers are shown during three symbol intervals. In reality, an OFDM receiver only sees the sum of all these signals, but showing the separate components makes it more clear what the effect of multipath is. From the figure, we can see that the OFDM subcarriers are BPSK modulated, which means that there can be 180-degree phase jumps at the symbol boundaries. For the dotted curve, these phase jumps occur at a certain delay after the first path. In this particular example, this multipath delay is smaller than the guard time, which means there are no phase transitions during the FFT interval. Hence, an OFDM receiver “sees” the sum of pure sine waves with some phase offsets. This summation does not destroy the orthogonality between the subcarriers, it only introduces a different phase shift for each subcarrier. The orthogonality does become lost if the multipath delay becomes larger than the

guard time. In that case, the phase transitions of the delayed path fall within the FFT interval of the receiver. The summation of the sine waves of the first path with the phase modulated waves of the delayed path no longer gives a set of orthogonal pure sine waves, resulting in a certain level of interference.

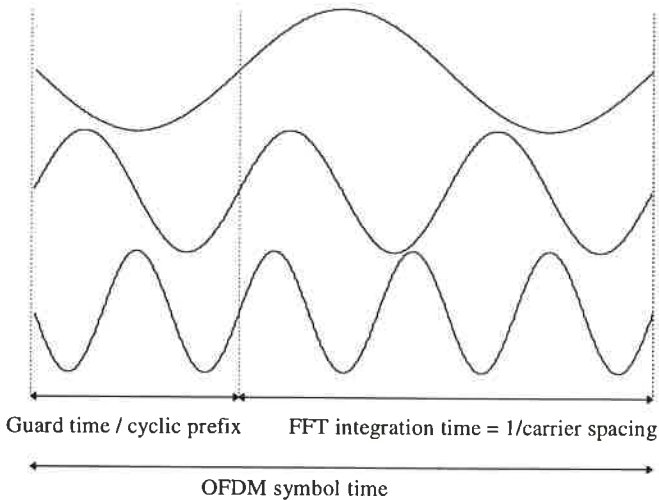


Figure 2.7 OFDM symbol with cyclic extension.

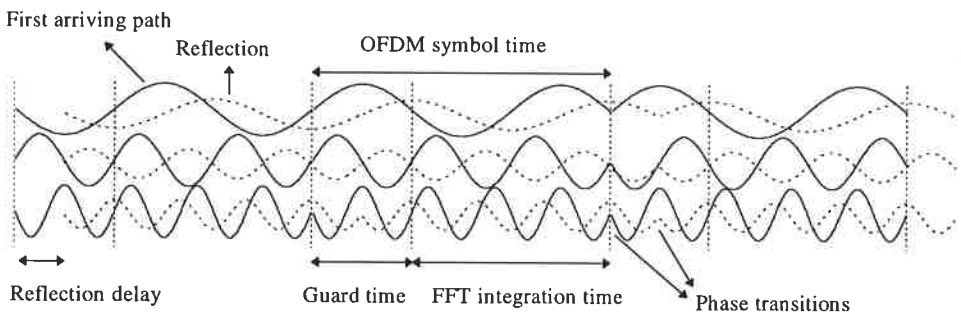


Figure 2.8 Example of an OFDM signal with three subcarriers in a two-ray multipath channel. The dashed line represents a delayed multipath component.

To get an idea what level of interference is introduced when the multipath delay exceeds the guard time, Figure 2.9 depicts three constellation diagrams that were derived from a simulation of an OFDM link with 48 subcarriers, each modulated by using 16-QAM. Figure 2.9(a) shows the undistorted 16-QAM constellation, which is observed whenever the multipath delay is below the guard time. In Figure 2.9(b), the multipath delay exceeds the guard time by a small 3% fraction of the FFT interval.

Hence, the subcarriers are not orthogonal anymore, but the interference is still small enough to get a reasonable received constellation. In Figure 2.9(c), the multipath delay exceeds the guard time by 10% of the FFT interval. The interference is now so large that the constellation is seriously blurred, causing an unacceptable error rate.

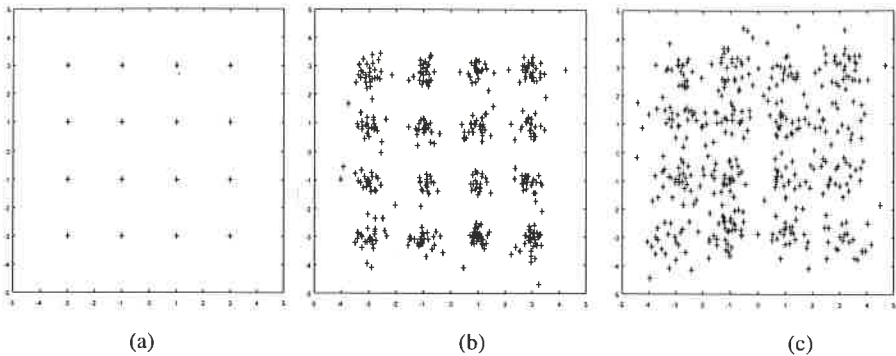


Figure 2.9 16-QAM constellation for a 48-subcarrier OFDM link with a two-ray multipath channel, the second ray being 6 dB lower than the first one. (a) delay < guard time; (b) delay exceeds guard time by 3% of the FFT interval; (c) delay exceeds guard time by 10% of the FFT interval.

2.4 WINDOWING

In the previous sections, it was explained how an OFDM symbol is formed by performing an IFFT and adding a cyclic extension. Looking at an example OFDM signal like in Figure 2.8, sharp phase transitions caused by the modulation can be seen at the symbol boundaries. Essentially, an OFDM signal like the one depicted in Figure 2.8 consists of a number of unfiltered QAM subcarriers. As a result, the out-of-band spectrum decreases rather slowly, according to a sinc function. As an example of this, the spectra for 16, 64, and 256 subcarriers are plotted in Figure 2.10. For larger number of subcarriers, the spectrum goes down more rapidly in the beginning, which is caused by the fact that the sidelobes are closer together. However, even the spectrum for 256 subcarriers has a relatively large -40-dB bandwidth that is almost four times the -3-dB bandwidth.

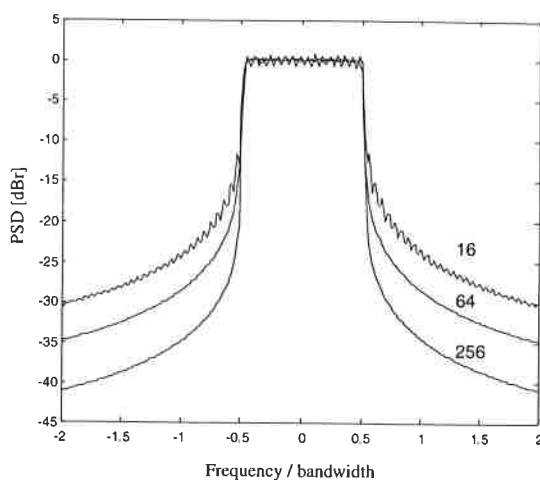


Figure 2.10 Power spectral density (PSD) without windowing for 16, 64, and 256 subcarriers.

To make the spectrum go down more rapidly, windowing can be applied to the individual OFDM symbols. Windowing an OFDM symbol makes the amplitude go smoothly to zero at the symbol boundaries. A commonly used window type is the raised cosine window, which is defined as

$$w(t) = \begin{cases} 0.5 + 0.5 \cos(\pi + t\pi / (\beta T_s)) & 0 \leq t \leq \beta T_s \\ 1.0 & \beta T_s \leq t \leq T_s \\ 0.5 + 0.5 \cos((t - T_s)\pi / (\beta T_s)) & T_s \leq t \leq (1 + \beta)T_s \end{cases} \quad (2.6)$$

Here, T_s is the symbol interval, which is shorter than the total symbol duration because we allow adjacent symbols to partially overlap in the roll-off region. The time structure of the OFDM signal now looks like Figure 2.11.

In equation form, an OFDM symbol starting at time $t = t_s = kT_s$ is defined as

$$s_k(t) = \text{Re} \left\{ w(t - t_s) \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s(k+1/2)} \exp(j2\pi(f_c - \frac{i+0.5}{T})(t - t_s - T_{\text{prefix}})) \right\}, \quad t_s \leq t \leq t_s + T_s(1 + \beta)$$

$$s_k(t) = 0, \quad t < t_s \wedge t > t_s + T_s(1 + \beta)$$
(2.7)

In practice, the OFDM signal is generated as follows: first, N_c input QAM values are padded with zeros to get N input samples that are used to calculate an IFFT. Then, the last T_{prefix} samples of the IFFT output are inserted at the start of the OFDM symbol, and the first $T_{postfix}$ samples are appended at the end. The OFDM symbol is then multiplied by a raised cosine window $w(t)$ to more quickly reduce the power of out-of-band subcarriers. The OFDM symbol is then added to the output of the previous OFDM symbol with a delay of T_s , such that there is an overlap region of βT_s , where β is the rolloff factor of the raised cosine window.

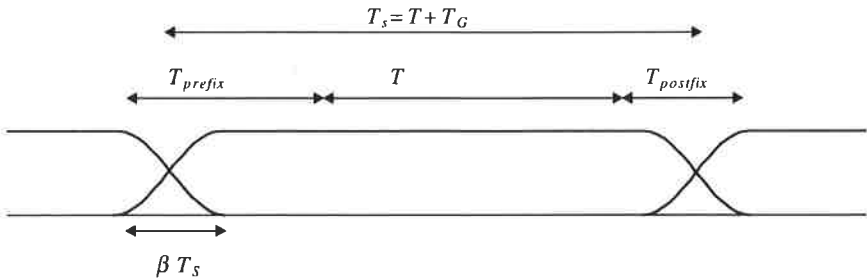


Figure 2.11 OFDM cyclic extension and windowing. T_s is the symbol time, T the FFT interval, T_G the guard time, T_{prefix} the preguard interval, $T_{postfix}$ the postguard interval, and β is the rolloff factor.

Figure 2.12 shows spectra for 64 subcarriers and different values of the rolloff factor β . It can be seen that a rolloff factor of 0.025—so the rolloff region is only 2.5% of the symbol interval—already makes a large improvement in the out-of-band spectrum. For instance, the -40 -dB bandwidth is more than halved to about twice the -3 -dB bandwidth. Larger rolloff factors improve the spectrum further, at the cost, however, of a decreased delay spread tolerance. The latter effect is demonstrated in Figure 2.13, which shows the signal structure of an OFDM signal for a two-ray multipath channel. The receiver demodulates the subcarriers by taking an FFT over the T -second interval between the dotted lines. Although the relative delay between the two multipath signals is smaller than the guard time, ICI and ISI are introduced because of the amplitude modulation in the gray part of the delayed OFDM symbol. The orthogonality between subcarriers as proved by (2.3) only holds when amplitude and phase of the subcarriers are constant during the entire T -second interval. Hence, a rolloff factor of β reduces the effective guard time by βT_s .

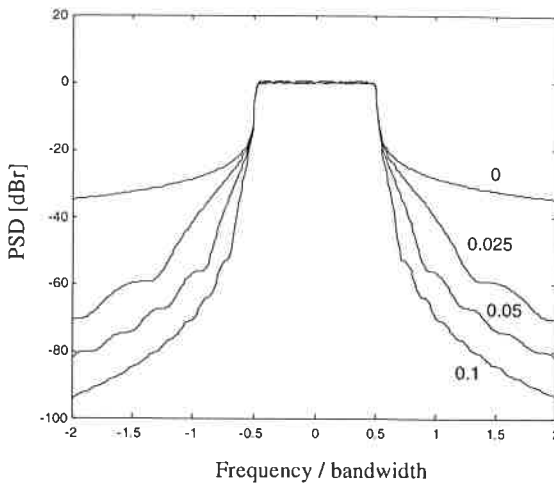


Figure 2.12 Spectra for raised cosine windowing with rolloff factors of 0 (rectangular window), 0.025, 0.05, and 0.1.

Instead of windowing, it is also possible to use conventional filtering techniques to reduce the out-of-band spectrum. Windowing and filtering are dual techniques; multiplying an OFDM symbol by a window means the spectrum is going to be a convolution of the spectrum of the window function with a set of impulses at the subcarrier frequencies. When filtering is applied, a convolution is done in the time domain and the OFDM spectrum is multiplied by the frequency response of the filter. When using filters, care has to be taken not to introduce rippling effects on the envelope of the OFDM symbols over a timespan that is larger than the rolloff region of the windowing approach. Too much rippling means the undistorted part of the OFDM envelope is smaller, and this directly translates into less delay spread tolerance. Notice that digital filtering techniques are more complex to implement than windowing. A digital filter requires at least a few multiplications per sample, while windowing only requires a few multiplications per symbol, for those samples which fall into the rolloff region. Hence, because only a few percent of the samples are in the rolloff region, windowing is an order of magnitude less complex than digital filtering.

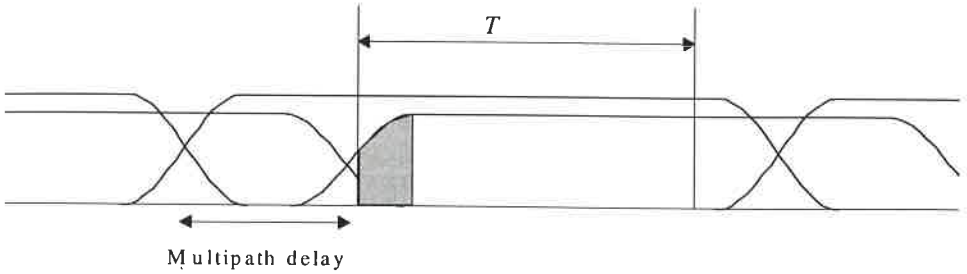


Figure 2.13 OFDM symbol windows for a two-ray multipath channel, showing ICI and ISI, because in the gray part, the amplitude of the delayed subcarriers is not constant.

2.5 CHOICE OF OFDM PARAMETERS

The choice of various OFDM parameters is a tradeoff between various, often conflicting requirements. Usually, there are three main requirements to start with: bandwidth, bit rate, and delay spread. The delay spread directly dictates the guard time. As a rule, the guard time should be about two to four times the root-mean-squared delay spread. This value depends on the type of coding and QAM modulation. Higher order QAM (like 64-QAM) is more sensitive to ICI and ISI than QPSK; while heavier coding obviously reduces the sensitivity to such interference.

Now that the guard time has been set, the symbol duration can be fixed. To minimize the signal-to-noise ratio (SNR) loss caused by the guard time, it is desirable to have the symbol duration much larger than the guard time. It cannot be arbitrarily large, however, because a larger symbol duration means more subcarriers with a smaller subcarrier spacing, a larger implementation complexity, and more sensitivity to phase noise and frequency offset [2], as well as an increased peak-to-average power ratio [3, 4]. Hence, a practical design choice is to make the symbol duration at least five times the guard time, which implies a 1-dB SNR loss because of the guard time.

After the symbol duration and guard time are fixed, the number of subcarriers follows directly as the required -3 -dB bandwidth divided by the subcarrier spacing, which is the inverse of the symbol duration less the guard time. Alternatively, the number of subcarriers may be determined by the required bit rate divided by the bit rate per subcarrier. The bit rate per subcarrier is defined by the modulation type (e.g., 16-QAM), coding rate, and symbol rate.

As an example, suppose we want to design a system with the following requirements:

- Bit rate: 20 Mbps
- Tolerable delay spread: 200 ns
- Bandwidth: < 15 MHz

The delay-spread requirement of 200 ns suggests that 800 ns is a safe value for the guard time. By choosing the OFDM symbol duration 6 times the guard time ($4.8 \mu\text{s}$), the guard time loss is made smaller than 1 dB. The subcarrier spacing is now the inverse of $4.8 - 0.8 = 4 \mu\text{s}$, which gives 250 kHz. To determine the number of subcarriers needed, we can look at the ratio of the required bit rate and the OFDM symbol rate. To achieve 20 Mbps, each OFDM symbol has to carry 96 bits of information ($96/4.8 \mu\text{s} = 20 \text{ Mbps}$). To do this, there are several options. One is to use 16-QAM together with rate $\frac{1}{2}$ coding to get 2 bits per symbol per subcarrier. In this case, 48 subcarriers are needed to get the required 96 bits per symbol. Another option is to use QPSK with rate $\frac{3}{4}$ coding, which gives 1.5 bits per symbol per subcarrier. In this case, 64 subcarriers are needed to reach the 96 bits per symbol. However, 64 subcarriers means a bandwidth of $64 \cdot 250 \text{ kHz} = 16 \text{ MHz}$, which is larger than the target bandwidth. To achieve a bandwidth smaller than 15 MHz, the number of subcarriers needs to be smaller than 60. Hence, the first option with 48 subcarriers and 16-QAM fulfills all the requirements. It has the additional advantage that an efficient 64-point radix-4 FFT/IFFT can be used, leaving 16 zero subcarriers to provide oversampling necessary to avoid aliasing.

An additional requirement that can affect the chosen parameters is the demand for an integer number of samples both within the FFT/IFFT interval and in the symbol interval. For instance, in the above example we want to have exactly 64 samples in the FFT/IFFT interval to preserve orthogonality among the subcarriers. This can be achieved by making the sampling rate $64/4 \mu\text{s} = 16 \text{ MHz}$. However, for that particular sampling rate, there is no integer number of samples with the symbol interval of $4.8 \mu\text{s}$. The only solution to this problem is to change one of the parameters slightly to meet the integer constraint. For instance, the number of samples per symbol can be set to 78, which gives a sampling rate of $78/4.8 \mu\text{s} = 16.25 \text{ MHz}$. Now, the FFT interval becomes $64/16.25 \text{ MHz} = 3.9385 \mu\text{s}$, so both guard time and subcarrier spacing are slightly larger than in the case of the original FFT interval of $4 \mu\text{s}$.

2.6 OFDM SIGNAL PROCESSING

The previous sections described how the basic OFDM signal is formed using the IFFT, adding a cyclic extension and performing windowing to get a steeper spectral rolloff. However, there is more to it to build a complete OFDM modem. Figure 2.14 shows the block diagram of an OFDM modem, where the upper path is the transmitter chain and the lower path corresponds to the receiver chain. In the center we see the IFFT, which modulates a block of input QAM values onto a number of subcarriers. In the receiver, the subcarriers are demodulated by an FFT, which performs the reverse operation of an IFFT. An interesting feature of the FFT/IFFT is that the FFT is almost identical to an IFFT. In fact, an IFFT can be made using an FFT by conjugating input and output of the FFT and dividing the output by the FFT size. This makes it possible to use the same hardware for both transmitter and receiver. Of course, this saving in complexity is only possible when the modem does not have to transmit and receive simultaneously.

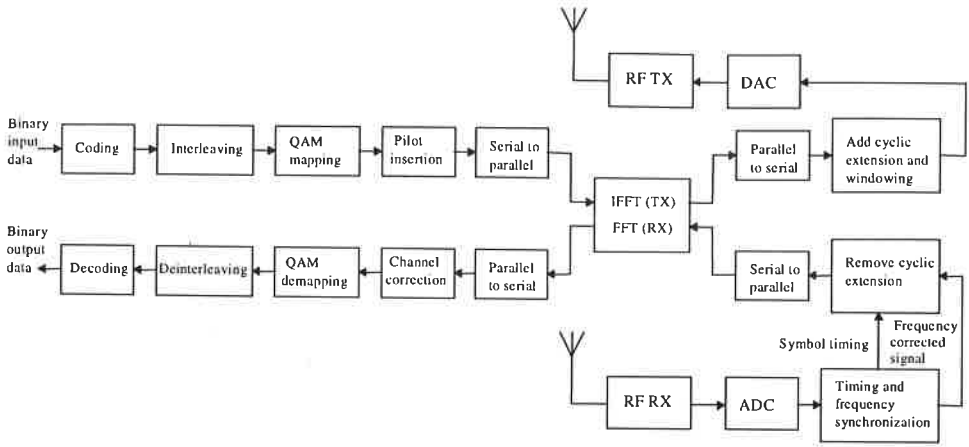


Figure 2.14 Block diagram of an OFDM transceiver.

The functions before the IFFT have not been discussed till now. Binary input data is first encoded by a forward error correction code. The encoded data is then interleaved and mapped onto QAM values. These functions are discussed in more detail in the next chapter.

In the receiver path, after passing the RF part and the analog-to-digital conversion, the digital signal processing starts with a training phase to determine symbol timing and frequency offset. An FFT is used to demodulate all subcarriers. The output of the FFT contains N_s QAM values, which are mapped onto binary values and decoded to produce binary output data. To successfully map the QAM values onto binary values, first the reference phases and amplitudes of all subcarriers have to be acquired. Alternatively, differential techniques can be applied. All of these receiver functions are highlighted in the following chapters.

2.7 IMPLEMENTATION COMPLEXITY OF OFDM VERSUS SINGLE-CARRIER MODULATION

One of the main reasons to use OFDM is its ability to deal with large delay spreads with a reasonable implementation complexity. In a single-carrier system, the implementation complexity is dominated by equalization, which is necessary when the delay spread is larger than about 10% of the symbol duration. OFDM does not require an equalizer. Instead, the complexity of an OFDM system is largely determined by the FFT, which is used to demodulate the various subcarriers. In the following example, it is demonstrated that the processing complexity of an OFDM modem can be significantly less than that of a single carrier modem, given the fact that both can deal with the same amount of delay spread. Notice that some papers mention the use of a

single-tap equalizer for OFDM, when they are referring to the phase correction in case of a coherent OFDM receiver. We will not use this somewhat confusing terminology, because the term equalization suggests that an attempt is made to invert the channel, while an OFDM receiver is doing the opposite; weak subcarriers are not being extra amplified to equalize the channel, but rather, they get a low weight in the decoding. In this way, OFDM avoids the problem of noise enhancement that is present in linear equalization techniques.

Figure 2.15 shows the block diagram of a decision feedback equalizer with symbol-spaced taps. Such an equalizer can be used to combat delay spread in single-carrier systems using quadrature amplitude modulation (QAM) or some form of constant amplitude modulation such as Gaussian Minimum Shift Keying (GMSK) or offset-QPSK. From references [5, 6], it can be learned that at least 8 feedforward and 8 feedback taps are required to handle a delay spread of 100 ns for a GMSK modem at a data rate of 24 Mbps. We can use this information to compare a single carrier system with the 24 Mbps mode of the new IEEE 802.11 OFDM standard, which can handle delay spreads up to 250 ns using a 64-point FFT. In order to increase the delay spread tolerance of a GMSK modem to the same level, the equalizer length has to be increased by a factor of 250/100, so it has 20 feedforward and feedback taps. Fortunately, for GMSK only the real outputs of the complex multiplications are used, so each multiplier has to perform two real multiplications per sample. Hence, the number of real multiplications per second becomes $2 \cdot 20 \cdot 24 \cdot 10^6 = 960 \cdot 10^6$. Notice we only count the feedforward taps here, since the feedback taps consist of trivial rotations, while the feedforward taps consist of full multiplications. For the OFDM system, a 64-point FFT has to be processed by every OFDM symbol duration which is 4 μ s, see chapter 10. With a radix-4 algorithm, this requires 96 complex multiplications [1], which gives a processing load of $96 \cdot 10^6$ real multiplications per second. So, in terms of multiplications per second, the equalizer of the single carrier system is 10 times more complex than the FFT of the OFDM system! This complexity difference grows with the bit rate, or rather with the bandwidth – delay spread product, which is a measure of the relative amount of inter-symbol interference. For instance, if we want to double the bit rate in the previous example by doubling the bandwidth, but the delay spread tolerance has to stay the same, then the number of subcarriers and the guard time has to be doubled for OFDM, while for the single-carrier system, both the number of equalizer taps and the sampling rate are doubled. The latter means that the number of multiplications per second is quadrupled, so the equalizer complexity grows quadratically with the bandwidth – delay spread product. For OFDM, a double-sized FFT has to be calculated in the same amount of time in order to double the rate. For the radix-2 algorithm, this means an increase in the number of multiplications of $N \log_2(2N) / (N/2) \log_2 N = 2(1 + 1/\log_2 N)$, where N is the FFT size for the half-rate system. The complexity of the FFT grows only slightly faster than linear with the bandwidth – delay spread product, which explains why OFDM is more attractive than a single carrier system with equalization for relatively large bandwidth – delay spread products (values around 1 or larger). It should be noted that the complexity difference between the FFT and the equalizer is less if the equalization is done in the frequency domain, as

described in [7]. In this case, equalization is twice as complex, because both an FFT and an IFFT have to be performed to do frequency domain equalization on a signal block.

Another complexity advantage of OFDM is the fact that the FFT does not really require full multiplications, but rather phase rotations, which can be efficiently implemented by the CORDIC algorithm [8]. Because phase rotations do not change the amplitude, they do not increase the dynamic range of the signals, which simplifies the fixed point design.

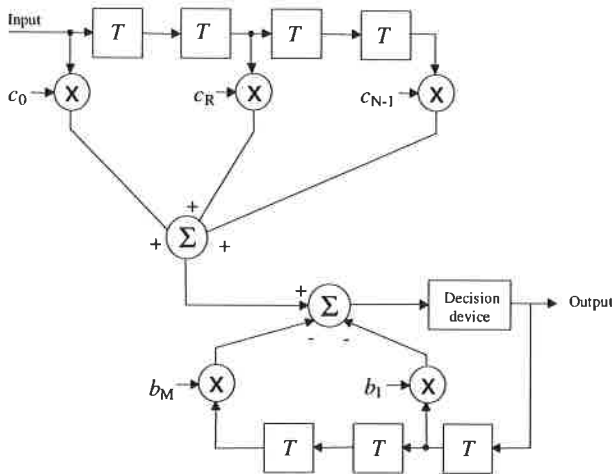


Figure 2.15 Decision-feedback equalizer.

Except for the difference in complexity, OFDM has another advantage over single carrier systems with equalizers. For the latter systems, the performance degrades abruptly if the delay spread exceeds the value for which the equalizer is designed. Because of error propagation, the raw bit error probability increases so quickly that introducing lower rate coding or a lower constellation size does not significantly improve the delay spread robustness. For OFDM, however, there are no such nonlinear effects as error propagation, and coding and lower constellation sizes can be employed to provide fallback rates that are significantly more robust against delay spread. This is an important consideration, as it enhances the coverage area and avoids the situation that users in bad spots cannot get any connection at all.

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