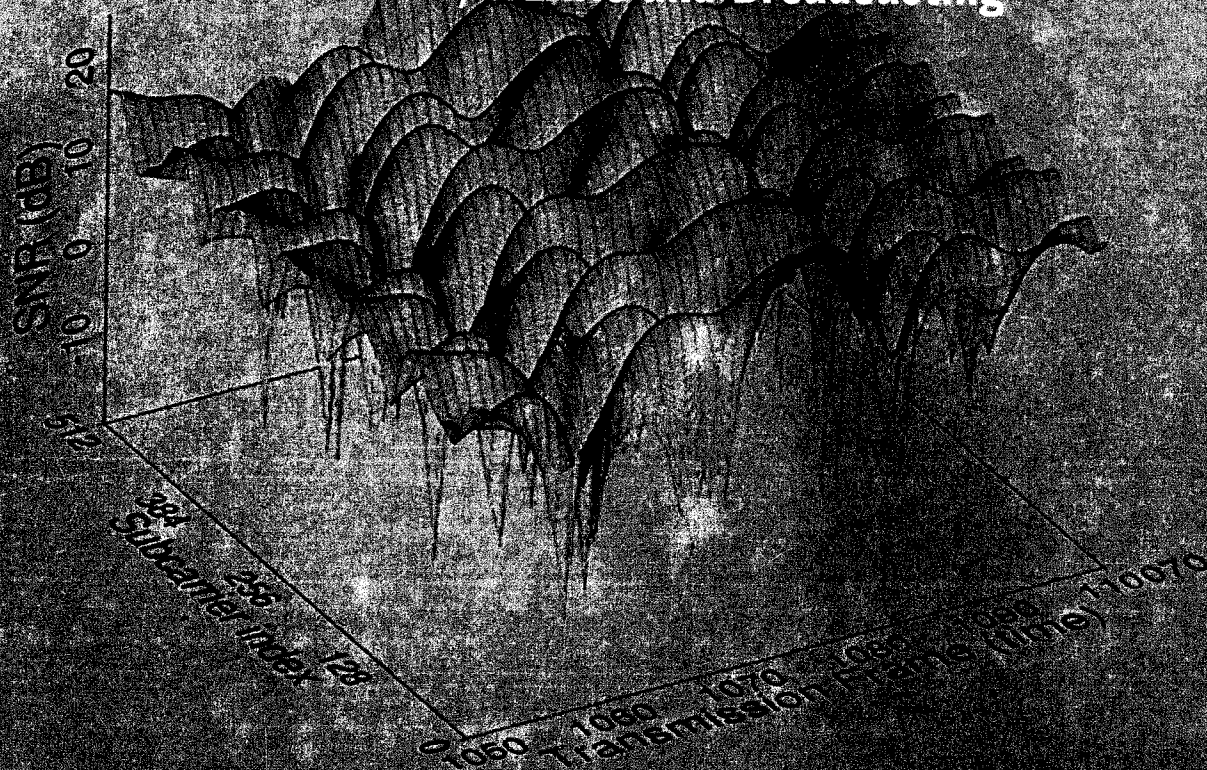


# Single- and Multi-carrier Quadrature Amplitude Modulation

Principles and Applications for Personal  
Communications, WLANs and Broadcasting



L. Hanzo • W. Webb • T. Keller

Hanzo  
Webb  
Keller

**Single- and Multi-carrier  
Quadrature Amplitude Modulation**



WILEY

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We dedicate this book to

**Rita, Alison and Katie**

as well as to the numerous contributors of the field,  
many of whom are listed in the acknowledgements and author index sections



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Communications, WLANs and Broadcasting

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# Preface

Since its discovery in the early 1960s, quadrature amplitude modulation (QAM) has continued to gain interest and practical application. Particularly in recent years many new ideas and techniques have been proposed, allowing its employment over fading mobile channels. This book attempts to provide an overview of most major QAM techniques, commencing with simple QAM schemes for the uninitiated, while endeavouring to pave the way towards complex, rapidly evolving areas, such as trellis-coded pilot-symbol and transparent-tone-in-band assisted schemes, or arrangements for wide-band mobile channels. The second half of the book is targetted at the more advanced reader, providing a research-oriented outlook using a variety of novel QAM-based single- and multi-carrier arrangements.

The book is structured in five parts. Part I - constituted by Chapters 1-4 - is a rudimentary introduction for those requiring a background in the field of modulation and radio wave propagation. Part II is comprised of Chapters 5-9 and concentrates mainly on classic QAM transmission issues relevant to Gaussian channels. Readers familiar with the fundamentals of QAM and the characteristics of propagation channels, as well as with basic pulse shaping techniques may decide to skip Chapters 1-5. Commencing with Chapter 6, each chapter describes individual aspects of QAM. Readers wishing to familiarize themselves with a particular subsystem, including clock and carrier recovery, equalisation, trellis coded modulation, standardised telephone-line modem features, etc. can turn directly to the relevant chapters, whereas those who desire a more complete treatment might like to read all the remaining chapters.

Parts III-V, including Chapters 10-24, are concerned with QAM-based transmissions over mobile radio channels. These chapters provide a research-based perspective and are dedicated to the more advanced reader. Specifically, Chapter 10 concentrates mainly on coherent QAM schemes, including reference-aided transparent-tone-in-band and pilot-symbol assisted modulation arrangements. In contrast, Chapter 11 focuses on low-complexity differentially encoded QAM schemes and on their performance with and without forward error correction coding and trellis coded modulation. Chapter 12 details various timing recovery schemes.

Part IV of the book commences with Chapter 13, which is concerned with variable rate QAM using one- to six-bits per symbol signal constellations. Chapter 14 is dedicated to high-rate wide-band transmissions and proposes a novel equaliser arrangement. Various QAM-related orthogonal signaling techniques are proposed in Chapter 15, while the spectral efficiency of QAM in cellular frequency re-use structures is detailed in Chapter 16. This is followed by Chapter 17, which concentrates on the employment of QAM in a source-matched speech communications system, in-

cluding various speech codecs, error correction codecs, a voice activity detector and packet reservation multiple access, providing performance figures in contrast to one and two bits per symbol bench-mark schemes.

Part V first appeared in this new edition of the book, concentrating on multi-carrier modulation. Specifically, following a rudimentary introduction to Orthogonal Frequency Division Multiplexing (OFDM) in Chapter 18, Chapters 19-23 detail a range of implementational and performance aspects of OFDM over both Gaussian and wideband fading channels. Lastly, Chapter 24 concentrates on the performance aspects of various standard-compliant and enhanced OFDM-based Digital Video Broadcasting (DVB) systems designed for transmission to mobile receivers.

To the original text of the first edition dealing with many of the fundamentals of single-carrier QAM and QAM-based systems we have added six new chapters dealing with the complexities of the exciting subject of multi-carrier modulation, which has found wide-ranging applications in a past decade, ranging from Wireless Local Area Network (WLAN) to broadcast systems. Whilst the book aims to portray a rapidly evolving area, where research results are promptly translated into products, it is our hope that you will find this second edition comprehensive, technically challenging and above all, enjoyable.

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*William Webb*

*Thomas Keller*

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Much of the results in Chapters 19-22 are based on our work conducted as a sub-contractor of Motorola ECID, Swindon, UK; as part of our involvement in a collaborative Pan-European Wireless Asynchronous Transfer Mode (WATM) project known as Median, which was generously supported by the European Commission (EC), Brussels, Belgium. We would like to acknowledge all our valued friends and colleagues - too numerous to mention individually - who at some stage were associated with the Median consortium and with whom we have enjoyed a stimulating collaboration under the stirring management of IMST, Germany. Our gratitude is due to Andy Wilton and to Paul Crichton of Motorola, who have whole-heartedly sponsored our research. Further thanks are also due to Dr. Joao Da Silva, Bartolome Aroyo, Bernard Barani, Dr. Jorge Pereira, Demosthenes Ikonomou and to the other equally supportive members of the EC's programme management team in Brussels for their enthusiastic support. Furthermore, we enjoyed the valuable support of EPSRC, Swindon UK, and the Mobile VCE, for which we are equally grateful.

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# Introduction and Background

This book is concerned with the issues of transmitting digital signals via multilevel modulation. We will be concerned with digital signals originating from a range of sources such as from speech or video encoding, or data from computers. A typical digital information transmission system is shown in Figure 1.1. The source encoder may be used to remove some of the redundancy which occurs in many sources such as speech, typically reducing the transmission rate of the source. The forward error correction (FEC) block then paradoxically inserts redundancy, but in a carefully controlled manner. This redundancy is in the form of parity check bits, and allows the FEC decoder to remove transmission errors caused by channel impairments, but at the cost of an increase in transmission rate. The interleaver systematically rearranges the bits to be transmitted, which has the effect of dispersing a short burst of errors at the receiver, allowing the FEC to work more effectively. Lastly, the modulator generates bandlimited waveforms which can be transmitted over the bandwidth-limited channel to the receiver, where the reverse functions are performed. Whilst we will discuss all aspects of Figure 1.1, it is the generation of waveforms in the modulator in a manner which reduces errors and increases the transmission rate within a given bandwidth, and the subsequent decoding in the demodulator, which will be our main concern in this book.

It is assumed that the majority of readers will be familiar with binary modulation schemes such as binary phase shift keying (BPSK), frequency shift keying (FSK), etc. Those readers who possess this knowledge might like to jump to Section 1.2. For those who are not familiar with modulation schemes we give a short non-mathematical explanation of modulation and constellation diagrams before detailing the history of QAM.

## 1.1 Modulation Methods

Suppose the data we wish to transmit is digitally encoded speech having a bit rate of 16 kbit/s, and after FEC coding the data rate becomes 32 kb/s. If the radio

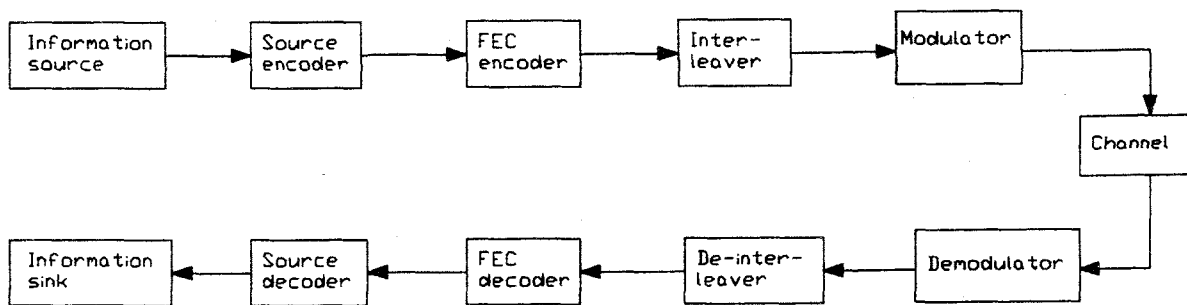


Figure 1.1: Typical digital information transmission system

channel to be used is centred around 1 GHz, then in order to transmit the 32 kb/s we must arrange for some feature of a 1 GHz carrier to change at the data rate of 32 kb/s. If the phase of the carrier is switched at the rate of 32 kb/s, being at 0 deg and 180 deg for bits having logical 0 or logical 1, respectively, then there are  $1 \times 10^9 / 32 \times 10^3 = 31,250$  radio frequency (RF) oscillations per bit transmitted. Figure 1.2(a) and (b) show the waveforms at the output of the modulator when the data is a logical 0 and a logical 1, respectively. On the left is the phasor diagram for a logical 0 and a logical 1 where the logical 0 is represented by a phasor along the positive  $x$ -axis, and the logical 1 by a phasor along the negative  $x$ -axis. This negative phasor represents a phase shift of the carrier by 180 deg. Figure 1.2(c) shows the modulator output for a sequence of data bits. Note that no filtering is used in this introductory example. Here the waveform can be seen to change abruptly at the boundary between some of the data symbols. We will see later that such abrupt changes can be problematic since they theoretically require an infinite bandwidth, and ways are sought to avoid them. Figure 1.2(d) is called a constellation diagram of phasor points, and as we are transmitting binary data there are only two points. As these two points are at equal distance from the origin we would expect them to represent equal magnitude carriers, and that the magnitude is indeed constant can be seen in Figure 1.2(c). The bandwidth of the modulated signal in this example will be in excess of the signalling rate of 32 kb/s due to the sudden transition between phase states. Later we will consider the bandwidth of modulated signals in-depth. Suffice to say here that if we decreased the signalling rate to 16 kb/s the bandwidth of the modulated signal will decrease. If the data rate is 32 kb/s, and the signalling rate becomes 16 kb/s, then every symbol transmitted must carry two bits of information. This means that we must have four points on the constellation, and clearly this can be done in many ways. Figure 1.3 shows some four-point constellations. The two bits of information associated with every constellation point are marked on the figure. In Figure 1.3(a) and (b) so-called quadrature modulation has been used as the points can only be uniquely described using two orthogonal coordinate axes, each passing through the origin. The orthogonal coordinate axes have a phase rotation of 90 deg with respect to each other, and hence they have a so-called quadrature relationship. The pair of coordinate axes can be associated with a pair of quadrature carriers, normally a sine and a cosine waveform, which can be independently modulated and then transmitted within the same frequency band. Due to their orthogonality they can be separated by the receiver. This implies that whatever symbol is chosen on one



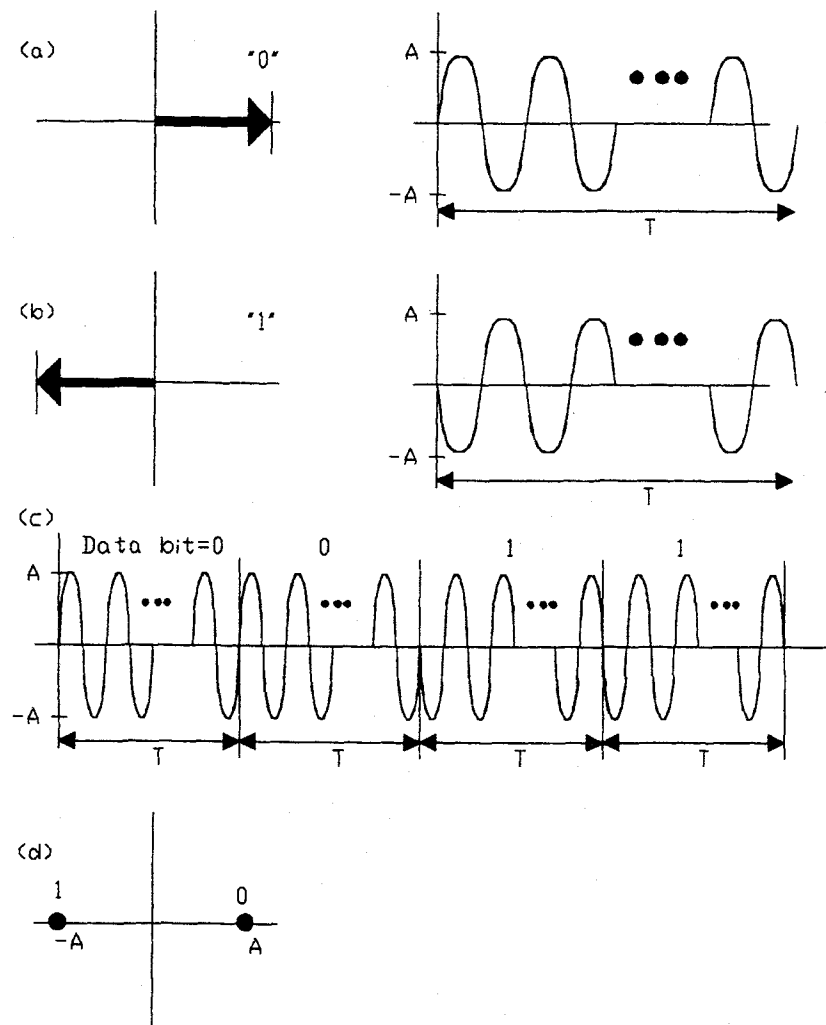


Figure 1.2: Carrier waveform for binary input data

axis (say the  $x$ -axis in the case of Figure 1.3(a)) it will have no effect on the data demodulated on the  $y$ -axis. Data can therefore be independently transmitted via these two quadrature or orthogonal channels without any increase in error rate, although some increase may result in practice and this is considered in later chapters. We have used I and Q to signify the in-phase and quadrature components, respectively, where in-phase normally represents the  $x$ -axis and quadrature the  $y$ -axis. Figure 1.3(c) and (d) show constellations where the four points are only on one line. These are not quadrature constellations but actually represent multilevel amplitude and phase modulation where both the carrier amplitude and phase can take two discrete values.

For the constellations in Figure 1.3(a) and (b) we have a constant amplitude signal, but the carrier phase values at the beginning of each symbol period in Figure 1.3(b) would be either 45 deg, 135 deg, 225 deg or 315 deg. There are two magnitude values and two phase values for the constellations in Figure 1.3(c) and (d).

In order to reduce the bandwidth of the modulated signal whilst maintaining the same information transmission rate we can further decrease the symbol rate by adding

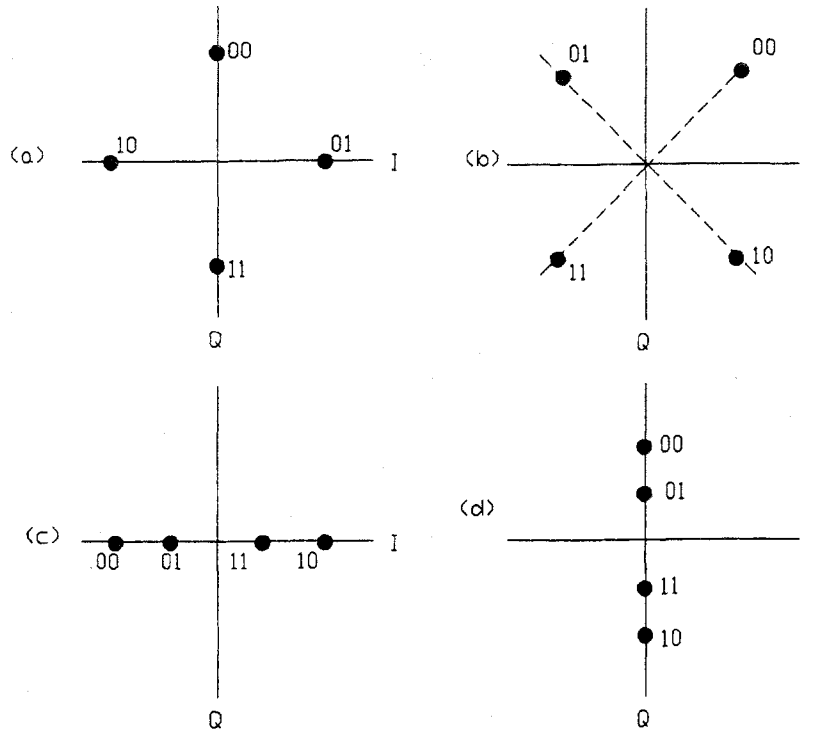


Figure 1.3: Examples of four-level constellations

more points in the constellation. Such a reduction in the bandwidth requirement will allow us to transmit more information in the spectrum we have been allocated. Such capability is normally considered advantageous. If we combine the constellations of Figure 1.3(c) and (d) we obtain the square QAM constellation having four bits per constellation point as displayed in Figure 1.4. We will spend much time in dealing with this constellation in this book. In general, grouping  $n$  bits into one signalling symbol yields  $2^n$  constellation points, which are often referred to as phasors, or complex vectors. The phasors associated with these points may have different amplitude and/or phase values, and this type of modulation is therefore referred to as multi-level modulation, where the number of levels is equal to the number of constellation points. After transmission through the channel the receiver must identify the phasor transmitted, and in doing so can determine the constellation point and hence the bits associated with this point. In this way the data is recovered. There are many problems with attempting to recover data transmitted over both fixed channels such as telephone lines and radio channels and many of these problems are given a whole chapter in this book. These problems are generally exacerbated by changing from binary to multilevel modulation, and this is why binary modulation is often preferred, despite its lower capacity. In order to introduce these problems, and to provide a historical perspective to quadrature amplitude modulation (QAM), a brief history of the development of QAM is presented.

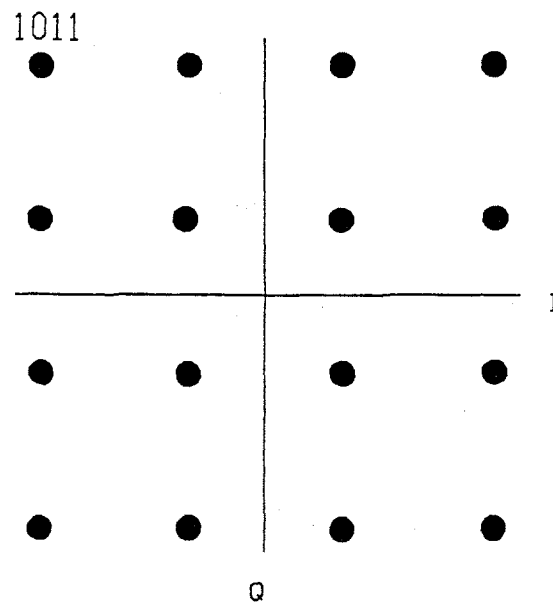


Figure 1.4: Square QAM constellation

## 1.2 History of Quadrature Amplitude Modulation

### 1.2.1 Determining the optimum constellation

Towards the end of the 1950s there was a considerable amount of interest in digital phase modulation transmission schemes [13] as an alternative to digital amplitude modulation. Digital phase modulation schemes are those whereby the amplitude of the transmitted carrier is held constant but the phase changed in response to the modulating signal. Such schemes have constellation diagrams of the form shown in Figure 1.3(a). It was a natural extension of this trend to consider the simultaneous use of both amplitude and phase modulation. The first paper to suggest this idea was by C.R. Cahn in 1960, who described a combined phase and amplitude modulation system [14]. He simply extended phase modulation to the multilevel case by allowing there to be more than one transmitted amplitude at any allowed phase. This had the effect of duplicating the original phase modulation or phase shift keying (PSK) constellation which essentially formed a circle. Such duplication led to a number of concentric circles depending on the number of amplitude levels selected. Each circle had the same number of phase points on each of its rings. Only Gaussian channels characteristic of telephone lines impaired by thermal noise were considered. Using a series of approximations and a wholly theoretical approach, he came to the conclusion that these amplitude and phase modulation (AM-PM) systems allowed an increased throughput compared to phase modulation systems when 16 or more states were used and suggested that such a system was practical to construct.

### 1.2.1.1 Coherent and non-coherent reception

The fundamental problem with PSK is that of determining the phase of the transmitted signal and hence decoding the transmitted information. This problem is also known as carrier recovery as an attempt is made to recover the phase of the carrier. When a phase point at, say, 90 deg is selected to reflect the information being transmitted, the phase of the transmitted carrier is set to 90 deg. However, the phase of the carrier is often changed by the transmission channel with the result that the receiver measures a different phase. This means that unless the receiver knew what the phase change imposed by the channel was it would be unable to determine the encoded information.

This problem can be overcome in one of two ways. The first is to measure the phase change imposed by the channel by a variety of means. The receiver can then determine the transmitted phase. This is known as coherent detection. The second is to transmit differences in phase, rather than absolute phase. The receiver then merely compares the previous phase with the current phase and the phase change of the channel is removed. This assumes that any phase change within the channel is relatively slow. This differential system is known as non-coherent transmission. In his paper, Cahn considered both coherent and non-coherent transmission, although for coherent transmission he assumed a hypothetical and unrealisable perfect carrier recovery device. The process of carrier recovery is considered in Chapter 6, and the details of differential transmission are explained in Chapter 5.

### 1.2.1.2 Clock recovery

Alongside carrier recovery runs the problem of clock recovery. The recovered clock signal is used to ensure appropriate sampling of the received signal. In Figure 1.2(c) a carrier signal was shown which had vertical lines indicating each bit or symbol period. It was the phase at the start of this period which was indicative of the encoded information. Unfortunately, the receiver has no knowledge of when these periods occur although it might know their approximate duration. It is determining these *symbol periods* which is the task of clock recovery. So carrier recovery estimates the phase of the transmitted carrier and clock recovery estimates the instances at which the data changes from one symbol to another. Whilst the need for carrier recovery can be removed through differential or non-coherent detection, there is no way to remove the requirement for clock recovery.

Clock recovery schemes tend to seek certain periodicities in the received signal and use these to estimate the start of a symbol (actually they often attempt to select the centre of a symbol for reasons which will be explained in later chapters). Clock recovery is often a complex procedure, and poor clock recovery can substantially increase the bit error rate (BER). The issue of clock recovery is considered in Chapter 6. In his work, Cahn overcame the problem of clock recovery by assuming that he had some device capable of perfect clock recovery. Such devices do not exist, so Cahn acknowledged that the error rate experienced in practice would be worse than the value he had calculated, but as he was unable to compute the errors introduced by a practical clock recovery system, this was the only course open to him.

### 1.2.1.3 The Type I, II and III constellations

A few months later a paper was published by Hancock and Lucky [15] in which they expanded upon the work of Cahn. In this paper they realised that the performance of the circular type constellation could be improved by having more points on the outer ring than on the inner ring. The rationale for this was that errors were caused when noise introduced into the signal moved the received phasor from the transmitted constellation point to a different one. The further apart constellation points could be placed, the less likely this was to happen. In Cahn's constellation, points on the inner ring were closest together in distance terms and so most vulnerable to errors. They conceded that a system with unequal numbers of points on each amplitude ring would be more complicated to implement, particularly in the case of non-coherent detection. They called the constellation proposed by Cahn a Type I system, and theirs a Type II system. Again using a mathematical approach they derived results similar to Cahn's for Type I systems and a 3 dB improvement for the Type II over the Type I system.

The next major publication was some 18 months later, in 1962, by Campopiano and Glazer [16]. They developed on the work of the previous papers but also introduced a new constellation - the square QAM system, which they termed a Type III system. They described this system as "essentially the amplitude modulation and demodulation of two carriers that have the same frequency but are in quadrature with each other" - the first time that combined amplitude and phase modulation had been thought of as amplitude modulation on quadrature carriers, although the acronym QAM was not suggested. They realised that the problem with their Type III system was that it had to be used in a phase coherent mode, that is non-coherent detection was not possible and so carrier recovery was necessary. Again, a theoretical analysis was performed for Gaussian noise channels and the authors came to the conclusion that the Type III system offered a very small improvement in performance over the Type II system, but thought that the implementation of the Type III system would be considerably simpler than that of Types I and II. Examples of the different types of constellation are shown in Figure 1.5.

Three months later another paper was published by Hancock and Lucky [17] in which they were probably unaware of the work done by Campopiano and Glazer. They attempted to improve on their previous work on the Type II system by carrying out a theoretical analysis, supposedly leading to the optimal constellation for Gaussian channels. In this paper they decided that the optimum 16-level constellation had two amplitude rings with eight equispaced points on each ring but with the rings shifted by 22.5 deg from each other. This constellation is shown in Figure 1.6.

Again, they concluded that 16 was the minimum number of levels for AM-PM modulation and that a SNR of at least 11 dB was required for efficient operation with a low probability of bit error.

After this paper there was a gap of nine years before any further significant advances were published. This was probably due to the difficulties in implementing QAM systems with the technology available and also because the need for increased data throughput was not yet pressing. During this period the work discussed in the above papers was consolidated into a number of books, particularly that by Lucky, Salz and Weldon [18]. Here they clearly distinguished between quadrature amplitude

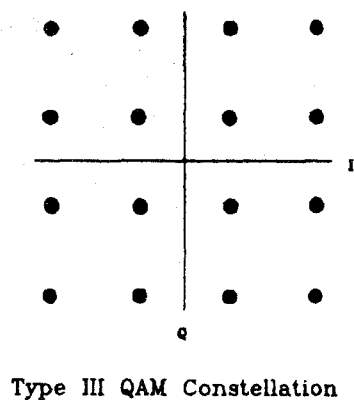
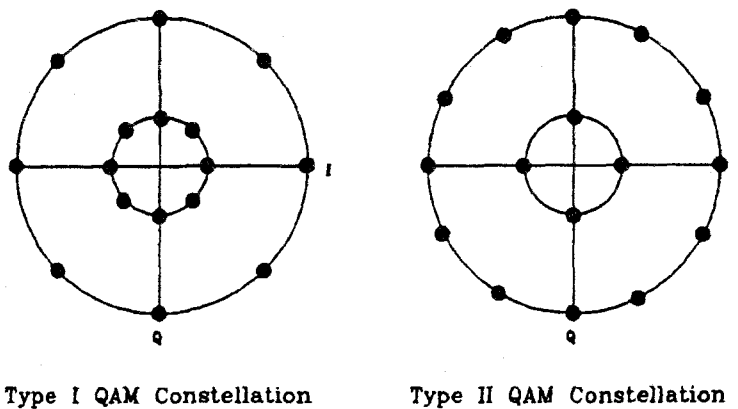


Figure 1.5: Examples of types I, II and III QAM constellations

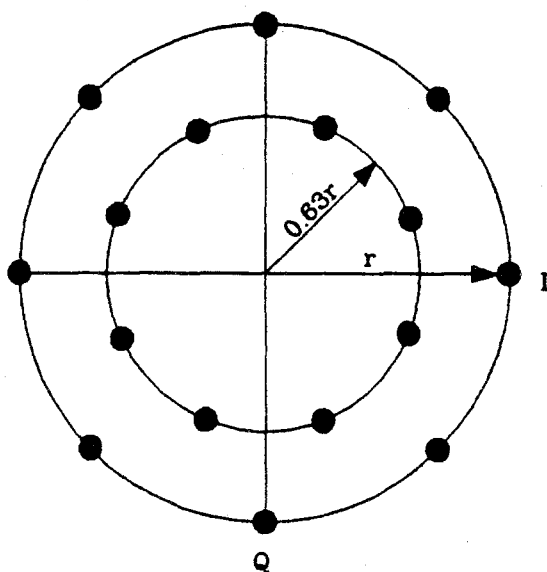


Figure 1.6: "Optimum" 16-level constellation according to Lucky [17] ©IRE, 1962

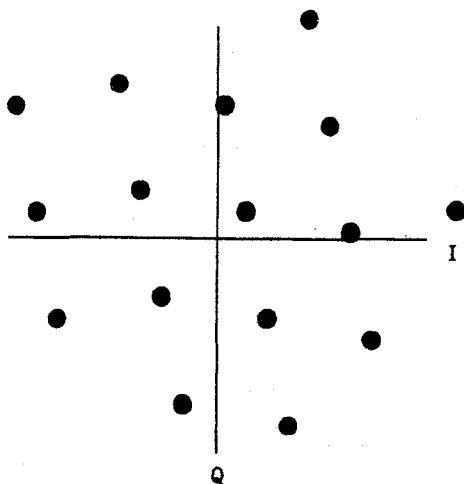


Figure 1.7: "Optimum" 16-level constellation according to Foschini [21] ©IEEE, 1974

modulation (QAM) schemes using square constellations and combined amplitude and phase modulation schemes using circular constellations. It was around this period that the acronym QAM started appearing in common usage along with AM-PM to describe the different constellations.

One of the earliest reports of the actual construction of a QAM system came from Salz, Sheenhan and Paris [19] of Bell Labs in 1971. They implemented circular constellations with 4 and 8 phase positions and 2 and 4 amplitude levels using coherent and non-coherent demodulation. Neither carrier nor clock recovery was attempted. Their results showed reasonable agreement with the theoretical results derived up to that time. This work was accompanied by that of Ho and Yeh [20] who improved the theory of circular AM-PM systems with algorithms that could be solved on digital computers which were by that time becoming increasingly available.

Interest in QAM remained relatively low, however, until 1974. In that year there was a number of significant papers published, considerably extending knowledge about quadrature amplitude modulation schemes. At this time, interest into optimum constellations was revived with two papers, one from Foschini, Gitlin and Weinstein [21] and the other from Thomas, Weidner and Durrani [22]. Foschini *et al.* attempted a theoretical derivation of the ideal constellation using a gradient calculation approach. They came to the conclusion that the ideal constellation was based around an equilateral triangle construction leading to the unusual 16-level constellation shown in Figure 1.7. This constellation has not found favour in practical applications, since the complexities involved in its employment outweigh the associated gains that were claimed for it.

Their conclusions were that this constellation, when limited in terms of power and operated over Gaussian channels, offered a performance improvement of 0.5 dB over square QAM constellations. Meanwhile Thomas *et al.*, working at COMSAT, empirically generated 29 constellations and compared their error probabilities.



### 1.2.2 Satellite links

In the paper by Thomas *et al.* [22] they also mentioned the first application of QAM, for use in satellite links. Satellite links have a particular problem in that satellites only have a limited power available to them. Efficient amplifiers are necessary to use this power carefully, and these had tended to employ a device known as a travelling wave tube (TWT). Such a device introduces significant distortion in the transmitted signal and Thomas *et al.* considered the effects of this distortion on the received waveform. They came to the interesting conclusion that the Type II constellation (in this case 3 points on the inner ring and 5 on the outer) was inferior to the Type I (4 points on inner and outer rings) due to the increased demand in peak-to-average power ratio of the Type II constellation. Their overall conclusion was that circular Type I constellations are superior in all cases. When they considered TWT distortion they discovered that AM-PM schemes were inferior to PSK schemes because of the need to significantly back off the amplifier to avoid severe amplitude distortion, and concluded that better linear amplifiers would be required before AM-PM techniques could be successfully used for satellite communications. They also considered the difficulties of implementing various carrier recovery techniques, advising that decision directed carrier recovery would be most appropriate, although few details were given as to how this was to be implemented. Decision directed carrier recovery is a process whereby the decoded signal is compared with the closest constellation point and the phase difference between them is used to estimate the error in the recovered carrier. This is discussed in more detail in Chapter 6.

Commensurate with the increasing interest as to possible applications for QAM were the two papers published in 1974 by Simon and Smith which concentrated on carrier recovery and detection techniques. In the first of these [23] they noted the interest in QAM that was then appearing for bandlimited systems, and addressed the problems of carrier recovery. They considered only the 16-level square constellation and noted that the generation of a highly accurate reconstructed carrier was essential for adequate performance. Their solution was to demodulate the signal, quantise it, and then establish the polarity of error from the nearest constellation point, and use it to update the voltage controlled oscillator (VCO) used in the carrier generation section. They provided a theoretical analysis and concluded that their carrier recovery technique worked well in the case of high signal-to-noise (SNR) ratio Gaussian noise, although they noted that gain control was required and would considerably complicate the implementation. They extended their work in Reference [24] where they considered offset QAM or O-QAM. In this modulation scheme the signal to one of the quadrature arms was delayed by half a symbol period in an attempt to prevent dramatic fluctuations of the signal envelope, which was particularly useful in satellite communications. They noted similar results for their decision directed carrier recovery scheme as when non-offset modulation was used.

#### 1.2.2.1 Odd-bit constellations

Despite all the work on optimum constellations, by 1975 interest had centred on the square QAM constellation. The shape of this was evident for even numbers of bits per symbol, but if there was a requirement for an odd number of bits per symbol to

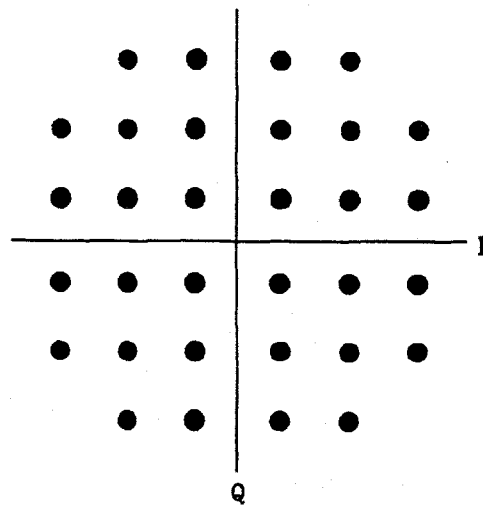


Figure 1.8: “Optimum” 32-level constellation according to Smith [25] ©IEEE, 1975

be transmitted, the ideal shape of the constellation was not obvious, with rectangular constellations having been tentatively suggested. Early in 1975 J.G. Smith, also working on satellite applications, published a paper addressing this problem [25]. He noted that for even numbers of bits per symbol “the square constellation was the only viable choice.” In this paper he showed that what he termed “symmetric” constellations offered about a 1 dB improvement over rectangular constellations and he considered both constellations to be of the same implementational complexity. Figure 1.8 shows an example of his symmetric constellation when there are 5 bits per symbol,

### 1.2.3 QAM modem implementation

About this time, the Japanese started to show interest in QAM schemes as they considered they might have application in both satellite and microwave radio links. In 1976 Miyauchi, Seki and Ishio published a paper devoted to implementation techniques [26]. They considered implementation by superimposing two 4-level PSK modulation techniques at different amplitudes to achieve a square QAM constellation and using a similar process in reverse at the demodulator, giving them the advantage of being able to use existing PSK modulator and demodulator circuits. This method of implementing QAM is discussed in Chapter 5. They implemented a prototype system without clock or carrier recovery and concluded that its performance was sufficiently good to merit further investigation. Further groundwork was covered in 1978 by W. Weber, again working on satellite applications, who considered differential (i.e. non-coherent) encoding systems for the various constellations still in favour [27]. His paper essentially added a theoretical basis to the differential techniques that had been in use at that point, although suggesting that non-coherent demodulation techniques deserved more attention.

In late 1979 evidence of the construction of QAM prototype systems worldwide started to emerge. A paper from the CNET laboratories in France by Dupuis *et al.* [28] considered a 140 Mbit/s modem for use in point-to-point links in the 10-11 GHz band. This prototype employed the square 16-level constellation and included carrier recovery, although no details were given. Theoretical calculations of impairments were presented followed by measurements made over a 58 km hop near Paris. Their conclusions were that QAM had a number of restrictions in its use, relating to its sensitivity to non-linearities, and that in the form they had implemented it, PSK offered improved performance. However, they suggested that with further work these problems might be overcome.

The Japanese simultaneously announced results from a prototype 200 Mbit/s 16-QAM system in a paper by Horikawa, Murase and Saito [29]. They used differential coding coupled with a new form of carrier recovery based on a decision feedback method (detailed in Chapter 6). Their modem was primarily designed for satellite applications and their experiment included the use of TWT amplifiers, but was only carried out back-to-back in the laboratory. Their conclusions were that their prototype had satisfactory performance and was an efficient way to increase bandwidth efficiency.

One of the last of the purely theoretical, as opposed to practical papers on QAM appeared in April 1980, marking the progression of QAM from a technical curiosity into a practical system, some twenty years after its introduction. This came from V. Prabhu of Bell Labs [30], further developing the theory to allow calculation of error probabilities in the presence of co-channel interference. Prabhu concluded that 16-QAM had a co-channel interference immunity superior to 16-PSK but inferior to 8-PSK.

### 1.2.3.1 Non-linear amplification

In 1982 there came a turning point in the use of QAM for satellite applications when Feher turned his attention to this problem. His first major publication in this field [31] introduced a new method of generation of QAM signals using highly non-linear amplifiers which he termed non-linear amplified QAM (NLA-QAM). Two separate amplifiers for the 16-QAM case were used, one operating with half the output power of the other. The higher power amplifier coded the two most significant bits of the 4-bit symbol only, and the lower power amplifier coded only the two least significant bits. The amplified coded signals were then summed at full output power to produce the QAM signal. Because both amplifiers were therefore able to use constant envelope modulation they could be run at full power with resulting high efficiency, although with increased complexity due to the need for two amplifiers and a hybrid combiner, compared to previous systems which used only a single amplifier. However, in satellite applications, complexity was relatively unimportant compared to power efficiency, and this NLA technique offered a very substantial 5 dB power gain, considerably increasing the potential of QAM in severely power limited applications.

### 1.2.3.2 Frequency selective fading and channel equalisers

This work was soon followed by a performance study of a NLA 64-state system [32] which extended the NLA scheme to 64 levels by using three amplifiers all operating at different power levels. Performance estimates were achieved using computer simulation techniques which included the effects of frequency selective fading.

Frequency selective fading is essentially caused when there are a number of propagation paths between the transmitter and receiver, e.g. by reflection off nearby mountains. When the time delay of the longest path compared to that of the shortest path becomes comparable to a symbol period, intersymbol interference (ISI) is said to result. As every time domain effect has an equivalent frequency domain effect, this can also be termed frequency selective fading. The two are related through the Fourier transform. This fading caused severe problems, leading Feher to conclude that adaptive equaliser techniques would be required for adequate wideband performance. Equalisers are devices which attempt to remove ISI. They do so by calculating the ISI introduced and then subtracting it from the received signals. They are often extremely complex devices and are considered in detail in Chapter 7.

### 1.2.3.3 Filtering

In a book published around this time [33], Feher also suggested the use of non-linear filtering (NLF) for QAM satellite communications. Since the I and Q components of the time domain QAM signal change their amplitude abruptly at the signalling intervals their transmission would require an infinite bandwidth. These abrupt changes are typically smoothed by a bandlimiting filter. The design and implementation of such filters is critical, particularly when the NLA-QAM signal is filtered at high power level. In order to alleviate these problems Feher developed the NLF technique along with Huang in 1979 which simplified filter design by simply fitting a quarter raised cosine segment between two initially abruptly changing symbols for both of the quadrature carriers. We have already hinted that filtering is required to prevent abrupt changes in the transmitted signal. This issue is considered in more rigour in Chapter 5. This allowed the generation of jitter-free bandlimited signals, which had previously been a problem, improving clock recovery techniques. Feher's work continued to increase the number of levels used, with a paper on 256-QAM in May 1985 [34] noting the problems that linear group delay distortion caused, and a paper in April 1986 on 512-QAM [35] which came to similar conclusions.

### 1.2.4 Advanced prototypes

Work was still continuing in France, Japan and also in New Zealand. CNET were continuing their attempt to overcome the problems they had found in their initial trials reported in 1979. A paper published in 1985 by M. Borgne [36] compared the performance of 16, 32, 64 and 128 level QAM schemes using computer simulation with particular interest in the impairments likely to occur over point-to-point radio links. Borgne concluded that non-linearity cancellers and adaptive equalisers would

concerned with clock and carrier recovery without going into detail as to how these operations were performed. Details were provided of a fractional decision feedback equaliser (DFE) which they considered suitable for point-to-point radio links. They concluded that carrier recovery and clock timing was critical and likely to cause major problems, which were somewhat ameliorated by their fractionally spaced system.

Although lagging somewhat behind Feher, the Japanese made up for this delay by publishing a very detailed paper describing the development of a 256-level QAM modem in August 1986. In this paper from Saito and Nakamura [38], the authors developed on the work announced in 1979 by Saito *et al.* [26] which was discussed earlier. In this new paper they detailed automatic gain control (which he termed automatic threshold control) and carrier recovery methods. The carrier recovery was a slight enhancement to the system announced in 1979 and the AGC system was based on decision directed methods. Details were given as to how false lock problems were avoided (see Chapter 6) and the back-to-back prototype experiments gave results which the authors considered showed the feasibility of the 256-QAM modem. Evidence of the ever increasing interest in QAM was that in the IEEE special issue on advances in digital communications by radio there was a substantial section devoted to high-level modulation techniques. In a paper by Rustako *et al.* [39] which considered point-to-point applications, the standard times-two carrier recovery method for binary modulation was expanded for QAM. The authors claimed the advantage of not requiring accurate data decisions or interacting with any equaliser. They acknowledged that their system had the disadvantage of slow reacquisition after fades and suggested that it would only be superior in certain situations. This form of carrier recovery is considered in some detail in Chapter 6.

Clearly, until the late 1980s developments were mainly targeted at telephone line and point-to-point radio applications, which led to the definition of the CCITT telephone circuit modem standards V.29 to V.33 based on various QAM constellations ranging from uncoded 16-QAM to trellis coded (TC) 128-QAM. Coded modulation is analysed in Chapter 8, while the CCITT standard modem schemes V.29 to V.33 for telephone lines will be considered in Chapter 9.

### 1.2.5 QAM for mobile radio

Another major development occurred in 1987 when Sundberg, Wong and Steele published a pair of papers [40, 41] considering QAM for voice transmission over Rayleigh fading channels, the first major paper considering QAM for mobile radio applications. In these papers, it was recognized that when a Gray code mapping scheme was used, some of the bits constituting a symbol had different error rates from other bits. Gray coding is a method of assigning bits to be transmitted to constellation points in an optimum manner and is discussed in Chapter 5. For the 16-level constellation two classes of bits occurred, for the 64-level three classes and so on. Efficient mapping schemes for pulse code modulated (PCM) speech coding were discussed where the most significant bits (MSBs) were mapped onto the class with the highest integrity. A number of other schemes including variable threshold systems and weighted systems were also discussed. Simulation and theoretical results were compared and found to be in reasonable agreement. They used no carrier recovery, clock recovery or AGC,

assuming these to be ideal, and came to the conclusion that channel coding and post-enhancement techniques would be required to achieve acceptable performance.

This work was continued, resulting in a publication in 1990 by Hanzo, Steele and Fortune [42], again considering QAM for mobile radio transmission, where again a theoretical argument was used to show that with a Gray encoded square constellation, the bits encoded onto a single symbol could be split into a number of subclasses, each subclass having a different average BER. The authors then showed that the difference in BER of these different subclasses could be reduced by constellation distortion at the cost of slightly increased total BER, but was best dealt with by using different error correction powers on the different 16-QAM subclasses. A 16 kbit/s sub-band speech coder was subjected to bit sensitivity analysis and the most sensitive bits identified were mapped onto the higher integrity 16-QAM subclasses, relegating the less sensitive speech bits to the lower integrity classes. Furthermore, different error correction coding powers were considered for each class of bits to optimise performance. Again ideal clock and carrier recovery were used, although this time the problem of automatic gain control (AGC) was addressed. It was suggested that as bandwidth became increasingly congested in mobile radio, microcells would be introduced supplying the required high SNRs with the lack of bandwidth being an incentive to use QAM.

In the meantime, CNET were still continuing their study of QAM for point-to-point applications, and Sari and Moridi published a paper [43] detailing an improved carrier recovery system using a novel combination of phase and frequency detectors which seemed promising. However, interest was now increasing in QAM for mobile radio usage and a paper was published in 1989 by J. Chuang of Bell Labs [44] considering NLF-QAM for mobile radio and concluding that NLF offered slight improvements over raised cosine filtering when there was mild intersymbol interference (ISI).

A technique, known as the transparent tone in band method (TTIB) was proposed by McGeehan and Bateman [9] from Bristol University, UK, which facilitated coherent detection of the square QAM scheme over fading channels and was shown to give good performance but at the cost of an increase in spectral occupancy. This important technique is discussed in depth in Chapter 11. At an IEE colloquium on multilevel modulation techniques in March 1990 a number of papers were presented considering QAM for mobile radio and point-to-point applications. Matthews [45] proposed the use of a pilot tone located in the centre of the frequency band for QAM transmissions over mobile channels.

Huish discussed the use of QAM over fixed links, which was becoming increasingly widespread [46]. Webb *et al.* presented two papers describing the problems of square QAM constellations when used for mobile radio transmissions and introduced the star QAM constellation with its inherent robustness in fading channels [47, 48].

There followed a number of publications by Webb *et al.* describing a number of techniques for enhancing QAM transmissions over mobile radio channels. All of these techniques are described in detail in Chapters 10 to 15 of this book. In December 1991 a paper appeared in the *IEE Proceedings* [49] which considered the effects of channel coding, trellis coding and block coding when applied to the star QAM constellation. This was followed by another paper in the *IEE Proceedings* [50] considering equaliser techniques for QAM transmissions over dispersive mobile radio channels. A review paper appearing in July 1992 [51] considered areas where QAM could be put to most

beneficial use within the mobile radio environment, and concluded that its advantages would be greatest in microcells. Further work on spectral efficiency, particularly of multilevel modulation schemes [52] concluded that variable level QAM modulation was substantially more efficient than all the other modulation schemes simulated. Variable level QAM was first discussed in a paper by Steele and Webb in 1991 [53].

Further QAM schemes for hostile fading channels characteristic of mobile telephony can be found in the following recent references [1, 1, 12, 54–86]. If Feher's previously mentioned NLA concept cannot be applied, then power-inefficient class A or AB linear amplification has to be used, which might become an impediment in lightweight, low-consumption handsets. However, the power consumption of the low-efficiency class A amplifier [63, 64] is less critical than that of the digital speech and channel codecs. In many applications 16-QAM, transmitting 4 bits per symbol reduces the signalling rate by a factor of 4 and hence mitigates channel dispersion, thereby removing the need for an equaliser, while the higher SNR demand can be compensated by diversity reception.

Significant contributions were made by Cavers, Stapleton *et al.* at Simon Fraser University, Burnaby, Canada in the field of pre- and post-distorter design. Out-of-band emissions due to class AB amplifier non-linearities and hence adjacent channel interferences can be reduced by some 15–20 dB using Stapleton's adaptive predistorter [1, 1, 65] and a class AB amplifier with 6 dB back-off, by adjusting the predistorter's adaptive coefficients using the complex convolution of the predistorter's input signal and the amplifier's output signal. Further aspects of linearised power amplifier design are considered in references [66] and [67].

A further important research trend is hallmarked by Cavers' work targeted at pilot symbol assisted modulation (PSAM) [12], where known pilot symbols are inserted in the information stream in order to allow the derivation of channel measurement information. The recovered received symbols are then used to linearly predict the channel's attenuation and phase. This arrangement will be considered in Chapter 11. A range of advanced QAM modems have also been proposed by Japanese researchers doing cutting-edge research in the field, including Sampei, Sunaga [68, 69], Adachi [70] and Sasaoka [62].

Since QAM research has reached a mature stage, a number of mobile speech, audio and video transmission schemes have been proposed [85–95]. These system design examples demonstrated that substantial system performance benefits accrue, when the entire system is jointly optimised, rather than just a conglomerate of independent system components. Some of these system aspects will be considered in Chapter 17 in the context of voice communications. A range of digital video broadcasting (DVB) schemes will be the topic of Chapter 24.

## 1.3 Orthogonal Frequency Division Multiplexing Based QAM

### 1.3.1 History of orthogonal frequency division multiplexing

The first QAM-related so-called orthogonal frequency division multiplexing (OFDM) scheme was proposed by Chang in 1966 [71] for dispersive fading channels, which has also undergone a dramatic evolution due to the efforts of Weinstein, Peled, Ruiz, Hirosaki, Kolb, Cimini, Schüssler, Preuss, Rückriem, Kalet *et al.* [71–84]. OFDM was standardised as the European digital audio broadcast (DAB) as well as digital video broadcast (DVB) scheme. It constituted also a credible proposal for the recent third-generation mobile radio standard competition in Europe. It was recently selected as the high performance local area network (HIPERLAN) transmission technique.

The system's operational principle is that the original bandwidth is divided in a high number of narrow sub-bands, in which the mobile channel can be considered non-dispersive. Hence no channel equaliser is required and instead of implementing a bank of sub-channel modems they can be conveniently implemented by the help of a single fast Fourier Transformer (FFT). This scheme will be the topic of Chapters 18–23.

These OFDM systems - often also termed as frequency division multiplexing (FDM) or multi-tone systems - have been employed in military applications since the 1960s, for example by Bello [96], Zimmerman [72], Powers and Zimmerman [97], Chang and Gibby [98] and others. Saltzberg [99] studied a multi-carrier system employing orthogonal time-staggered quadrature amplitude modulation (O-QAM) of the carriers.

The employment of the discrete Fourier transform (DFT) to replace the banks of sinusoidal generators and the demodulators was suggested by Weinstein and Ebert [73] in 1971, which significantly reduces the implementation complexity of OFDM modems. In 1980, Hirosaki [84] suggested an equalisation algorithm in order to suppress both intersymbol and intersubcarrier interference caused by the channel impulse response or timing and frequency errors. Simplified OFDM modem implementations were studied by Peled [77] in 1980, while Hirosaki [78] introduced the DFT based implementation of Saltzberg's O-QAM OFDM system. From Erlangen University, Kolb [79], Schüßler [80], Preuss [81] and Rückriem [82] conducted further research into the application of OFDM. Cimini [74] and Kalet [83] published analytical and early seminal experimental results on the performance of OFDM modems in mobile communications channels.

More recent advances in OFDM transmission were presented in the impressive state-of-the-art collection of works edited by Fazel and Fettweis [100], including the research by Fettweis *et al.* at Dresden University, Rohling *et al.* at Braunschweig University, Vandendorp at Loeven University, Huber *et al.* at Erlangen University, Lindner *et al.* at Ulm University, Kammeyer *et al.* at Brehmen University and Meyr *et al.* [101,102] at Aachen University, but the individual contributions are too numerous to mention.

While OFDM transmission over mobile communications channels can alleviate the problem of multipath propagation, recent research efforts have focused on solving a set of inherent difficulties regarding OFDM, namely the peak-to-mean power ratio,



time and frequency synchronisation, and on mitigating the effects of the frequency selective fading channel. These issues are addressed below in slightly more depth, while a treatment is given in Chapters 18-23.

### 1.3.2 Peak-to-mean power ratio

It is plausible that the OFDM signal - which is the superposition of a high number of modulated sub-channel signals - may exhibit a high instantaneous signal peak with respect to the average signal level. Furthermore, large signal amplitude swings are encountered, when the time domain signal traverses from a low instantaneous power waveform to a high power waveform, which may result in a high out-of-band (OOB) harmonic distortion power, unless the transmitter's power amplifier exhibits an extremely high linearity across the entire signal level range (Section 4.5.1). This then potentially contaminates the adjacent channels with adjacent channel interference. Practical amplifiers exhibit a finite amplitude range, in which they can be considered almost linear. In order to prevent severe clipping of the high OFDM signal peaks - which is the main source of OOB emissions - the power amplifier must not be driven into saturation and hence they are typically operated with a certain so-called back-off, creating a certain "head room" for the signal peaks, which reduces the risk of amplifier saturation and OOB emission. Two different families of solutions have been suggested in the literature, in order to mitigate these problems, either reducing the peak-to-mean power ratio, or improving the amplification stage of the transmitter.

More explicitly, Shepherd [103], Jones [104], and Wulich [105] suggested different coding techniques which aim to minimise the peak power of the OFDM signal by employing different data encoding schemes before modulation, with the philosophy of choosing block codes whose legitimate code words exhibit low so-called Crest factors or peak-to-mean power envelope fluctuation. Müller [106], Pauli [107], May [108] and Wulich [109] suggested different algorithms for post-processing the time domain OFDM signal prior to amplification, while Schmidt and Kammeyer [110] employed adaptive subcarrier allocation in order to reduce the crest factor. Dinis and Gusmão [111-113] researched the use of two-branch amplifiers, while the clustered OFDM technique introduced by Daneshrad, Cimini and Carloni [114] operates with a set of parallel partial FFT processors with associated transmitting chains. OFDM systems with increased robustness to non-linear distortion have been proposed by Okada, Nishijima and Komaki [115] as well as by Dinis and Gusmão [116].

### 1.3.3 Synchronisation

Time and frequency synchronisation between the transmitter and receiver are of crucial importance as regards to the performance of an OFDM link [117, 118]. A wide variety of techniques have been proposed for estimating and correcting both timing and carrier frequency offsets at the OFDM receiver. Rough timing and frequency acquisition algorithms relying on known pilot symbols or pilot tones embedded into the OFDM symbols have been suggested by Claßen [101], Warner [119], Sari [120], Moose [121], as well as Brüninghaus and Rohling [122]. Fine frequency and timing tracking algorithms exploiting the OFDM signal's cyclic extension were published by

Moose [121], Daffara [123] and Sandell [124].

#### 1.3.4 OFDM/CDMA

Combining OFDM transmissions with code division multiple access (CDMA) allows us to exploit the wideband channel's inherent frequency diversity by spreading each symbol across multiple subcarriers. This technique has been pioneered by Yee, Lin-nartz and Fettweis [125], by Chouly, Brajal and Jourdan [126], as well as by Fettweis, Bahai and Anvari [127]. Fazel and Papke [128] investigated convolutional coding in conjunction with OFDM/CDMA. Prasad and Hara [129] compared various methods of combining the two techniques, identifying three different structures, namely multi-carrier CDMA (MC-CDMA), multi-carrier direct sequence CDMA (MC-DS-CDMA) and multi-tone CDMA (MT-CDMA). Like non-spread OFDM transmission, OFDM/CDMA methods suffer from high peak-to-mean power ratios, which are dependent on the frequency domain spreading scheme, as investigated by Choi, Kuan and Hanzo [130].

#### 1.3.5 Adaptive antennas

Combining adaptive antenna techniques with OFDM transmissions was shown to be advantageous in suppressing co-channel interference in cellular communications systems. Li, Cimini and Sollenberger [131–133], Kim, Choi and Cho [134], Lin, Cimini and Chuang [135] as well as Münster *et al.* [136] have investigated algorithms for multi-user channel estimation and interference suppression.

#### 1.3.6 OFDM applications

Due to their implementational complexity, OFDM applications have been scarce until quite recently. Recently, however, OFDM has been adopted as the new European digital audio broadcasting (DAB) standard [75, 76, 137–139] as well as for the terrestrial digital video broadcasting (DVB) system [120, 140].

For fixed-wire applications, OFDM is employed in the asynchronous digital subscriber line (ADSL) and high-bit-rate digital subscriber line (HDSL) systems [141–144] and it has also been suggested for power line communications systems [145, 146] due to its resilience to time dispersive channels and narrow band interferers.

More recently, OFDM applications were studied within the European 4th Framework Advanced Communications Technologies and Services (ACTS) programme [147]. The MEDIAN project investigated a 155 Mbps wireless asynchronous transfer mode (WATM) network [148–151], while the Magic WAND group [152, 153] developed a wireless local area network (LAN). Hallmann and Rohling [154] presented a range of different OFDM systems that were applicable to the European Telecommunications Standardisation Institute's (ETSI) recent personal communications oriented air interface concept [155].

## 1.4 Summary

This concludes our introduction to QAM and the review of publications concerning QAM, spanning a period of over thirty years between the first theoretical study for Gaussian channels through to implementation within microcellular mobile radio environments. We now embark on a detailed investigation of the topics introduced in this chapter.

## 1.5 Outline of Topics

In Chapter 2 we consider the communications channels over which we wish to send our data. These channels are divided into Gaussian and mobile radio channels, and the characteristics of each are explained.

Chapter 3 forms an introduction to modems, considering the manner in which speech or other source waveforms are converted into a form suitable for transmission over a channel, and introducing some of the fundamentals of modems.

Chapter 4 provides a more detailed description of modems, specifically that of the modulator, considering QAM constellations, pulse shaping techniques, methods of generating and detecting QAM, as well as amplifier techniques to reduce the problems associated with non-linearities.

Chapter 5 provides details of decision theory and highlights the theoretical aspects of QAM transmission, showing how the BER can be mathematically computed for transmission over Gaussian channels.

Chapter 6 considers clock and carrier recovery schemes whilst Chapter 7 looks at equalisers. Chapter 8 considers trellis coded modulation and provides an introduction to Chapter 9 which deals with modems for Gaussian channels such as telephone lines.

Chapter 10 starts the section on mobile radio by providing a theoretical analysis of QAM transmission over Rayleigh fading mobile radio channels. Chapter 11 then considers some of the practicalities of QAM transmission over these links, including constellation distortions and hardware imperfections.

Chapter 12 details clock and carrier recovery schemes for mobile radio, Chapter 13 considers an advanced variable rate QAM scheme, while Chapter 14 proposes various equalisers for wideband transmissions. In Chapter 15 we consider a range of orthogonal transmission and pulse shaping techniques in the context of quadrature-quadrature amplitude modulation.

Chapter 16 considers the spectral efficiency gains that can be achieved when using QAM instead of conventional binary modulation techniques, whilst Chapter 17 provides details of a plethora of practical QAM speech communications systems and compares these with their equivalent binary systems.

In Chapter 18 we focus our attention on Fourier transforms in order to implement orthogonal frequency division multiplexing schemes. In Chapter 19 the BER performance of OFDM modems in AWGN channels is studied for a set of different modulation schemes in the subcarriers. The effects of amplitude limiting of the transmitter's output signal, caused by a simple clipping amplifier model, and of finite resolution D/A and A/D conversion on the system performance are investigated.

Oscillator phase noise is considered as a source of intersubcarrier interference and its effects on the system performance are demonstrated.

In Chapter 20 the effects of time-dispersive frequency-selective Rayleigh fading channels on OFDM transmissions are demonstrated. Channel estimation techniques are presented which support the employment of coherent detection in frequency selective channels. Additionally, differential detection is investigated, and the resulting system performance over the different channels is compared.

Chapter 21 focuses our attention on the time and frequency synchronisation requirements of OFDM transmissions and the effects of synchronisation errors are demonstrated. Two novel synchronisation algorithms for frame and OFDM symbol synchronisation are suggested and compared. The resulting system performance over fading wideband channels is examined.

In Chapter 22, based on the results of Chapter 20, the employment of adaptive modulation schemes is suggested for duplex point-to-point links over frequency-selective time-varying channels. Different bit allocation schemes are investigated and a simplified sub-band adaptivity OFDM scheme is suggested for alleviating the associated signalling constraints. A range of blind modulation scheme detection algorithms are also investigated and compared. The employment of long-block-length convolutional turbo codes is suggested for improving the system's throughput and the turbo coded adaptive OFDM modem's performance is compared using different sets of parameters. Then the effects of using pre-equalisation at the transmitter are examined, and a set of different pre-equalisation algorithms is introduced. A joint pre-equalisation and adaptive modulation algorithm is proposed and its BER and throughput performance is studied.

In Chapter 23 the adaptive OFDM transmission ideas of Chapter 22 are extended further, in order to include adaptive error correction coding, based on redundant residual number system (RRNS) and turbo BCH codes. A joint modulation and code rate adaptation scheme is presented.

Lastly, in Chapter 24 a variety of advanced QAM and OFDM assisted turbo-coded DVB schemes are proposed and analysed. Specifically, it compares the performance of schemes that use blind-equalised QAM modems. It is demonstrated that in comparison to the standard-based DVB systems the employment of turbo coding can provide an extra 5-6 dB channel SNR gain and this can be exploited for example to double the number of transmitted bits in a given bandwidth. This then ultimately allows us to improve for example the associated video quality that can be guaranteed in a given bandwidth at the cost of the associated additional implementational complexity.



# Chapter 22

## Adaptive Single- and Multi-user OFDM Techniques

### 22.1 Introduction

Steele and Webb [53] proposed adaptive modulation for exploiting the time-variant Shannonian channel capacity of fading narrowband channels, and further research was conducted at Osaka University by Sampei *et al.* [420], at the University of Stanford by Goldsmith *et al.* [421], by Pearce, Burr and Tozer at the University of York [422], Lau and McLeod at the University of Cambridge [423], and at Southampton University [424,425]. The associated principles can also be invoked in the context of parallel modems, as it has been demonstrated by Kalet [83], Czylik *et al.* [426] as well as by Chow, Cioffi and Bingham [427].

#### 22.1.1 Motivation

We have seen in Figure 20.12 that the bit error probability of different OFDM subcarriers transmitted in time dispersive channels depends on the frequency domain channel transfer function. The occurrence of bit errors is normally concentrated in a set of severely faded subcarriers, while in the rest of the OFDM spectrum often no bit errors are observed. If the subcarriers that will exhibit high bit error probabilities in the OFDM symbol to be transmitted can be identified and excluded from data transmission, the overall BER can be improved in exchange for a slight loss of system throughput. As the frequency domain fading deteriorates the SNR of certain subcarriers, but improves others' above the average SNR value, the potential loss of throughput due to the exclusion of faded subcarriers can be mitigated by employing higher-order modulation modes on the subcarriers exhibiting high SNR values.

In addition to excluding sets of faded subcarriers and varying the modulation modes employed, other parameters such as the coding rate of error correction coding schemes can be adapted at the transmitter according to the perceived channel transfer

function. This issue will be addressed in Chapter 23.

Adaptation of the transmission parameters is based on the transmitter's perception of the channel conditions in the forthcoming time slot. Clearly, this estimation of future channel parameters can only be obtained by extrapolation of previous channel estimations, which are acquired upon detecting each received OFDM symbol. The channel characteristics therefore have to be varying sufficiently slowly compared to the estimation interval.

Adapting the transmission technique to the channel conditions on a time slot by time slot basis for serial modems in narrowband fading channels has been shown to considerably improve the BER performance [428] for time division duplex (TDD) systems assuming duplex reciprocal channels. However, the Doppler fading rate of the narrowband channel has a strong effect on the achievable system performance: if the fading is rapid, then the prediction of the channel conditions for the next transmit time slot is inaccurate, and therefore the wrong set of transmission parameters may be chosen. If, however, the channel varies slowly, then the data throughput of the system is varying dramatically over time, and large data buffers are required at the transmitters in order to smoothen the bit rate fluctuation. For time critical applications, such as interactive speech transmission, the potential delays can become problematic. A given single-carrier adaptive system in narrowband channels will therefore operate efficiently only in a limited range of channel conditions.

Adaptive OFDM modem channels can ease the problem of slowly time-varying channels, since the variation of the signal quality can be exploited in the time domain and the frequency domain. The channel conditions still have to be monitored based on the received OFDM symbols, and relatively slowly varying channels have to be assumed, since we have seen in Section 20.2.2 that OFDM transmissions are not well suited to rapidly varying channel conditions.

### 22.1.2 Adaptive techniques

Adaptive modulation is only suitable for duplex communication between two stations, since the transmission parameters have to be adapted using some form of two-way transmission in order to allow channel measurements and signalling to take place.

Transmission parameter adaptation is a response of the transmitter to time-varying channel conditions. In order to efficiently react to the changes in channel quality, the following steps have to be taken:

- (1) *Channel quality estimation:* In order to appropriately select the transmission parameters to be employed for the next transmission, a reliable estimation of the channel transfer function during the next active transmit time slot is necessary.
- (2) *Choice of the appropriate parameters for the next transmission:* Based on the prediction of the channel conditions for the next time slot, the transmitter has to select the appropriate modulation modes for the subcarriers.
- (3) *Signalling or blind detection of the employed parameters:* The receiver has to be informed, as to which demodulator parameters to employ for the received packet. This information can either be conveyed within the OFDM symbol

itself, with the loss of effective data throughput, or the receiver can attempt to estimate the parameters employed by the remote transmitter by means of blind detection mechanisms.

### 22.1.2.1 Channel quality estimation

The transmitter requires an estimate of the expected channel conditions for the time when the next OFDM symbol is to be transmitted. Since this knowledge can only be gained by prediction from past channel quality estimations, the adaptive system can only operate efficiently in an environment exhibiting relatively slowly varying channel conditions.

The channel quality estimation can be acquired from a range of different sources. If the communication between the two stations is bidirectional and the channel can be considered reciprocal, then each station can estimate the channel quality on the basis of the received OFDM symbols, and adapt the parameters of the local transmitter to this estimation. We will refer to such a regime as *open-loop adaptation*, since there is no feedback between the receiver of a given OFDM symbol and the choice of the modulation parameters. A time division duplex (TDD) system in absence of interference is an example of such a system, and hence a TDD regime is assumed for generating the performance results below. Channel reciprocity issues were addressed for example in [429, 430].

If the channel is not reciprocal, as in a frequency division duplex (FDD) system, then the stations cannot determine the parameters for the next OFDM symbol's transmission from the received symbols. In this case, the receiver has to estimate the channel quality and explicitly signal this perceived channel quality information to the transmitter in the reverse link. Since in this case the receiver explicitly instructs the remote transmitter as to which modem modes to invoke, this regime is referred to as *closed-loop adaptation*. With the aid of this technique the adaptation algorithms can take into account effects such as interference as well as non-reciprocal channels. If the communication between the stations is essentially unidirectional, then a low-rate signalling channel must be implemented from the receiver to the transmitter. If such a channel exists, then the same technique as for non-reciprocal channels can be employed.

Different techniques can be employed to estimate the channel quality. For OFDM modems, the bit error probability in each subcarrier is determined by the fluctuations of the channel's instantaneous frequency domain channel transfer function  $H_n$ , if no interference is present. The estimate of the channel transfer function  $\hat{H}_n$  can be acquired by means of pilot tone based channel estimation, as demonstrated in Section 20.3.1.1. More accurate measures of the channel transfer function can be gained by means of decision directed or time domain training sequence based techniques. The estimate of the channel transfer function  $\hat{H}_n$  does not take into account effects, such as co-channel or intersubcarrier interference. Alternative channel quality measures that include interference effects can be devised on the basis of the error correction decoder's soft output information or by means of decision feedback local SNR estimations.

The delay between the channel quality estimate and the actual transmission of



the OFDM symbol in relation to the maximal Doppler frequency of the channel is crucial to the adaptive system's performance. If the channel estimate is obsolete at the time of transmission, then poor system performance will result. For a closed-loop adaptive system the delays between channel estimation and transmission of the packet are generally longer than for an open-loop adaptive system, and therefore the Doppler frequency of the channel is a more critical parameter for the system's performance than in the context of open-loop adaptive systems.

### 22.1.2.2 Parameter adaptation

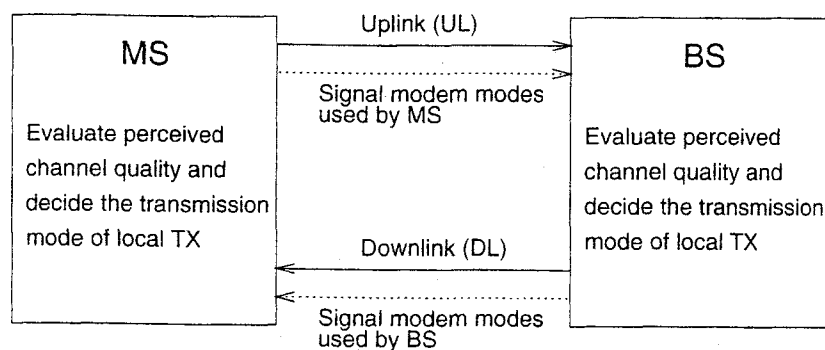
Different transmission parameters can be adapted to the anticipated channel conditions, such as the modulation and coding modes. Adapting the number of modulation levels in response to the anticipated local SNR encountered in each subcarrier can be employed, in order to achieve a wide range of different trade-offs between the received data integrity and throughput. Corrupted subcarriers can be excluded from data transmission and left blank or perhaps used for crest factor reduction.

The adaptive channel coding parameters entail code rate, adaptive interleaving and puncturing for convolutional and turbo codes, or varying block lengths for block codes [403]. These techniques can be combined with adaptive modulation mode selection and will be discussed in Chapter 23.

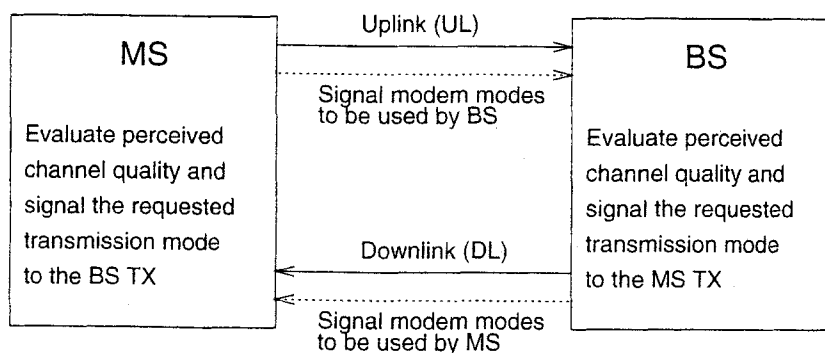
Based on the estimated frequency domain channel transfer function, spectral pre-distortion at the transmitter of one or both communicating stations can be invoked, in order to partially or fully counteract the frequency selective fading of the time dispersive channel. Unlike frequency domain equalisation at the receiver – which corrects for the amplitude and phase errors inflicted upon the subcarriers by the channel but cannot improve the signal-to-noise ratio in poor quality channels – spectral pre-distortion at the OFDM transmitter can deliver near-constant signal-to-noise levels for all subcarriers and can be thought of as power control on a subcarrier-by-subcarrier basis.

In addition to improving the system's BER performance in time dispersive channels, spectral pre-distortion can be employed in order to perform all channel estimation and equalisation functions at only one of the two communicating duplex stations. Low cost, low power consumption mobile stations can communicate with a base station that performs the channel estimation and frequency domain equalisation of the uplink, and uses the estimated channel transfer function for pre-distorting the down-link OFDM symbol. This set-up would lead to different overall channel quality on the uplink and the downlink, and the superior downlink channel quality could be exploited by using a computationally less complex channel decoder having weaker error correction capabilities in the mobile station than in the base station.

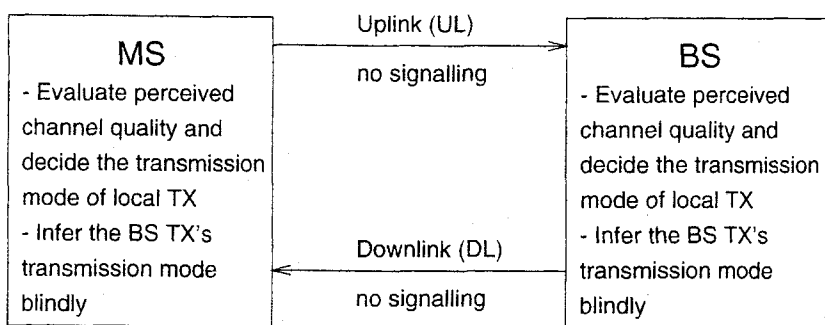
If the channel's frequency domain transfer function is to be fully counteracted by the spectral pre-distortion upon adapting the subcarrier power to the inverse of the channel transfer function, then the output power of the transmitter can become excessive, if heavily faded subcarriers are present in the system's frequency range. In order to limit the transmitter's maximal output power, hybrid channel pre-distortion and adaptive modulation schemes can be devised, which would deactivate transmission in deeply faded sub-channels, while retaining the benefits of pre-distortion in the



(a) Reciprocal channel, open-loop control



(b) Non-reciprocal channel, closed-loop signalling



(c) Reciprocal channel, blind modem-mode detection

**Figure 22.1:** Signalling scenarios in adaptive modems

remaining subcarriers.

### 22.1.2.3 Signalling the parameters

Signalling plays an important role in adaptive systems and the range of signalling options is summarised in Figure 22.1 for both open-loop and closed-loop signalling, as

well as for blind detection. If the channel quality estimation and parameter adaptation have been performed at the transmitter of a particular link, based on open-loop adaptation, then the resulting set of parameters has to be communicated to the receiver in order to successfully demodulate and decode the OFDM symbol. If the receiver itself determines the requested parameter set to be used by the remote transmitter, the closed-loop scenario, then the same amount of information has to be transported to the remote transmitter in the reverse link. If this signalling information is corrupted, then the receiver is generally unable to correctly decode the OFDM symbol corresponding to the incorrect signalling information.

Unlike adaptive serial systems, which employ the same set of parameters for all data symbols in a transmission packet [424, 425], adaptive OFDM systems have to react to the frequency selective nature of the channel, by adapting the modem parameters across the subcarriers. The resulting signalling overhead may become significantly higher than that for serial modems, and can be prohibitive for subcarrier-by-subcarrier modulation mode adaptation. In order to overcome these limitations, efficient and reliable signalling techniques have to be employed for practical implementation of adaptive OFDM modems.

If some flexibility in choosing the transmission parameters is sacrificed in an adaptation scheme, like in the sub-band adaptive OFDM schemes described below, then the amount of signalling can be reduced. Alternatively, blind parameter detection schemes can be devised, which require little or no signalling information, respectively. Two simple blind modulation scheme detection algorithms are investigated in Section 22.2.6.2 [431].

### 22.1.3 System aspects

The effects of transmission parameter adaptation for OFDM systems on the overall communication system have to be investigated in at least the following areas: data buffering and latency due to varying data throughput, the effects of co-channel interference and bandwidth efficiency.

## 22.2 Adaptive Modulation for OFDM

### 22.2.1 System model

The system model of the  $N$ -subcarrier orthogonal frequency division multiplexing (OFDM) modem is shown in Figure 22.2 [71]. At the transmitter, the modulator generates  $N$  data symbols  $S_n$ ,  $0 \leq n \leq N - 1$ , which are multiplexed to the  $N$  subcarriers. The time domain samples  $s_n$  transmitted during one OFDM symbol are generated by the inverse fast Fourier transform (IFFT) and transmitted over the channel after the cyclic extension (C. Ext.) has been inserted. The channel is modelled by its time-variant impulse response  $h(\tau, t)$  and AWGN. At the receiver, the cyclic extension is removed from the received time domain samples, and the data samples  $r_n$  are fast Fourier transformed, in order to yield the received frequency-domain data symbols  $R_n$ .

The channel's impulse response is assumed to be time invariant for the duration of

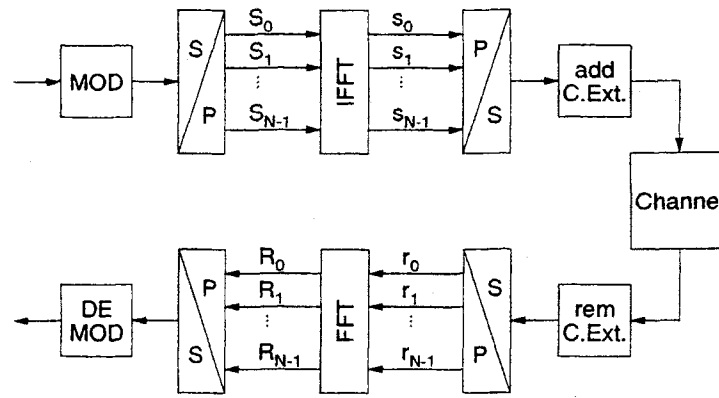


Figure 22.2: Schematic model of the OFDM system

one OFDM symbol, therefore it can be characterised for each OFDM symbol period by the  $N$ -point Fourier transform of the impulse response, which is referred to as the frequency domain channel transfer function  $H_n$ . The received data symbols  $R_n$  can be expressed as:

$$R_n = S_n \cdot H_n + n_n,$$

where  $n_n$  is an AWGN sample. Coherent detection is assumed for the system, therefore the received data symbols  $R_n$  have to be defaded in the frequency domain with the aid of an estimate of the channel transfer function  $H_n$ . This estimate  $\hat{H}_n$  can be obtained by the use of pilot subcarriers in the OFDM symbol, or by employing time domain channel sounding training sequences embedded in the transmitted signal. Since the noise energy in each subcarrier is independent of the channel's frequency domain transfer function  $H_n$ , the "local" signal-to-noise ratio SNR in subcarrier  $n$  can be expressed as

$$\gamma_n = |H_n|^2 \cdot \gamma,$$

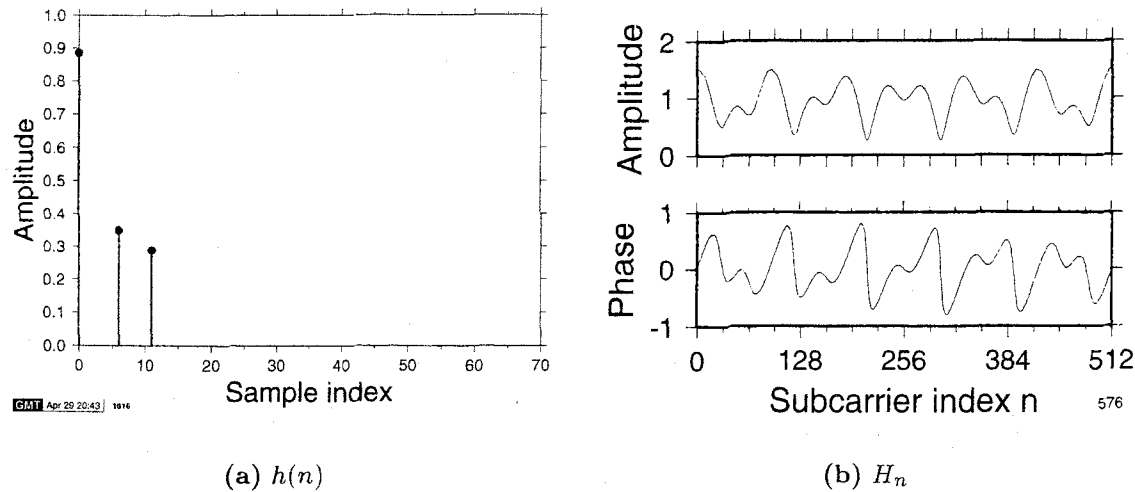
where  $\gamma$  is the overall SNR. If no signal degradation due to intersubcarrier interference (ISI) or interference from other sources appears, then the value of  $\gamma_n$  determines the bit error probability for the transmission of data symbols over the subcarrier  $n$ .

The goal of adaptive modulation is to choose the appropriate modulation mode for transmission in each subcarrier, given the local SNR  $\gamma_n$ , in order to achieve a good trade-off between throughput and overall BER. The acceptable overall BER varies depending on other systems parameters, such as the correction capability of the error correction coding and the nature of the service supported by this particular link.

The adaptive system has to fulfill these requirements:

- (1) Channel quality estimation
- (2) Choice of the appropriate modulation modes
- (3) Signalling or blind detection of the modulation modes

We will examine these three points with reference to Figure 22.1 in the following sections for the example of a 512-subcarrier OFDM modem in the shortened WATM channel of Section 20.1.1.2.



**Figure 22.3:** WATM wideband channel: (a) unfaded symbol spaced impulse response, (b) the corresponding frequency domain channel transfer function

### 22.2.2 Channel model

The impulse response  $h(\tau, t)$  used in our experiments was generated on the basis of the symbol-spaced impulse response shown in Figure 22.3(a) by fading each of the impulses obeying a Rayleigh distribution of a normalised maximal Doppler frequency of  $f'_d = 1.235 \cdot 10^{-5}$ , which corresponds to the WLAN channel experienced by a modem transmitting at a carrier frequency of 60 GHz with a sample rate of 225 MHz and a vehicular velocity of 50 km/h. The complex frequency domain channel transfer function  $H_n$  corresponding to the unfaded impulse response is shown in Figure 22.3(b).

### 22.2.3 Channel estimation

The most convenient setting for an adaptive OFDM (AOFDM) system is a time division duplex (TDD) system in a slowly varying reciprocal channel, allowing open-loop adaptation. Both stations transmit an OFDM symbol in turn, and at each station the most recent received symbol is used for the channel estimation employed for the modulation mode adaptation for the next transmitted OFDM symbol. The channel estimation on the basis of the received symbol can be performed by PSAM (see Section 20.3.1.1), or upon invoking more sophisticated methods, such as decision directed channel estimation. Initially, we will assume perfect knowledge of the channel transfer function during the received time slot.

### 22.2.4 Choice of the modulation modes

The two communicating stations use the open-loop predicted channel transfer function acquired from the most recent received OFDM symbol, in order to allocate the appropriate modulation modes to the subcarriers. The modulation modes were chosen from the set of binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), and 16-QAM. Page 53 of 163

(QPSK), 16-quadrature amplitude modulation (16-QAM), as well as “no transmission”, for which no signal was transmitted. These modulation modes are denoted by  $M_m$ , where  $m \in (0, 1, 2, 4)$  is the number of data bits associated with one data symbol of each mode.

In order to keep the system complexity low, the modulation mode is not varied on a subcarrier-by-subcarrier basis, but instead the total OFDM bandwidth of 512 subcarriers is split into blocks of adjacent subcarriers, referred to as sub-bands, and the same modulation scheme is employed for all subcarriers of the same sub-band. This substantially simplifies the task of signalling the modem mode and renders the employment of alternative blind detection mechanisms feasible, which will be discussed in Section 22.2.6.

Three modulation mode allocation algorithms were investigated in the sub-bands: a fixed threshold controlled algorithm, an upper bound BER estimator and a fixed throughput adaptation algorithm.

#### 22.2.4.1 Fixed threshold adaptation algorithm

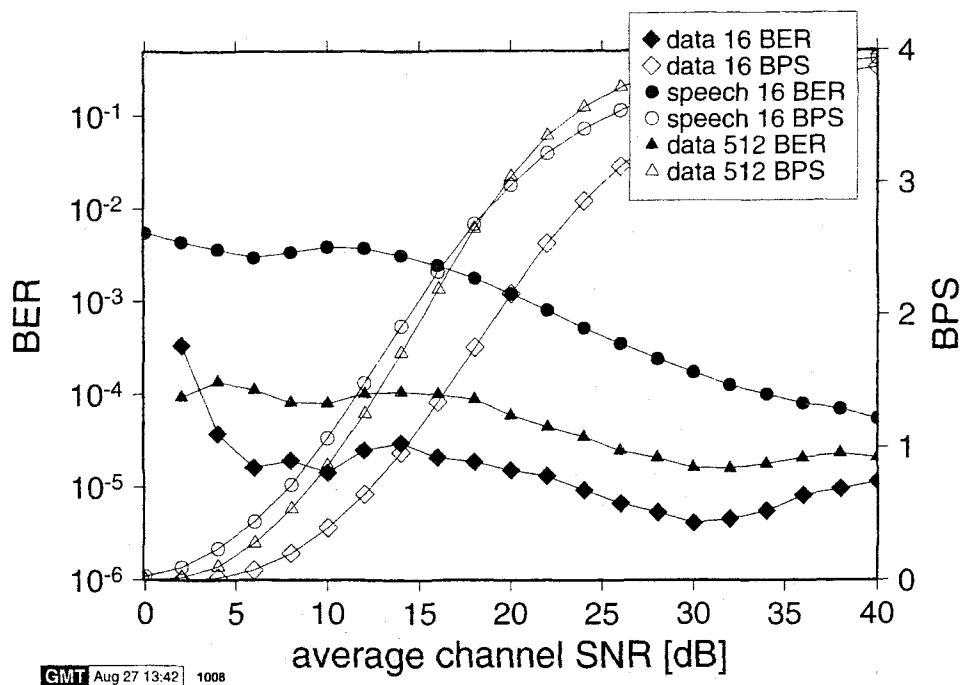
The fixed threshold algorithm was derived from the adaptation algorithm proposed by Torrance for serial modems [428]. In the case of a serial modem, the channel quality is assumed to be constant for all symbols in the time slot, and hence the channel has to be slowly varying in order to allow accurate channel quality prediction. Under these circumstances, all data symbols in the transmit time slot employ the same modulation mode, chosen according to the predicted SNR. The SNR thresholds for a given long-term target BER were determined by Powell optimisation [432]. Torrance assumed two uncoded target bit error rates: 1% for a high data rate “speech” system, and  $10^{-4}$  for a higher integrity, lower data rate “data” system. The resulting SNR thresholds  $l_n$  for activating a given modulation mode  $M_n$  in a slowly Rayleigh fading narrowband channel for both systems are given in Table 22.1. Specifically, the modulation mode

	$l_0$	$l_1$	$l_2$	$l_4$
speech system	$-\infty$	3.31	6.48	11.61
data system	$-\infty$	7.98	10.42	16.76

**Table 22.1:** Optimised switching levels for adaptive modulation over Rayleigh fading channels for the “speech” and “data” systems, shown in instantaneous channel SNR [dB] (from [432])

$M_n$  is selected if the instantaneous channel SNR exceeds the switching level  $l_n$ .

This adaptation algorithm originally assumed a constant instantaneous SNR over all of the block’s symbols, but in the case of an OFDM system in a frequency selective channel the channel quality varies across the different subcarriers. For sub-band adaptive OFDM transmission, this implies that if the sub-band width is wider than the channel’s coherence bandwidth [403], then the original switching algorithm cannot be employed. For our investigations, we have therefore employed the lowest quality subcarrier in the sub-band for the adaptation algorithm based on the thresholds given in Table 22.1. The performance of the 16 sub-band adaptive system over the shortened

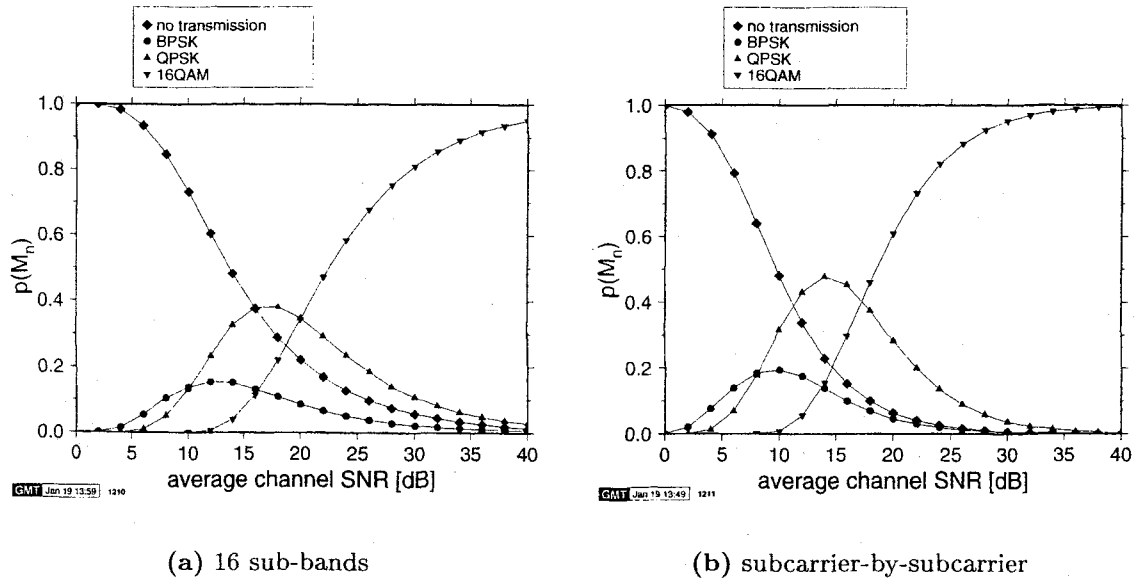


**Figure 22.4:** BER and BPS throughput performance of the 16 sub-band, 512 subcarrier switching level adaptive OFDM modem employing BPSK, QPSK, 16-QAM and “no transmission” over the Rayleigh fading time dispersive channel of Figure 20.3 using the switching thresholds of Table 22.1

WATM Rayleigh fading channel of Figure 20.3 is shown in Figure 22.4.

Adjacent or consecutive time slots have been used for the uplink and downlink slots in these simulations, so that the delay between channel estimation and transmission was rendered as short as possible. Figure 22.4 shows the long-term average BER and throughput of the studied modem for the “speech” and “data” switching levels of Table 22.1 as well as for a subcarrier-by-subcarrier adaptive modem employing the “data” switching levels. The results show the typical behaviour of a variable throughput AOFDM system, which constitutes a trade-off between the best BER and best throughput performance. For low SNR values, the system achieves a low BER by transmitting very few bits and only when the channel conditions allow. With increasing long-term SNR, the throughput increases without significant change in the BER. For high SNR values the BER drops as the throughput approaches its maximum of 4 bits per symbol, since the highest-order constellation was 16-QAM.

It can be seen from the figure that the adaptive system performs better than its target bit error rates of  $10^{-2}$  and  $10^{-4}$  for the “speech” and “data” systems, respectively, resulting in measured bit error rates lower than the targets. This can be explained by the adaptation regime, which was based on the conservative principle of using the lowest quality subcarrier in each sub-band for channel quality estimation, leading to a pessimistic channel quality estimate for the entire sub-band. For low values of SNR, the throughput in bits per data symbol is low and exceeds the fixed BPSK throughput of 1 bit/symbol only for SNR values in excess of 9.5 dB and 14 dB



**Figure 22.5:** Histograms of modulation modes versus channel SNR for the “data” switching level adaptive 512 subcarrier, 16 sub-band OFDM modem over the Rayleigh fading time dispersive channel of Figure 20.3 using the switching thresholds of Table 22.1

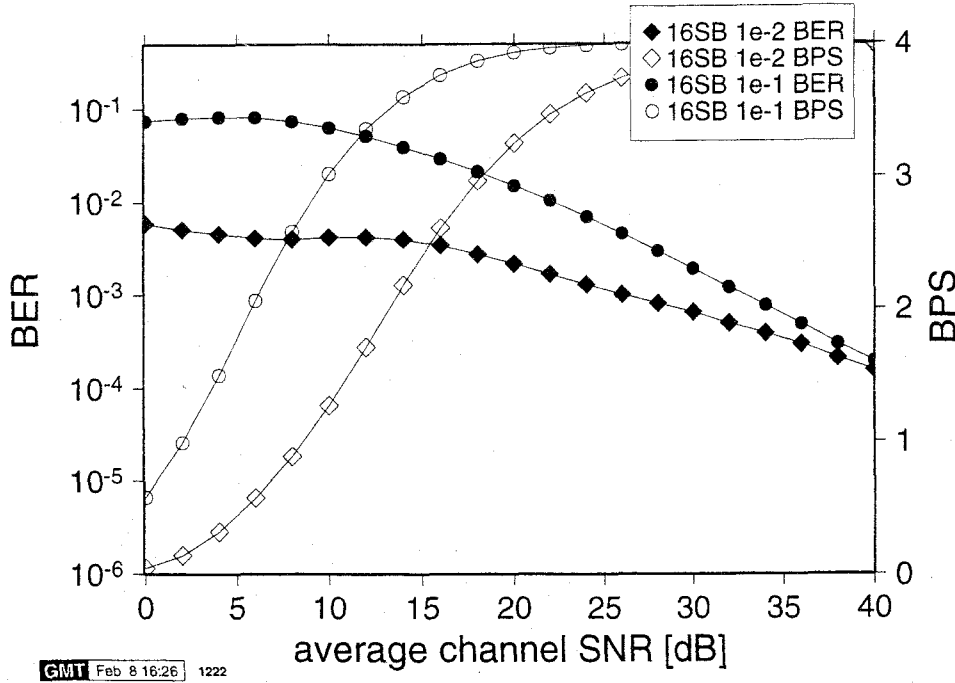
for the “speech” and “data” systems, respectively.

The upper bound performance of the system with subcarrier-by-subcarrier adaptation is also portrayed in the figure, shown as 512 independent sub-bands, for the “data” optimised set of threshold values. It can be seen that in this case the target BER of  $10^{-4}$  is closely met over a wide range of SNR values from about 2 dB to 20 dB, and that the throughput is considerably higher than in the case of the 16 sub-band modem. This is the result of more accurate subcarrier-by-subcarrier channel quality estimation and fine-grained adaptation, leading to better exploitation of the available channel capacity.

Figure 22.5 shows the long-term modulation mode histograms for a range of channel SNR values for the “data” switching levels in both the 16 sub-band and the subcarrier-by-subcarrier adaptive modems using the switching thresholds of Table 22.1. Comparison of the graphs shows that higher-order modulation modes are used more frequently by the subcarrier-by-subcarrier adaptation algorithm, which is in accordance with the overall throughput performance of the two modems in Figure 22.4.

The throughput penalty of employing sub-band adaptation depends on the frequency domain variation of the channel transfer function. If the sub-band bandwidth is lower than the channel’s coherence bandwidth, then the assumption of constant channel quality per sub-band is closely met, and the system performance is equivalent to that of a subcarrier-by-subcarrier adaptive scheme.





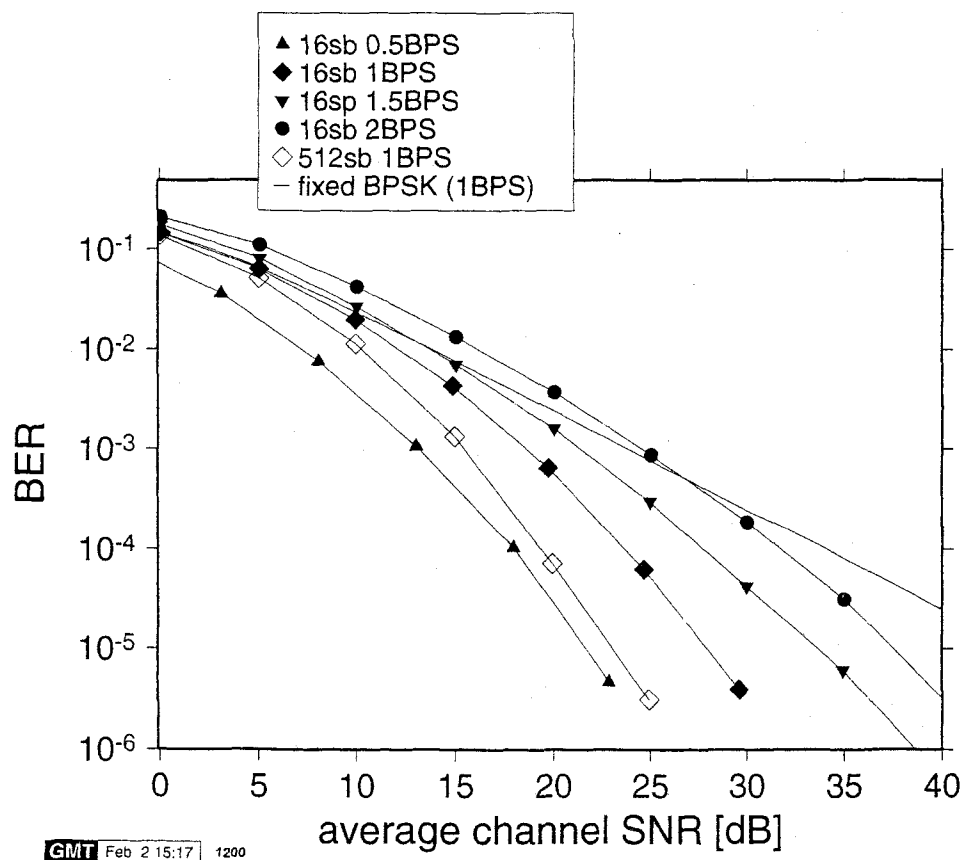
**Figure 22.6:** BER and BPS throughput performance of the 16 sub-band, 512 subcarrier BER estimator adaptive OFDM modem employing BPSK, QPSK, 16-QAM and “no transmission” over the Rayleigh fading time dispersive channel of Figure 20.3

#### 22.2.4.2 Sub-band BER estimator adaptation algorithm

We have seen above that the fixed switching level based algorithm leads to a throughput performance penalty, if used in a sub-band adaptive OFDM modem, when the channel quality is not constant throughout each sub-band. This is due to the conservative adaptation based on the subcarrier experiencing the most hostile channel in each sub-band.

An alternative scheme taking into account the non-constant SNR values  $\gamma_j$  across the  $N_s$  subcarriers in the  $j$ th sub-band can be devised by calculating the expected overall bit error probability for all available modulation modes  $M_n$  in each sub-band, which is denoted by  $\bar{p}_e(n) = 1/N_s \sum_j p_e(\gamma_j, M_n)$ . For each sub-band, the mode having the highest throughput, whose estimated BER is lower than a given threshold, is then chosen. While the adaptation granularity is still limited to the sub-band width, the channel quality estimation includes not only the lowest quality subcarrier, which leads to an improved throughput.

Figure 22.6 shows the BER and throughput performance for the 16 sub-band adaptive OFDM modem employing the BER estimator adaptation algorithm in the Rayleigh fading time dispersive channel of Figure 20.3. The two sets of curves in the figure correspond to target bit error rates of  $10^{-2}$  and  $10^{-1}$ , respectively. Comparing the modem’s performance for a target BER of  $10^{-2}$  with that of the “speech” modem in Figure 22.4, it can be seen that the BER estimator algorithm results in significantly higher throughput while meeting the BER requirements. The BER estimator algo-



**Figure 22.7:** BER performance versus SNR for the 512 subcarrier, 16 sub-band constant throughput adaptive OFDM modem employing BPSK, QPSK, 16-QAM and “no transmission” in the Rayleigh fading time dispersive channel of Figure 20.3 for 0.5, 1, 1.5 and 2 bits per symbol (BPS) target throughput

rithm is readily adjustable to different target bit error rates, which is demonstrated in the figure for a target BER of  $10^{-1}$ . Such adjustability is beneficial when combining adaptive modulation with channel coding, as will be discussed in Section 22.2.7.

### 22.2.5 Constant throughput adaptive OFDM

The time-varying data throughput of an adaptive OFDM modem operating with either of the two adaptation algorithms discussed above makes it difficult to employ such a scheme in a variety of constant rate applications. Torrance [428] studied the system implications of variable-throughput adaptive modems in the context of narrowband channels, stressing the importance of data buffering at the transmitter, in order to accommodate the variable data rate. The required length of the buffer is related to the Doppler frequency of the channel, and a slowly varying channel – as required for adaptive modulation – results in slowly varying data throughput and therefore the need for a high buffer capacity. Real-time interactive audio or video transmission is sensitive to delays, and therefore different modem mode adaptation algorithms are needed for such applications.

The constant throughput AOFDM scheme proposed here exploits the frequency selectivity of the channel, while offering a constant bit rate. Again, sub-band adaptivity is assumed, in order to simplify the signalling or the associated blind detection of the modem schemes.

The modulation mode allocation of the sub-bands is performed on the basis of a cost function to be introduced below, based on the expected number of bit errors in each sub-band. The expected number of bit errors  $e_{n,s}$ , for each sub-band  $n$  and for each possible modulation mode index  $s$ , is calculated on the basis of the estimated channel transfer function  $\hat{H}$ , taking into account the number of bits transmitted per sub-band and per modulation mode,  $b_{n,s}$ .

Each sub-band is assigned a state variable  $s_n$  holding the index of a modulation mode. Each state variable is initialised to the lowest-order modulation mode, which in our case is 0 for “no transmission”. A set of cost values  $c_{n,s}$  is calculated for each sub-band  $n$  and state  $s$ :

$$c_{n,s} = \frac{e_{n,s+1} - e_{n,s}}{b_{n,s+1} - b_{n,s}} \quad (22.1)$$

for all but the highest modulation mode index  $s$ . This cost value is related to the expected increase in the number of bit errors, divided by the increase of throughput, if the modulation mode having the next higher index is used instead of index  $s$  in sub-band  $n$ . In other words, Equation 22.1 quantifies the expected incremental bit error rate of the state transition  $s \rightarrow s + 1$  in sub-band  $n$ .

The modulation mode adaptation is performed by repeatedly searching for the block  $n$  having the lowest value of  $c_{n,s_n}$ , and incrementing its state  $s_n$ . This is repeated until the total number of bits in the OFDM symbol reaches the target number of bits. Because of the granularity in bit numbers introduced by the sub-bands, the total number of bits may exceed the target. In this case, the data is padded with dummy bits for transmission.

Figure 22.7 gives an overview of the BER performance of the fixed throughput 512-subcarrier OFDM modem over the time dispersive channel of Figure 20.3 for a range of target bit numbers. The curve without shapes represents the performance of a fixed BPSK OFDM modem over the same channel and where the modem transmits, i.e. 1 bit over each data subcarrier per OFDM symbol. The diamond shapes give the performance of the equivalent throughput adaptive scheme for the 16 sub-band arrangement filled shapes and for the subcarrier-by-subcarrier adaptive scheme open shapes. It can be seen that the 16 sub-band adaptive scheme yields a significant improvement in BER terms for SNR values above 10 dB. The SNR gain for a bit error rate of  $10^{-4}$  is 8 dB compared to the non-adaptive case. Subcarrier-by-subcarrier adaptivity increases this gain by a further 4 dB. The modem can readily be adapted to the system requirements by adjusting the target bit rate, as shown in Figure 22.7. Halving the throughput to 0.5 BPS, the required SNR is reduced by 6 dB for a BER of  $10^{-4}$ , while increasing the throughput to 2 BPS deteriorates the noise resilience by 8 dB at the same BER.

### 22.2.6 Signalling and blind detection

The adaptive OFDM receiver has to be informed of the modulation modes used for the different sub-bands. This information can either be conveyed using signalling subcarriers in the OFDM symbol itself, or the receiver can employ blind detection techniques in order to estimate the transmitted symbols' modulation modes, as seen in Figure 22.1.

#### 22.2.6.1 Signalling

The simplest way of signalling the modulation mode employed in a sub-band is to replace one data symbol by an  $M$ -PSK symbol, where  $M$  is the number of possible modulation modes. In this case, reception of each of the constellation points directly signals a particular modulation mode in the current sub-band. In our case, for four modulation modes and assuming perfect phase recovery, the probability of a signalling error  $p_s(\gamma)$ , when employing one signalling symbol, is the symbol error probability of QPSK. Then the correct sub-band mode signalling probability is:

$$(1 - p_s(\gamma)) = (1 - p_{b,QPSK}(\gamma))^2,$$

where  $p_{b,QPSK}$  is the bit error probability for QPSK:

$$p_{b,QPSK}(\gamma) = Q(\sqrt{\gamma}) = \frac{1}{2} \cdot \text{erfc} \left( \sqrt{\frac{\gamma}{2}} \right),$$

which leads to the expression for the modulation mode signalling error probability of

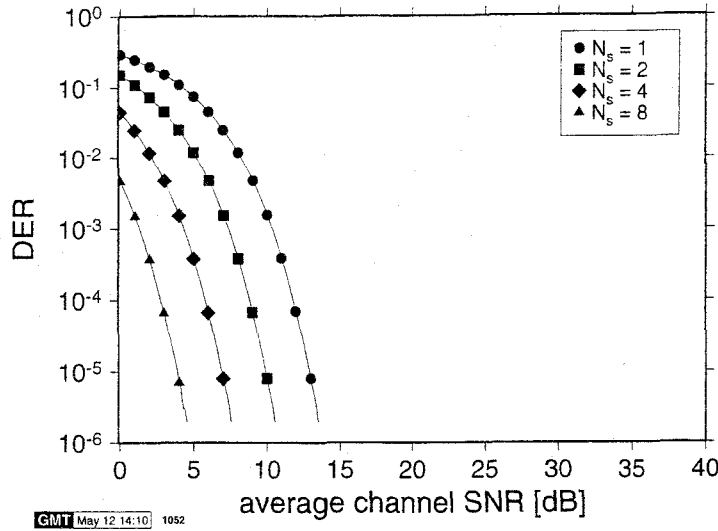
$$p_s(\gamma) = 1 - \left( 1 - \frac{1}{2} \cdot \text{erfc} \left( \sqrt{\frac{\gamma}{2}} \right) \right)^2.$$

The modulation mode signalling error probability can be reduced by employing multiple signalling symbols and maximum ratio combining of the received signalling symbols  $R_{s,n}$ , in order to generate the decision variable  $R'_s$  prior to decision:

$$R'_s = \sum_{n=1}^{N_s} R_{s,n} \cdot \hat{H}_{s,n}^*,$$

where  $N_s$  is the number of signalling symbols per sub-band, the quantities  $R_{s,n}$  are the received symbols in the signalling subcarriers, and  $\hat{H}_{s,n}$  represents the estimated values of the frequency domain channel transfer function at the signalling subcarriers. Assuming perfect channel estimation and constant values of the channel transfer function across the group of signalling subcarriers, the signalling error probability for  $N_s$  signalling symbols can be expressed as:

$$p'_s(\gamma, N_s) = 1 - \left( 1 - \frac{1}{2} \cdot \text{erfc} \left( \sqrt{\frac{N_s \gamma}{2}} \right) \right)^2. \quad (22.2)$$



**Figure 22.8:** Modulation mode detection error ratio (DER) if signalling with maximum ratio combining is employed for QPSK symbols in an AWGN channel for 1, 2, 4 and 8 signalling symbols per sub-band, evaluated from Equation 22.2

Figure 22.8 shows the signalling error rate in an AWGN channel for 1, 2, 4 and 8 signalling symbols per sub-band, respectively. It can be seen that doubling the number of signalling subcarriers improves the performance by 3 dB. Modem mode detection error ratios (DER) below  $10^{-5}$  can be achieved at 10 dB SNR over AWGN channels if two signalling symbols are used. The signalling symbols for a given sub-band can be interleaved across the entire OFDM symbol bandwidth, in order to benefit from frequency diversity in fading wideband channels.

As seen in Figure 22.1, blind detection algorithms aim to estimate the employed modulation mode directly from the received data symbols, therefore avoiding the loss of data capacity due to signalling subcarriers. Two algorithms have been investigated, one based on geometrical SNR estimation and another incorporating error correction coding.

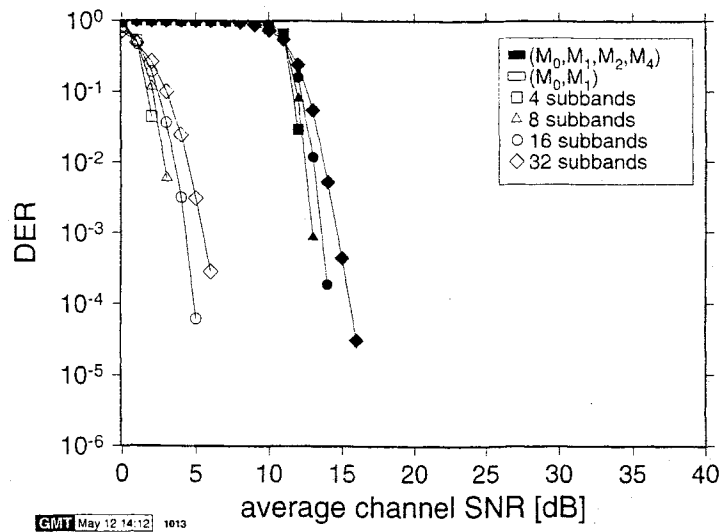
#### 22.2.6.2 Blind detection by SNR estimation

The receiver has no a-priori knowledge of the modulation mode employed in a particular received sub-band and estimates this parameter by quantising the defaded received data symbols  $R_n/\hat{H}_n$  in the sub-band to the closest symbol  $\hat{R}_{n,m}$  for all possible modulation modes  $M_m$  for each subcarrier index  $n$  in the current sub-band. The decision directed error energy  $e_m$  for each modulation mode is calculated:

$$e_m = \sum_n \left( R_n/\hat{H}_n - \hat{R}_{n,m} \right)^2$$

and the modulation mode  $M_m$  which minimises  $e_m$  is chosen for the demodulation of the sub-band.

The DER of the blind modulation mode detection algorithm described in this sec-

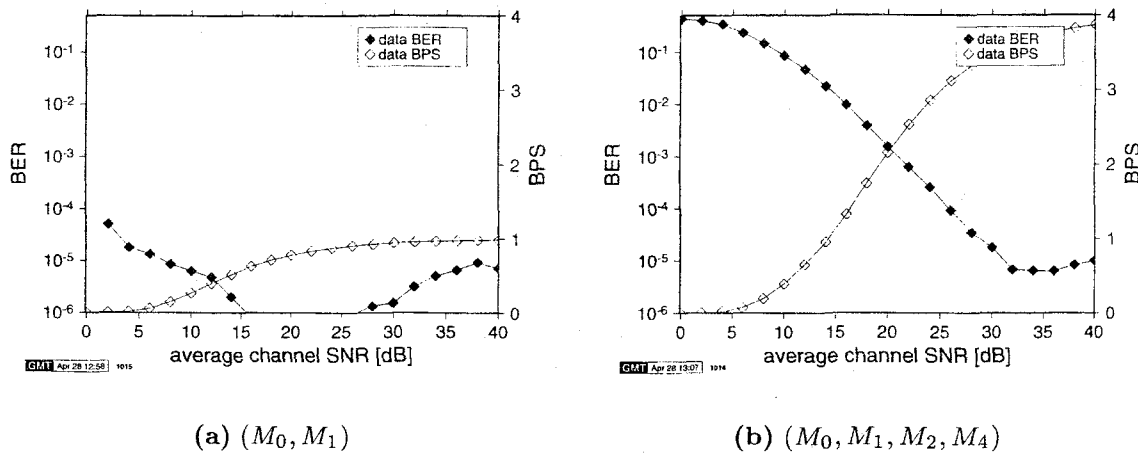


**Figure 22.9:** Blind modulation mode detection error ratio (DER) for 512-subcarrier OFDM systems employing  $(M_0, M_1)$  as well as for  $(M_0, M_1, M_2, M_4)$  for different numbers of sub-bands in an AWGN channel

tion for a 512-subcarrier OFDM modem in an AWGN channel is depicted in Figure 22.9. It can be seen that the detection performance depends on the number of symbols per sub-band, with fewer sub-bands and therefore longer symbol sequences per sub-band leading to a better detection performance. It is apparent, however, that the number of available modulation modes has a more significant effect on the detection reliability than the block length. If all four legitimate modem modes are employed, then reliable detection of the modulation mode is only guaranteed for AWGN SNR values of more than 15-18 dB, depending on the number of sub-bands per OFDM symbol. If only  $M_0$  and  $M_1$  are employed, however, the estimation accuracy is dramatically improved. In this case, AWGN SNR values above 5-7 dB are sufficient to ensure reliable detection. The estimation accuracy could be improved by using the estimate of the channel quality, in order to predict the modulation mode, which is likely to have been employed at the transmitter. For example, at an estimated channel SNR of 5 dB it is unlikely that 16-QAM was employed as the modem mode and hence this a priori knowledge can be exploited, in order to increase our confidence in the corresponding decision.

Figure 22.10 shows the BER performance of the fixed threshold “data” 16 sub-band adaptive system in the fading wideband channel of Figure 20.3 for both sets of modulation modes, namely for  $(M_0, M_1)$  and  $(M_0, M_1, M_2, M_4)$  with blind modulation mode detection. Erroneous modulation mode decisions were assumed to yield a BER of 50% in the received block. This is optimistic, since in a realistic scenario the receiver would have no knowledge of the number of bits actually transmitted, leading to loss of synchronisation in the data stream. This problem is faced by all systems having a variable throughput and not employing an ideal reliable signalling channel. This impediment must be mitigated by data synchronisation measures.

It can be seen from Figure 22.10 that while blind modulation mode detection yields



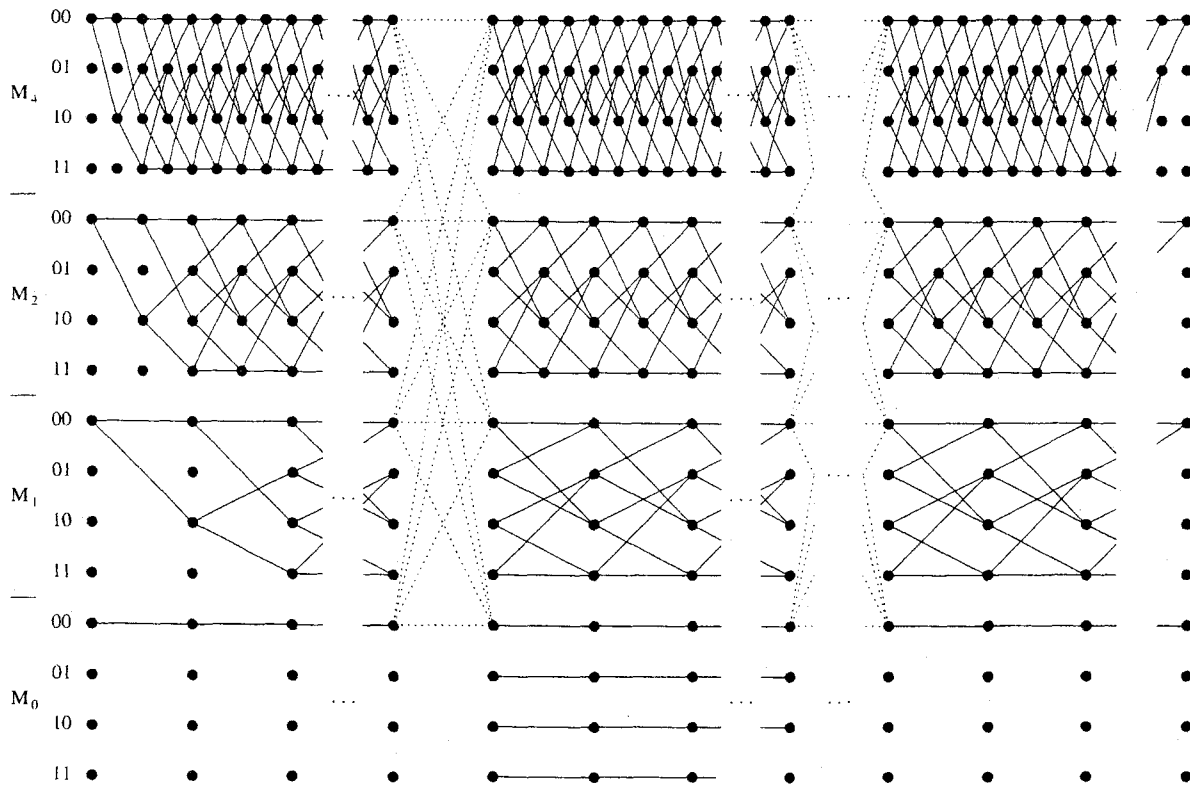
**Figure 22.10:** BER and BPS throughput performance of a 16 sub-band, 512 subcarrier adaptive OFDM modem employing (a) no transmission ( $M_0$ ) and BPSK ( $M_1$ ), or (b) ( $M_0, M_1, M_2, M_4$ ), both using the data-type switching levels of Table 22.1 and the SNR-based blind modulation mode detection of Section 22.2.6.2 over the Rayleigh fading time dispersive channel of Figure 20.3

poor performance for the quadruple-mode adaptive scheme, the twin-mode scheme exhibits BER results consistently better than  $10^{-4}$ .

### 22.2.6.3 Blind detection by multi-mode trellis decoder

If error correction coding is invoked in the system, then the channel decoder can be employed to estimate the most likely modulation mode per sub-band. Since the number of bits per OFDM symbol is varying in this adaptive scheme, and the channel encoder's block length is therefore not constant, for the sake of implementational convenience we have chosen a convolutional encoder at the transmitter. Once the modulation modes to be used are decided upon at the transmitter, the convolutional encoder is employed to generate a zero-terminated code word having the length of the OFDM symbol's capacity. This code word is modulated on the subcarriers according to the different modulation modes for the different sub-bands, and the OFDM symbol is transmitted over the channel.

At the receiver, each received data subcarrier is demodulated by all possible demodulators, and the resulting hard decision bits are fed into parallel trellises for Viterbi decoding. Figure 22.11 shows a schematic sketch of the resulting parallel trellis if 16-QAM ( $M_4$ ), QPSK ( $M_2$ ), BPSK ( $M_1$ ), and "no transmission" ( $M_0$ ) are employed, for a convolutional code having four states. Each sub-band in the adaptive scheme corresponds to a set of four parallel trellises, whose inputs are generated independently by the four demodulators of the legitimate modulation modes. The number of transitions in each of the trellises depends on the number of output bits received from the different demodulators, so that the 16-QAM ( $M_4$ ) trellis contains four times as many transitions as the BPSK and "no transmission" trellises. Since in



**Figure 22.11:** Schematic plot of the parallel trellises for blind modulation mode detection employing convolutional coding. In this example, a four-state 00-terminated convolutional encoder was assumed. The dotted lines indicate the intersub-band transitions for the 00 state, and are omitted for the other three states

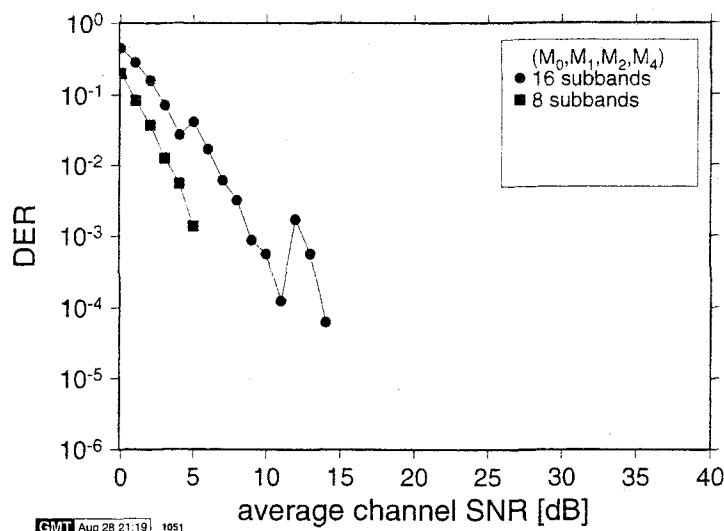
the case of “no transmission” no coded bits are transmitted, the state of the encoder does not change. Therefore, legitimate transitions for this case are only horizontal ones.

At sub-band boundaries, transitions are allowed between the same state of all the parallel trellises associated with the different modulation modes. This is not a transition due to a received bit, and therefore preserves the metric of the originating state. Note that in the figure only the possible allowed transitions for the state 00 are drawn; all other states originate the equivalent set of transitions. The initial state of the first sub-band is 00 for all modulation modes, and since the code is terminated in 00, the last sub-band’s final states are 00.

The receiver’s Viterbi decoder calculates the metrics for the transitions in the parallel trellises, and once all the data symbols have been processed, it traces back through the parallel trellis on the surviving path. This backtracing commences at the most likely 00 state at the end of the last sub-band. If no termination was used at the decoder, then the backtracing would start at the most likely of all the final states of the last block.

Figure 22.12 shows the modulation mode detection error ratio (DER) for the parallel trellis decoder in an AWGN channel for 16 and 8 sub-bands, if a convolutional code of constraint length 7 is used. Comparison with Figure 22.9 shows consider-





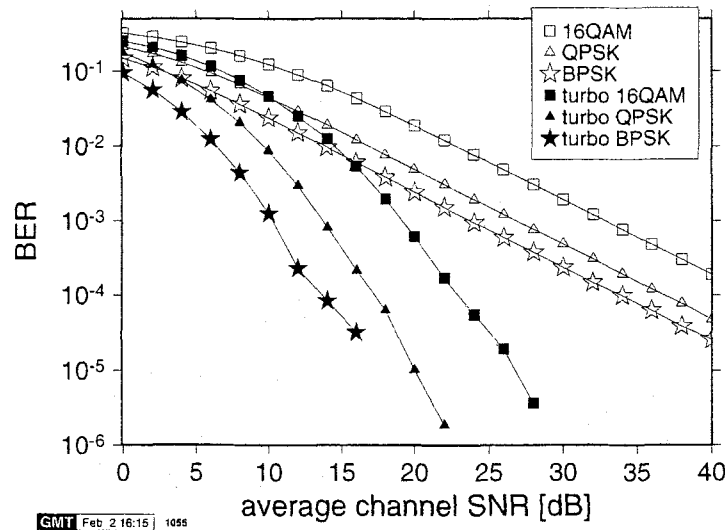
**Figure 22.12:** Blind modulation mode detection error ratio (DER) using the parallel trellis algorithm of Section 22.2.6.3 with a  $K = 7$  convolutional code in an AWGN channel for a 512-subcarrier OFDM modem

able improvements relative to the BER estimation based blind detection scheme of Section 22.2.6.2, both for 16 and 8 sub-bands. Higher sub-band lengths improve the estimation accuracy by a greater degree than has been observed for the BER estimation algorithm of Figure 22.9. A DER of less than  $10^{-5}$  was observed for AWGN SNR values of 6 dB and 15 dB in the 8 and 16 sub-band scenarios, respectively. The use of stronger codes could further improve the estimation accuracy, at the cost of higher complexity.

### 22.2.7 Sub-band adaptive OFDM and channel coding

Adaptive modulation can reduce the BER to a level where channel decoders can perform well. Figure 22.13 shows both the uncoded and coded BER performance of a 512-subcarrier OFDM modem in the fading wideband channel of Figure 20.3, assuming perfect channel estimation. The channel coding employed in this set of experiments was a turbo coder [433] with a data block length of 1000 bits, employing a random interleaver and 8 decoder iterations. The log-MAP decoding algorithm was used [434]. The constituent half-rate convolutional encoders were of constraint length 3, with octally represented generator polynomials of (7,5) [403]. It can be seen that the turbo decoder provides a considerable coding gain for the different fixed modulation schemes, with a BER of  $10^{-4}$  for SNR values of 13.8 dB, 17.3 dB and 23.2 dB for BPSK, QPSK and 16-QAM transmission, respectively.

Figure 22.14 depicts the BER and throughput performance of the same decoder employed in conjunction with the adaptive OFDM modem for different adaptation algorithms. Figure 22.14(a) shows the performance for the “speech” system employing the switching levels listed in Table 22.1. As expected, the half-rate channel coding results in a halved throughput compared to the uncoded case, but offers low BER transmission over the channel of Figure 20.3 for SNR values of down to 0 dB, main-



**Figure 22.13:** BER performance of the 512-subcarrier OFDM modem in the fading time dispersive channel of Figure 20.3 for both uncoded and half-rate turbo-coded transmission, using 8-iteration log-MAP turbo decoding, 1000-bit random interleaver, and a constraint length of 3

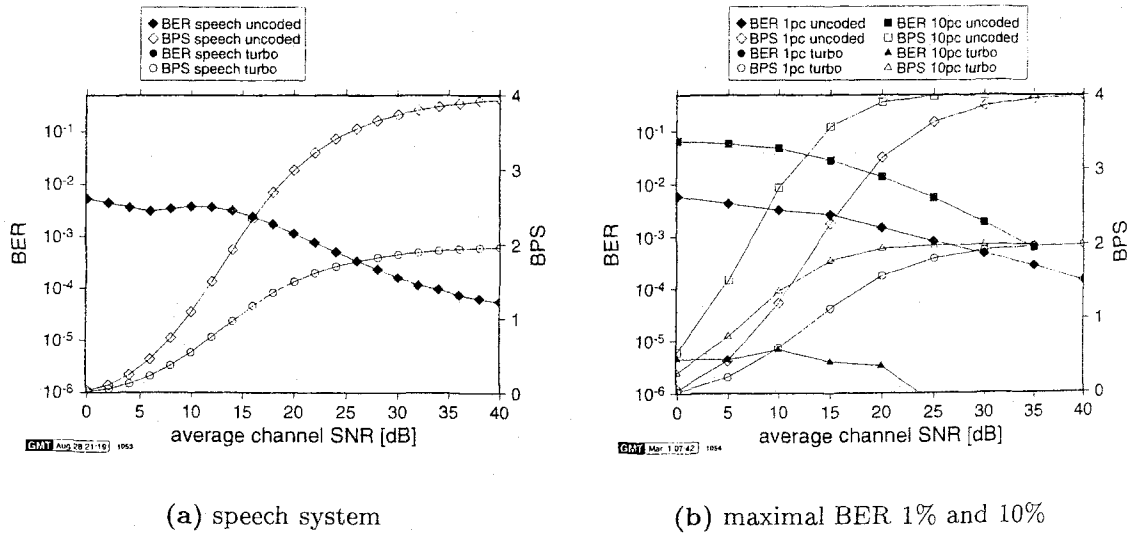
taining a BER below  $10^{-6}$ .

Further tuning of the adaptation parameters can ensure a better average throughput, while retaining error-free data transmission. The switching level based adaptation algorithm of Table 22.1 is difficult to control for arbitrary bit error rates, since the set of switching levels was determined by an optimisation process for uncoded transmission. Since the turbo codec has a non-linear BER versus SNR characteristic, direct switching level optimisation is an arduous task. The sub-band BER predictor of Section 22.2.4.2 is easier to adapt to a channel codec, and Figure 22.14(b) shows the performance for the same decoder, with the adaptation algorithm employing the BER prediction method having an upper BER bound of 1%. It can be seen that the less stringent uncoded BER constraints when compared to Figure 22.14(a) lead to a significantly higher throughput for low SNR values. The turbo-decoded data bits are error-free, hence a further increase in throughput is possible while maintaining a high degree of coded data integrity.

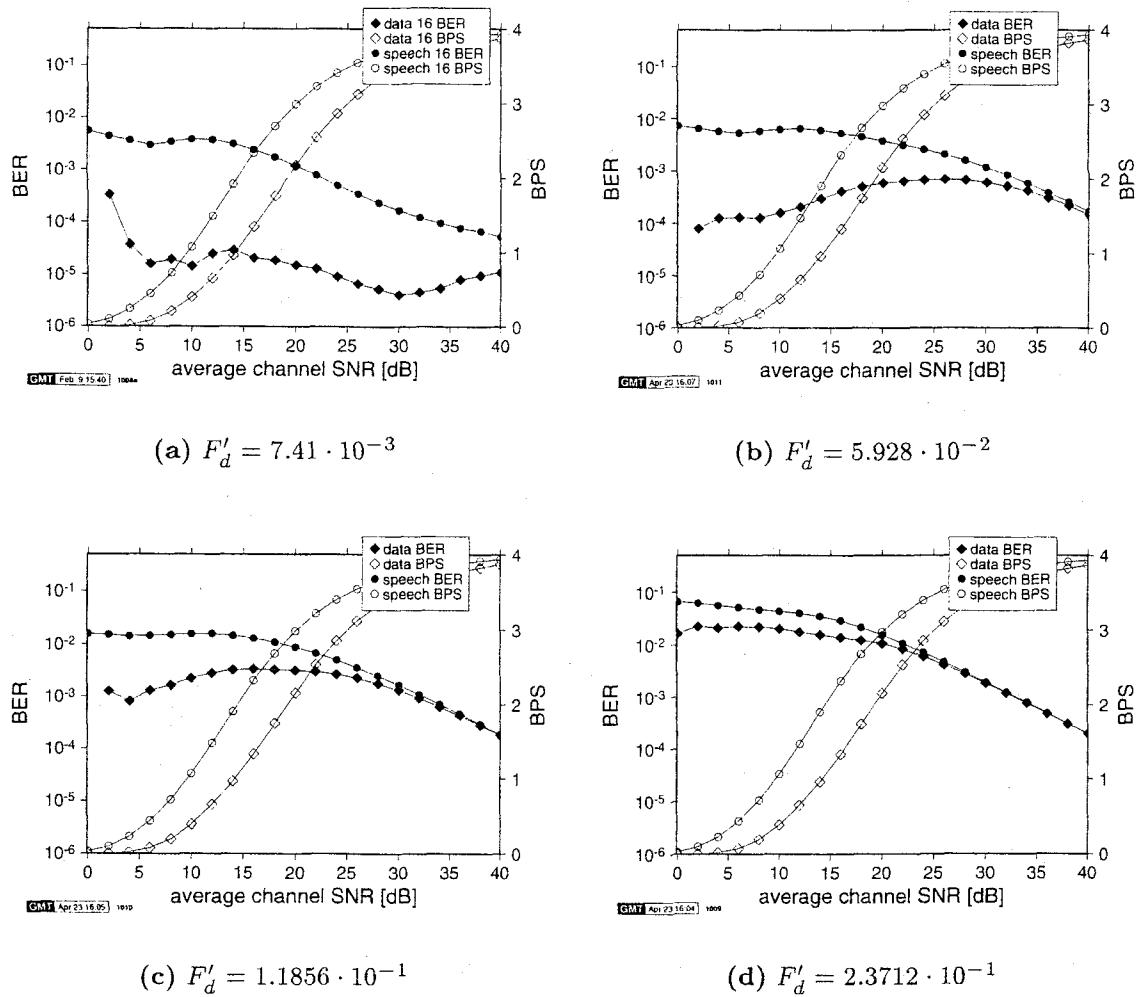
The second set of curves in Figure 22.14(b) show the system's performance, if an uncoded target BER of 10% is assumed. In this case, the turbo decoder's output BER is below  $10^{-5}$  for all the SNR values plotted, and shows a slow decrease for increasing values of SNR. The throughput of the system, however, exceeds 0.5 data bits per symbol for SNR values of more than 2 dB.

### 22.2.8 The effect of channel Doppler frequency

Since the adaptive OFDM modem employs the most recently received OFDM symbol in order to predict the frequency domain transfer function of the reverse channel for the next transmission, the quality of this prediction suffers from the time variance



**Figure 22.14:** BER and BPS throughput performance of 16 sub-band, 512 subcarrier adaptive turbo coded and uncoded OFDM modem employing  $(M_0, M_1, M_2, M_4)$  for (a) speech-type switching levels of Table 22.1 and (b) a maximal estimated sub-band BER of 1% and 10% over the channel of Figure 20.3. The turbo-coded transmission over the speech system and the 1% maximal BER system are error-free for all examined SNR values and therefore the corresponding BER curves are omitted from the graphs, hence the lack of black circles on (a) and (b)



**Figure 22.15:** BER and BPS throughput performance of 16 sub-band, 512 subcarrier adaptive OFDM modem employing  $(M_0, M_1, M_2, M_4)$  for both data-type and speech-type switching levels with perfect modulation mode detection and different frame normalised Doppler frequencies  $F'_d$  over the channel of Figure 20.3. The triangular markers in (a) show the performance of a subcarrier-by-subcarrier adaptive modem using the data-type switching levels of Table 22.1 for comparison

of the channel transfer function between the uplink and downlink time slots. We assume that the time delay between the uplink and downlink slots is the same as the delay between the downlink and uplink slots, and we refer to this time as the frame duration  $T_f$ . We normalise the maximal Doppler frequency  $f_d$  of the channel to the frame duration  $T_f$ , and define the *frame normalised Doppler frequency*  $F'_d$  as  $F'_d = f_d \cdot T_f$ . Figure 22.15 depicts the fixed switching level (see Table 22.1) modem's BER and throughput performance in bits per symbol (BPS) for values of  $F'_d$  between  $7.41 \cdot 10^{-3}$  and  $2.3712 \cdot 10^{-1}$ . These values stem from the studied WATM system with a time slot duration of  $2.67 \mu s$  and up/downlink delays of 1, 8, 16 and 32 time slots

corresponds to a system employing a carrier frequency of 60 GHz, a sampling rate of 225 Msamples/s and a vehicular velocity of 50 km/h or 13.8 m/s.

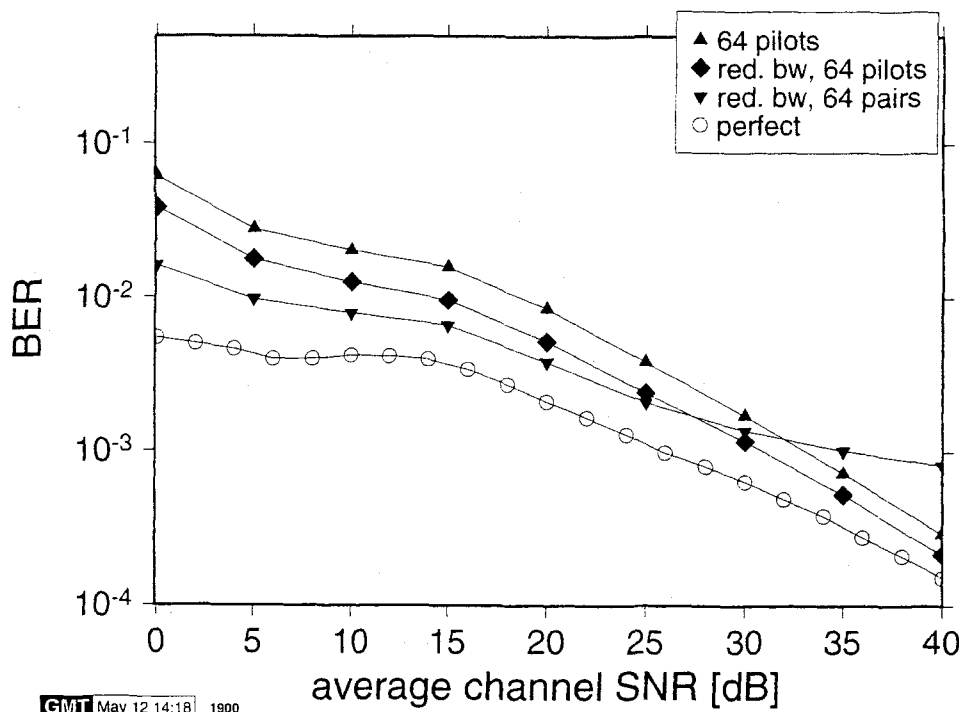
Figure 22.15(a) shows the BER and BPS throughput of the studied modems in a framework with consecutive uplink and downlink time slots. This corresponds to  $F'_d = 7.41 \cdot 10^{-3}$ , while the target bit error rates for the speech and data system are met for all SNR values above 4 dB, and the BER performance is generally better than the target error rates. This was explained above with the conservative choice of modulation modes based on the most corrupted subcarrier in each sub-band, resulting in lower throughput and lower bit error rates for the switching level based sub-band adaptive modem.

Comparing Figure 22.15(a) with the other performance curves, it can be seen that the bit error rate performance for both the speech and the data system suffer from increasing decorrelation of the predicted and actual channel transfer function for increasing values of  $F'_d$ . In Figure 22.15(b) an 8 time slot delay was assumed between uplink and downlink time slots, which corresponds to  $F'_d = 5.928 \cdot 10^{-2}$ , and therefore the BER performance of the modem was significantly deteriorated. The “speech” system still maintains its target BER, but the “data” system delivers a BER of up to  $10^{-3}$  for SNR values between 25 and 30 dB. It is interesting to observe that the delayed channel prediction mainly affects the higher-order modulation modes, which are employed more frequently at high SNR values. This explains the shape of the BER curve for the “data” system, which is rising from below  $10^{-4}$  at 2 dB SNR up to  $10^{-3}$  at 26 dB SNR. The average throughput of the modem is mainly determined by the statistics of the estimated channel transfer function at the receiver, and this is therefore not affected by the delay between the channel estimation and the packet transmission.

### 22.2.9 Channel estimation

All the adaptive modems above rely on the estimate of the frequency domain channel transfer function, both for equalisation of the received symbols at the receiver, as well as for the modem mode adaptation of the next transmitted OFDM symbol. Figure 22.16 shows the BER versus SNR curves for the 1% target BER modem, as it was presented above, if pilot symbol assisted channel estimation [12] is employed instead of the previously used delayed, but otherwise perfect, channel estimation.

Comparing the curves for perfect channel estimation and for the 64-pilot lowpass interpolation algorithm, it can be seen that the modem falls short of the target bit error rate of 1% for channel SNR values of up to 20 dB. More noise resilient channel estimation algorithms can improve the modem’s performance. If the passband width of the interpolation lowpass filter (see Section 20.3.1.1) is halved, which is indicated in Figure 22.16 as the reduced bandwidth (red. bw.) scenario, then the BER gap between the perfect and the pilot symbol assisted channel estimation narrows, and a BER of 1% is achieved at a SNR of 15 dB. Additionally, employing pairs of pilots with the above bandwidth-limited interpolation scheme further improves the modem’s performance, which results in BER values below 1% for SNR values above 5 dB. The averaging of the pilot pairs improves the noise resilience of the channel estimation, but introduces estimation errors for high SNR values. This can be observed in the



**Figure 22.16:** BER versus channel SNR performance for the 1% target BER adaptive 16 sub-band, 512 subcarrier OFDM modem employing pilot symbol assisted channel transfer function estimation over the channel of Figure 20.3

residual BER in the figure.

Having studied a range of different AOFDM modems, let us now embark on a system design study in the context of an adaptive interactive speech system.

## 22.3 Adaptive OFDM Speech System<sup>1</sup>

### 22.3.1 Introduction

In this section we introduce a bidirectional high-quality audio communications system, which will be used to highlight the systems aspects of adaptive OFDM transmissions over time dispersive channels. Specifically, the channel-coded adaptive transmission characteristics and a potential application for joint adaptation of modulation, channel coding and source coding are studied.

The basic principle of adaptive modulation is to react to the anticipated channel capacity for the next OFDM symbol transmission burst, by employing modulation modes of different robustness to channel impairments and of different data throughput. The trade-off between data throughput and integrity can be adapted to different system environments. For data transmission systems, which are not dependent on

<sup>1</sup>T. Keller, M. Münster, L. Hanzo: A Turbo-coded Burst-by-burst Adaptive Wideband Speech Transceiver, submitted to ©IEEE JSAC, 2000 [435]. Details concerning the source codec and its performance are discussed in the above publication.

a fixed data rate and do not require low transmission delays, variable-throughput adaptive schemes can be devised that operate efficiently with powerful error correction coders, such as long block length turbo codes [436]. Real-time audio or video communications employing source codecs, which allow variable bit rates, can also be used in conjunction with variable-rate adaptive schemes, but in this case block-based error correction coders cannot be readily employed.

Fixed-rate adaptive OFDM systems, which sacrifice a guaranteed BER performance for the sake of a fixed data throughput, are more readily integrated into interactive communications systems, and can coexist with long block based channel coders in real-time applications.

For these investigations, we propose a hybrid adaptive OFDM scheme, based on a multimode constant throughput algorithm, consisting of two adaptation loops: an inner constant throughput algorithm, having a bit rate consistent with the source and channel coders, and an outer mode switching control loop, which selects the target bit rate of the whole system from a set of distinct operating modes. These issues will become more explicit during our further discourse.

### 22.3.2 System overview

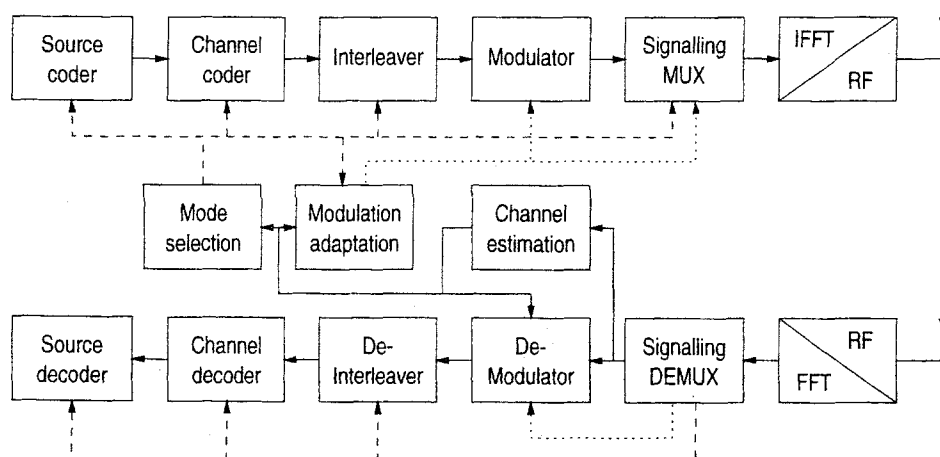


Figure 22.17: Schematic model of the multimode adaptive OFDM system

The structure of the studied adaptive OFDM modem is depicted schematically in Figure 22.17. The top half of the diagram is the transmitter chain, which consists of the source and channel coders, a channel interleaver decorrelating the channel's frequency domain fading, an adaptive OFDM modulator, a multiplexer adding signalling information to the transmit data, and an IFFT/RF OFDM block. The receiver, at the lower half of the schematic, consists of a RF/FFT OFDM receiver, a demultiplexer extracting the signalling information, an adaptive demodulator, a de-interleaver/channel decoder and the source decoder. The parameter adaptation linking the receiver and transmitter chain consists of a channel estimator, and the throughput mode selection as well as the modulation adaptation blocks.

The open-loop control structure of the adaptation algorithms can be seen in the

figure: the receiver's operation is controlled by the signalling information that is contained in the received OFDM symbol, while the channel quality information generated by the receiver is employed to determine the remote transmitter's matching parameter set by the modulation adaptation algorithms. The two distinct adaptation loops distinguished by the dotted and dashed lines are the inner and outer adaptation loops, respectively. The outer adaptation loop controls the overall throughput of the system, which is chosen from a finite set of predefined modes, so that a fixed delay decoding of the received OFDM data packets becomes possible. This outer loop controls the block length of the channel encoder and interleaver, and the target throughput of the inner adaptation loop. The operation of the adaptive modulator, controlled by the inner loop, is transparent to the rest of the system. The operation of the adaptation loops is described in more detail below.

### 22.3.2.1 System parameters

The transmission parameters have been adopted from the TDD mode of the UMTS system of Section 20.1.3, with a carrier frequency of 1.9 GHz, a time frame and time slot duration of 4.615 ms and 122  $\mu$ s, respectively. The sampling rate is assumed to be 3.78 MHz, leading to a 1024-subcarrier OFDM symbol with a cyclic extension of 64 samples in each time slot. For spectral shaping of the OFDM signal, there are a total of 206 virtual subcarriers at the bandwidth boundaries.

The 7 kHz bandwidth PictureTel audio codec<sup>2</sup> has been chosen for this system because of its good audio quality, robustness to packet dropping and adjustable bit rate. The channel encoder/interleaver combination is constituted by a convolutional turbo codec [433] employing block turbo interleavers in conjunction with a subsequent pseudorandom channel interleaver. The constituent half-rate recursive systematic convolutional (RSC) encoders are of constraint length 3, with octal generator polynomials of (7, 5) [403]. At the decoder, 8 iterations are performed, utilising the so-called maximum a posteriori (MAP) [434] algorithm and log-likelihood ratio soft inputs from the demodulator.

The channel model consists of a four-path COST 207 typical urban impulse response [410], where each impulse is subjected to independent Rayleigh fading with a normalised Doppler frequency of  $2.25 \cdot 10^{-6}$ , corresponding to a pedestrian scenario with a walking speed of 3 mph.

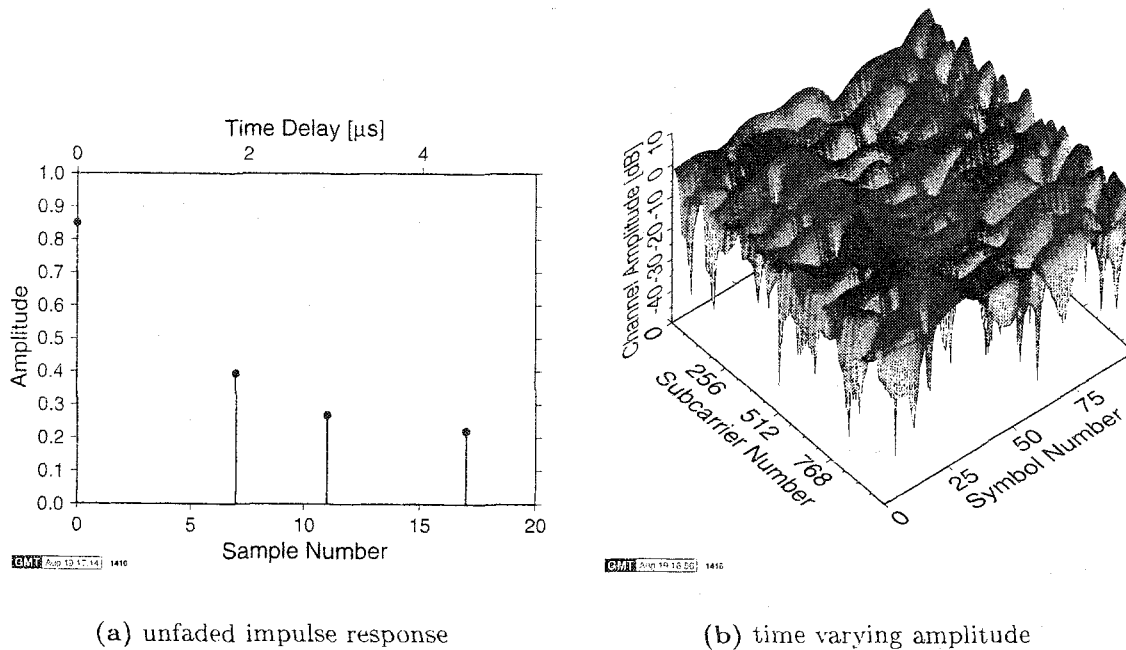
The unfaded impulse response and the time- and frequency-varying amplitude of the channel transfer function are depicted in Figure 22.18.

### 22.3.3 Constant throughput adaptive modulation

The constant throughput adaptive algorithm attempts to allocate a given number of bits for transmission in subcarriers exhibiting a low BER, while the use of high BER subcarriers is minimised. We employ the open-loop adaptive regime of Figure 22.1, basing the decision concerning the next transmitted OFDM symbol's modulation scheme allocation on the channel estimation gained at the reception of the most recently received OFDM symbol by the local station. Sub-band adaptive modulation

<sup>2</sup>see <http://www.picturetel.com>





**Figure 22.18:** Channel for PictureTel experiments: (a) unfaded channel impulse response  
(b) time-varying channel amplitude for 100 OFDM symbols

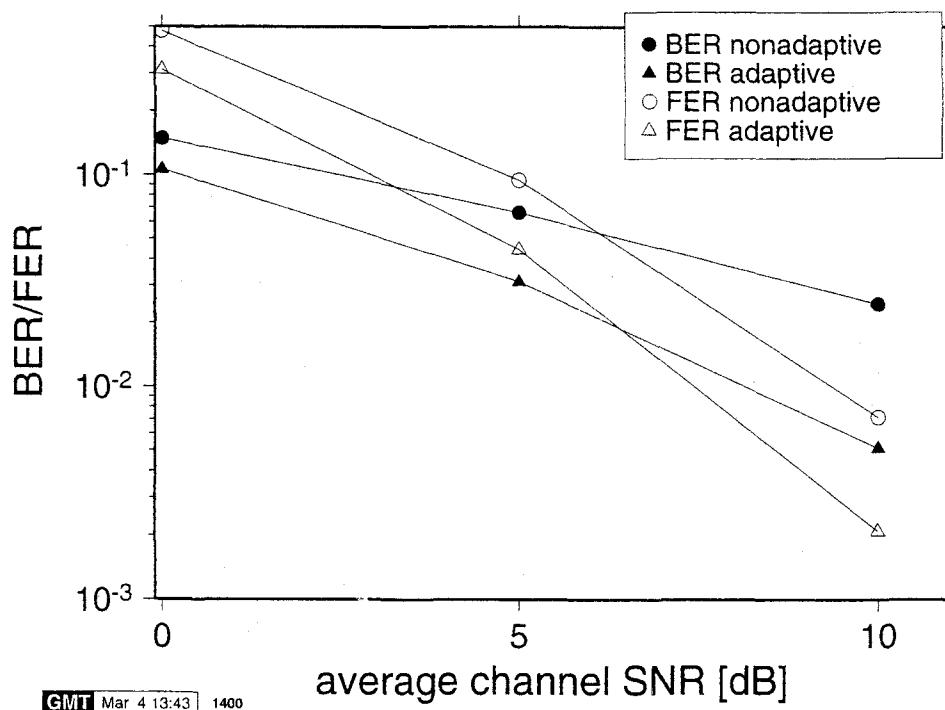
[437] – where the modulation scheme is adapted not on a subcarrier-by-subcarrier basis, but for sub-bands of adjacent subcarriers – is employed in order to simplify the signalling requirements. The adaptation algorithm was highlighted in Section 22.2.5. For these investigations we employed 32 sub-bands of 32 subcarriers in each OFDM symbol. Perfect channel estimation and signalling were used.

### 22.3.3.1 Constant-rate BER performance

Figure 22.19 characterises the fixed-throughput adaptive modulation scheme's performance under the channel conditions characterised above, for a block length of 578 coded bits. As a comparison, the BER curve of a fixed BPSK modem transmitting the same number of bits in the same channel, employing 578 out of 1024 subcarriers, is also depicted. The number of useful audio bits per OFDM symbol was based on a 200-bit target data throughput, which corresponds to a 10 kbps data rate, padded with 89 bits, which can contain a checksum for error detection and high-level signalling information. Furthermore, half-rate channel coding was used.

The BER plotted in the figure is the hard decision based bit error rate at the receiver before channel decoding. It can be seen that the adaptive modulation scheme yields a significantly improved performance, which is also reflected in the frame error rate (FER). This FER approximates the probability of a decoded block containing errors, in which case it is unusable for the audio source decoder and hence it is dropped. This error event can be detected by using the checksum in the OFDM data symbol.

As an example, the modulation mode allocation for the 578 data bit adaptive

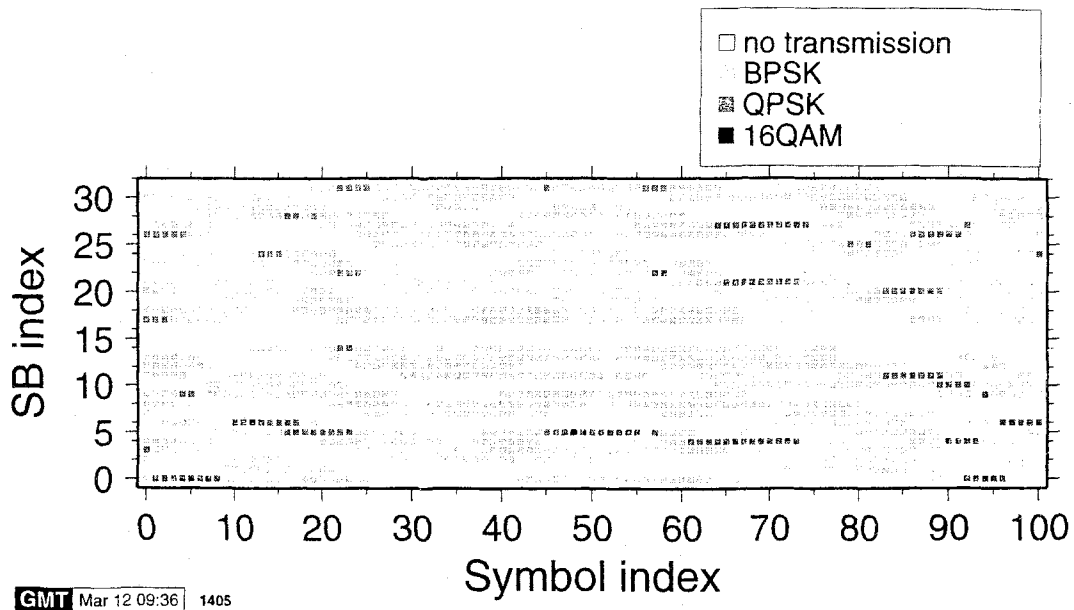


**Figure 22.19:** BER and FER performance for the fixed-throughput adaptive and non-adaptive OFDM modems in the fading time dispersive channel of Section 20.1.3 for a block length of 578 coded bits

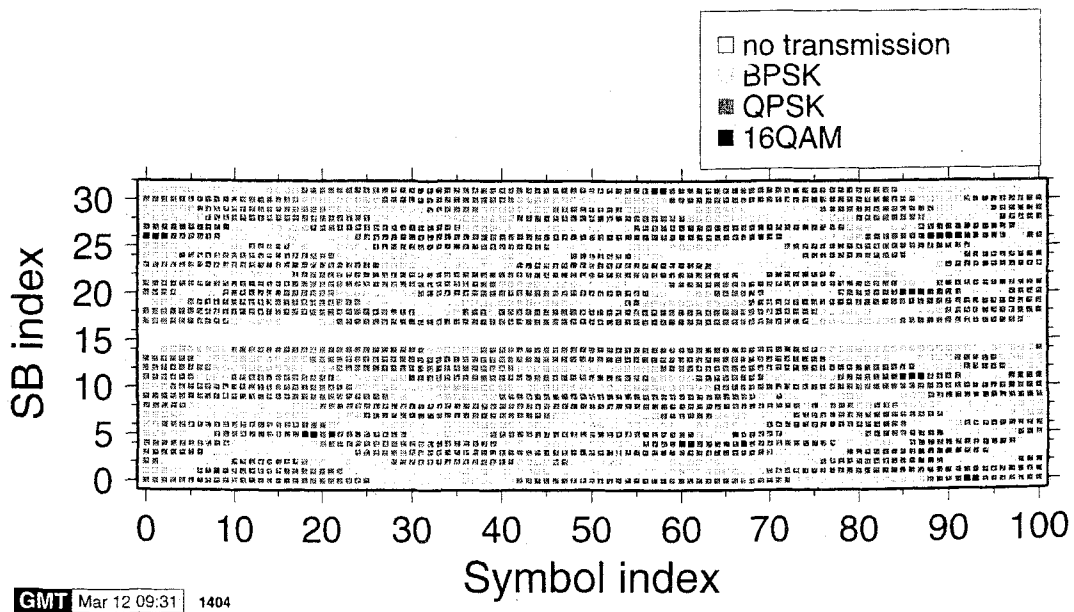
modem at an average channel SNR of 5 dB is given in Figure 22.20(a) for 100 consecutive OFDM symbols. The unused sub-bands with indexes 15 and 16 contain the virtual carriers, and therefore do not transmit any data. It can be seen that the constant throughput adaptation algorithm of Section 22.2.5 allocates data to the higher quality subcarriers on a symbol-by-symbol basis, while keeping the total number of bits per OFDM symbol constant. As a comparison, Figure 22.20(b) shows the equivalent overview of the modulation modes employed for a fixed bit rate of 1458 bits per OFDM symbol. It can be seen that in order to meet this increased throughput target, hardly any sub-bands are in “no transmission” mode, and overall higher-order modulation schemes have to be employed.

### 22.3.4 Multimode adaptation

While the fixed-throughput adaptive algorithm described above copes with the frequency domain fading of the channel, there is a medium-term variation of the overall channel capacity due to time domain channel quality fluctuations as indicated in Figure 22.18(b). While it is not straightforward to employ powerful block-based channel coding schemes, such as turbo coding, in variable-throughput adaptive OFDM schemes for real-time applications like voice or video telephony, a multimode adaptive system can be designed that allows us to switch between a set of different source and channel codecs as well as transmission parameters, depending on the overall channel quality. We have investigated the use of the estimated overall BER at the output



(a) 578 data bits per OFDM symbol



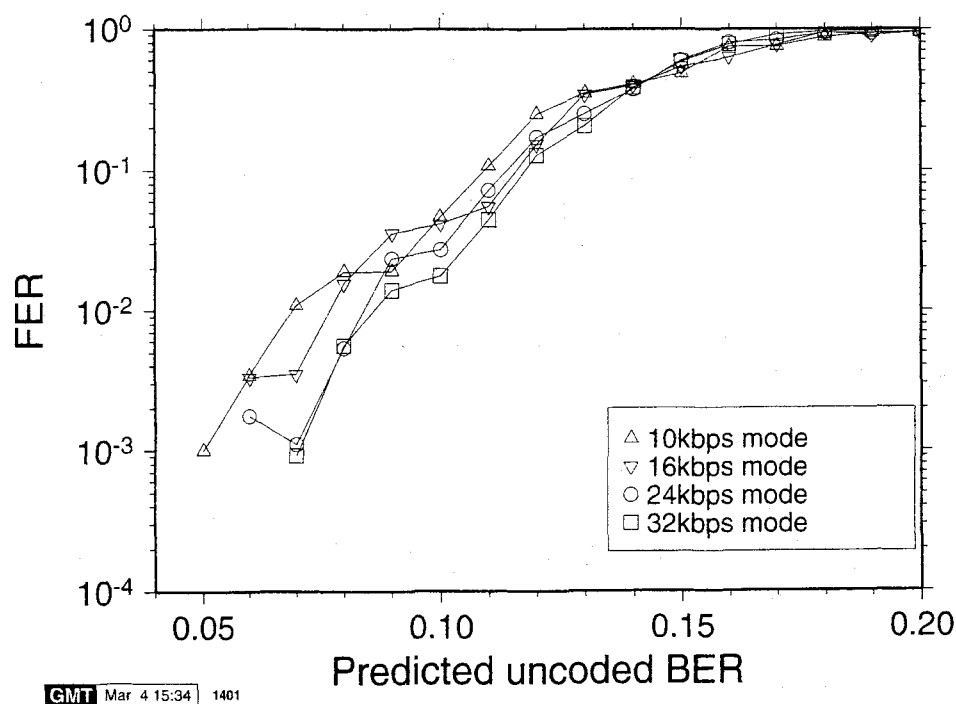
(b) 1458 data bits per OFDM symbol

**Figure 22.20:** Overview of modulation mode allocation for fixed-throughput adaptive modems over the fading time dispersive channel of Figure 22.18(b) at 5 dB average channel SNR

of the receiver, which is the sum of all the  $e(j, s_j)$  quantities of Equation 22.1 after adaptation. On the basis of this expected bit error rate at the input of the channel decoder, the probability of a frame error (FER) must be estimated and compared with the estimated FER of the other modem modes. Then the mode having the highest throughput exhibiting an estimated FER of less than  $10^{-6}$  – or alternatively the mode exhibiting the lowest FER – is selected and the source encoder, the channel encoder and the adaptive modem are set up accordingly.

We have defined four different operating modes, which correspond to unprotected audio data rates of 10, 16, 24, and 32 kbps at the source codec's interface. With half-rate channel coding and allowing for checksum and signalling overhead, the number of transmitted coded bits per OFDM symbol was 578, 722, 1058 and 1458 for the four modes, respectively.

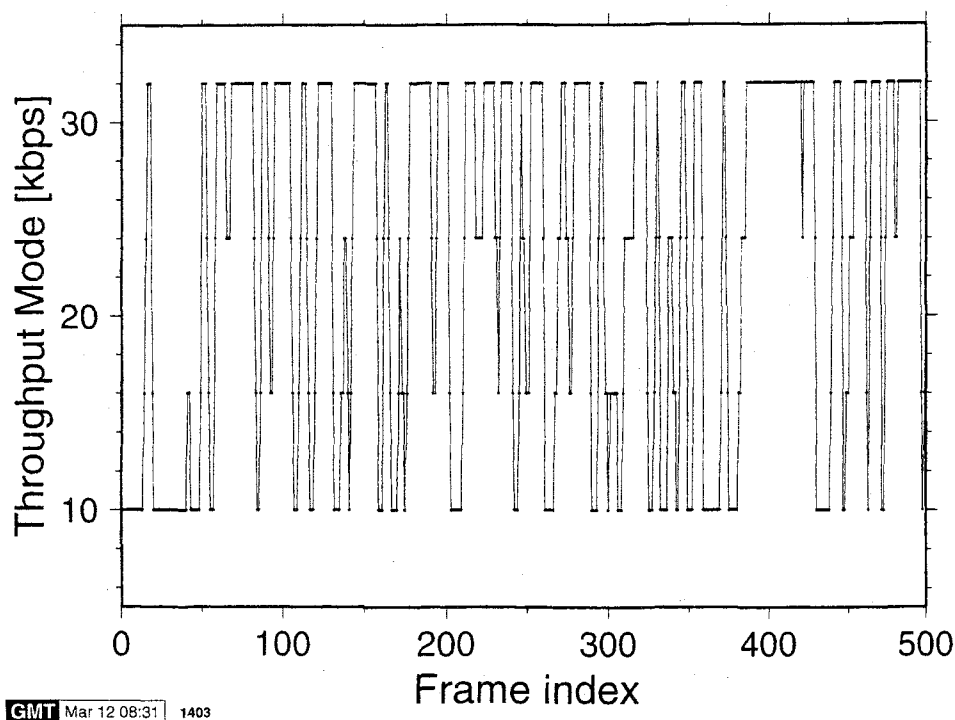
#### 22.3.4.1 Mode switching



**Figure 22.21:** Frame error rate versus the predicted unprotected BER for 10 kbps, 16 kbps, 24 kbps and 32 kbps modes

Figure 22.21 shows the *observed* FER for all four modes versus the unprotected BER that was *predicted* at the transmitter. The predicted unprotected BER was discretised into intervals of 1%, and the channel-coded FER was evaluated over these BER intervals. It can be seen from the figure that for estimated protected BER values below 5% no frame errors were observed for any of the modes. For higher estimated unprotected BER values, the higher throughput modes exhibited a lower FER than the lower throughput modes, which was consistent with the turbo coder's performance increase for longer block lengths. A FER of 1% was observed for a 7%

predicted unprotected error rate for the 10 kbps mode, and BER values of 8% to 9% were allowed for the longer blocks, whilst maintaining a FER of less than 1%



**Figure 22.22:** Mode switching pattern at 5 dB channel SNR over the channel of Figure 22.18(b)

For this experiment, we assumed the best-case scenario of using the actual measured FER statistics of Figure 22.21 for the mode switching algorithm rather than estimating the FER on the basis of the estimated uncoded BER. In this case, the previously observed FER corresponding to the predicted overall BER values for the different modes were compared, and the mode having the lowest FER was chosen for transmission. The mode switching sequence for the first 500 OFDM symbols at 5 dB channel SNR over the channel of Figure 22.18(b) is depicted in Figure 22.22. It can be seen that in this segment of the sequence 32 kbps transmission is the most frequently employed mode, followed by the 10 kbps mode. The intermediate modes are mostly transitory, as the improving or deteriorating channel conditions makes it necessary to switch between the 10 kbps and 32 kbps modes. This behaviour is consistent with Table 22.2, for the Switch-I scheme.

### 22.3.5 Simulation results

The comparison between the different adaptive schemes will be based on a channel SNR of 5 dB over the channel of Figure 22.18(b), since the audio codec's performance is unacceptable for SNR values around 0 dB, and as the adaptive modulation is most effective for channel SNR values below 10 dB.

Scheme	FER [%]	Rate-10kbps [%]	Rate-16kbps [%]	Rate-24kbps [%]	Rate-32kbps [%]
Fixed-10kbps	4.45	95.55	0.0	0.0	0.0
Fixed-16kbps	5.58	0.0	94.42	0.0	0.0
Fixed-24kbps	10.28	0.0	0.0	89.72	0.0
Fixed-32kbps	18.65	0.0	0.0	0.0	81.35
Switch-I	4.44	21.87	13.90	11.59	48.20
Switch-II	5.58	0.0	34.63	11.59	48.20

**Table 22.2:** FER and relative usage of different bit rates in the fixed bit rate schemes and the variable schemes Switch-I and Switch-II (successfully transmitted frames) for a channel SNR of 5 dB over the channel of Figure 22.18(b)

### 22.3.5.1 Frame error results

The audio experiments [435] have shown that the audio quality is acceptable for frame dropping rates of about 5%, and that the perceived audio quality increases with increasing throughput. Table 22.2 gives an overview of the frame error rates and mode switching statistics of the system for a channel SNR of 5 dB over the channel of Figure 22.18(b). It can be seen that for the fixed modes the FER increases with the throughput, from 4.45% in the 10 kbps mode up to 18.65% for the 32 kbps mode. This is because the turbo codec performance improves for longer interleavers, the OFDM symbol had to be loaded with more bits, hence there was a higher unprotected BER. The time-variant bit rate mode-switching schemes, referred to as Switch-I and Switch-II for the four- and three-mode switching regimes used, deliver frame dropping rates of 4.44% and 5.58%, respectively. Both these FER values are acceptable for the audio transmission. It can be seen that upon incorporating the 10 kbps mode in the switching regime Switch-I of Table 22.2 the overall FER is lowered only by an insignificant amount, while the associated average throughput is reduced considerably.

### 22.3.5.2 Audio segmental SNR

Figure 22.23 displays the cumulative density function (CDF) of the segmental SNR (SEGSNR) [403] obtained from the reconstructed signal of an audio test sound for all the modes of Table 22.2 discussed above at a channel SNR of 5 dB over the channel of Figure 22.18(b).

Focusing our attention on the figure, we can draw a whole range of interesting conclusions. As expected, for any given SEGSNR it is desirable to maintain as low a proportion as possible of the audio frames' SEGSNRs below a given abscissa value. Hence we conclude that the best SEGSNR CDF was attributable to the Switch-II scheme, while the worst performance was observed for the fixed 10 kbps scheme. Above a SEGSNR of 15 dB the CDFs of the fixed 16, 24 and 32 kbps modes follow our expectations. Viewing matters from a different perspective, the Switch-II scheme exhibits a SEGSNR of less than 20 dB with a probability of 0.8, compared to 0.95 for the fixed 10 kbps scheme.

Before concluding we also note that the CDFs do not have a smoothly tapered tail,

since for the erroneous audio frames a SEGSNR of 0 dB was registered. This results in the step-function-like behaviour associated with the discontinuities corresponding to the FER values in the FER column of Table 22.2.

## 22.4 Pre-equalisation

We have seen above how the receiver's estimate of the channel transfer function can be employed by the transmitter in order to dramatically improve the performance of an OFDM system by adapting the subcarrier modulation modes to the channel conditions. For sub-channels exhibiting a low signal-to-noise ratio, robust modulation modes were used, while for subcarriers having a high SNR, high throughput multi-level modulation modes can be employed. An alternative approach to combating the frequency selective channel behaviour is to apply pre-equalisation to the OFDM symbol prior to transmission on the basis of the anticipated channel transfer function. The optimum scheme for the power allocation to the subcarriers is Shannon's water-pouring solution [438]. We will investigate a range of related topics in this section.

### 22.4.1 Motivation

As discussed above, the received data symbol  $R_n$  of subcarrier  $n$  over a stationary time dispersive channel can be characterised by:

$$R_n = S_n \cdot H_n + n_n,$$

where  $S_n$  is the transmitted data symbol,  $H_n$  is the channel transfer function of subcarrier  $n$ , and  $n_n$  is a noise sample.

The frequency domain equalisation at the receiver, which is necessary for non-differential detection of the data symbols, corrects the phase and amplitude of the received data symbols using the estimate of the channel transfer function  $\hat{H}_n$  as follows:

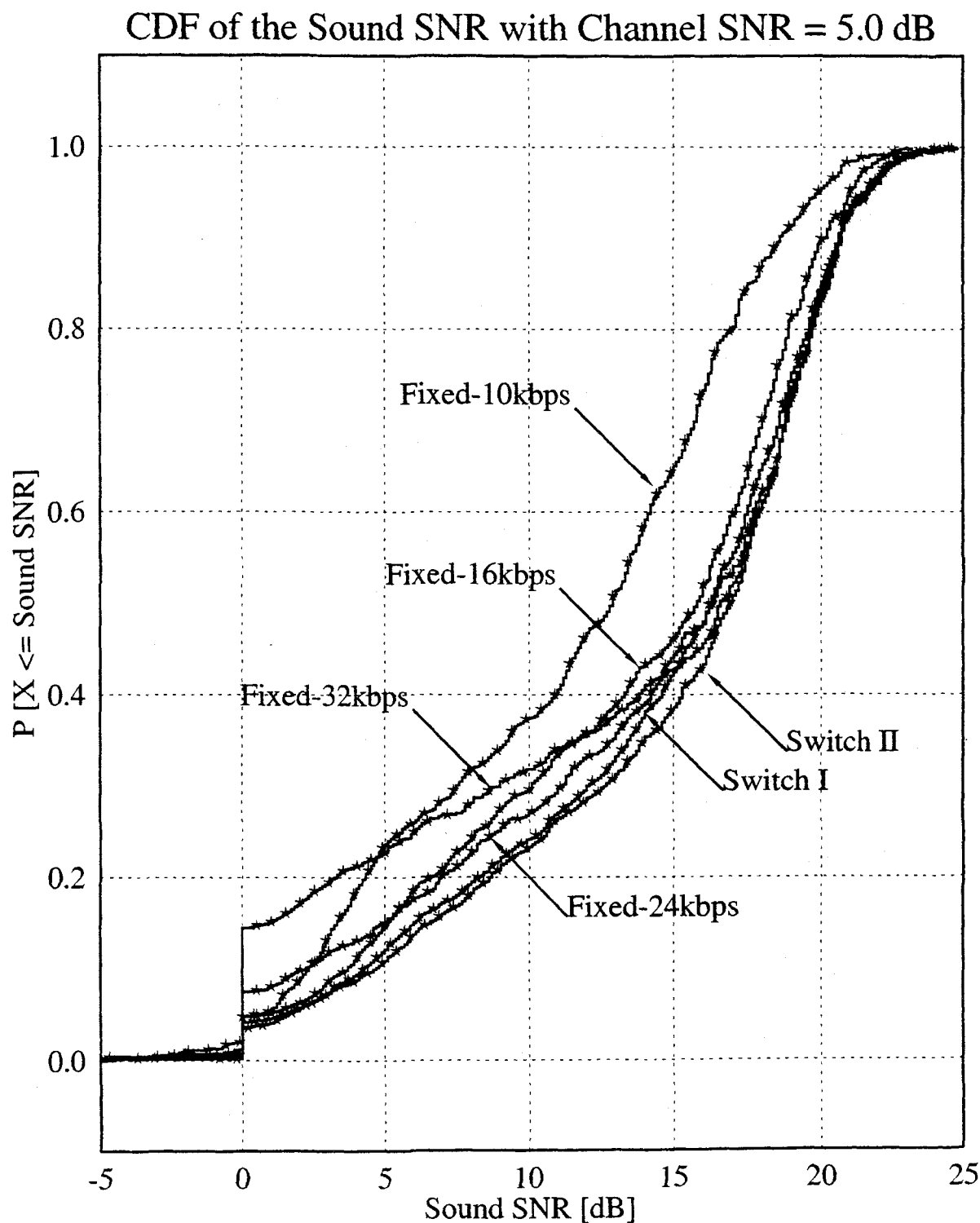
$$R'_n = R_n / \hat{H}_n = S_n \cdot H_n / \hat{H}_n + n_n / \hat{H}_n.$$

If the estimate  $\hat{H}_n$  is accurate, this operation defades the constellation points before decision. However, upon defading, the noise sample  $n_n$  is amplified by the same amount as the signal, therefore preserving the SNR of the received sample.

Pre-equalisation for the OFDM modem operates by scaling the data symbol of subcarrier  $n$ ,  $S_n$ , by a predistortion function  $E_n$ , computed from the inverse of the anticipated channel transfer function, prior to transmission. At the receiver, no equalisation is performed, hence the received symbols can be expressed as:

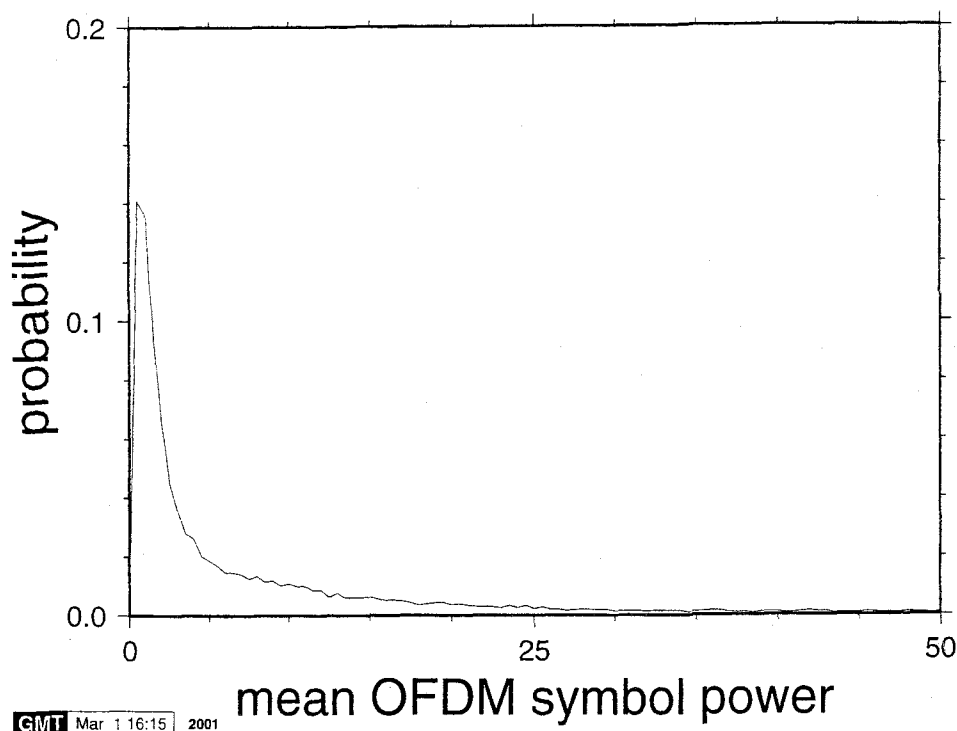
$$R_n = S_n \cdot E_n \cdot H_n + n_n.$$

Since no equalisation is performed, there is no noise amplification at the receiver. Similarly to the adaptive modulation techniques illustrated above, pre-equalisation is only applicable to a duplex link, since the transmitted signal is adapted to the specific channel conditions perceived by the receiver. As in other adaptive schemes, the transmitter needs an estimate of the current frequency domain channel transfer



**Figure 22.23:** Typical CDF of the segmental SNR of a reconstructed audio signal transmitted over the fading time dispersive channel of Section 20.1.3 at a channel SNR of 5 dB (from [435])



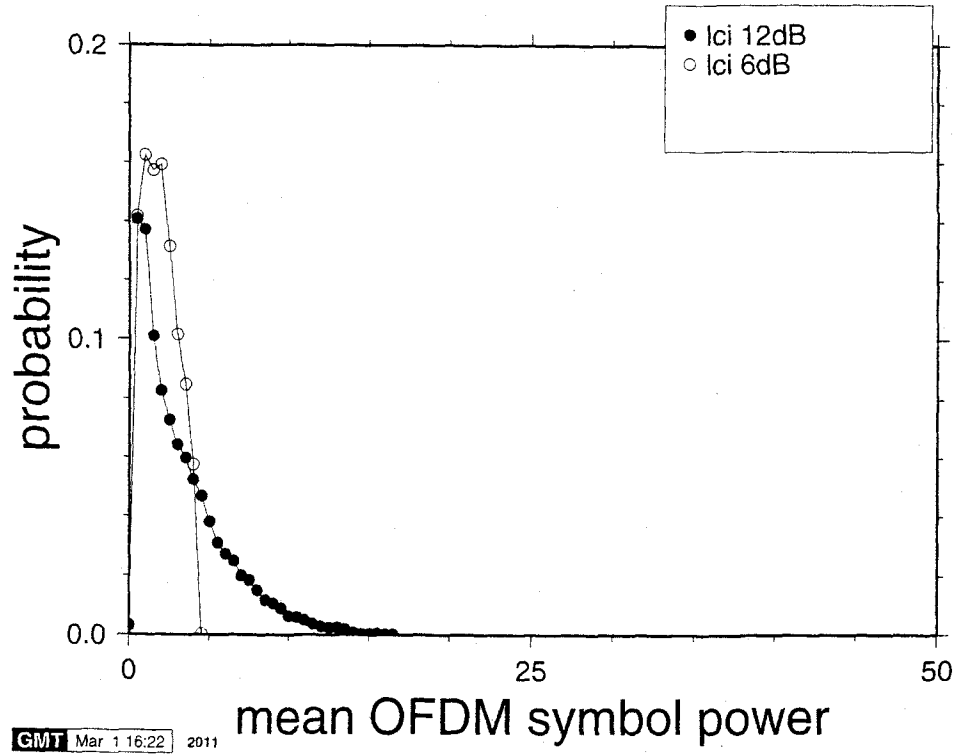


**Figure 22.24:** OFDM symbol energy histogram for 512-subcarrier 16-QAM with full channel inversion over the short WATM channel of Figure 20.3

function, which can be obtained from the received signal in the reverse link, as seen in Figure 22.1.

The simplest choice of the pre-equalisation transfer function  $E_n$  is the inverse of the estimated frequency domain channel transfer function,  $E_n = 1/\hat{H}_n$ . If the estimation of the channel transfer function is accurate, then perfect channel inversion would result in an AWGN-like channel perceived at the receiver, since the anticipated time and frequency dependent behaviour of the channel is precompensated at the transmitter. The BER performance of such a system, accordingly, is identical to that of the equivalent modem in an AWGN channel with respect to the received signal power.

Since the pre-equalisation algorithm amplifies the power in each subcarrier by the corresponding estimate of the channel transfer function, the transmitter's output power fluctuates in an inverse fashion with respect to time variant channel. The fades in the frequency domain channel transfer function can be deep, hence the transmit power in the corresponding subcarriers may be high. Figure 22.24 shows the histogram of the total OFDM symbol energy at the transmitter's output for the short WLAN channel of Figure 20.3 in conjunction with "full channel inversion", normalised to the fixed average output energy. It can be seen that the OFDM symbol energy fluctuates widely, with observed peak values in excess of 55. The long-term mean symbol energy was measured to be 22.9, which corresponds to an average output power increase of 13.6 dB. This naturally would impose unacceptable constraints on the required perfectly linear dynamic range of the power amplifier. Hence in practice only limited



**Figure 22.25:** OFDM symbol energy histogram for 512-subcarrier 16-QAM transmission with limited dynamic range channel inversion (lci) over the WATM channel of Figure 20.3.

dynamic scenarios can be considered.

Explicitly, in order to limit the associated transmit power fluctuations, the dynamic range of the pre-equalisation algorithm can be limited to a value  $l$ , so that the following relations apply:

$$E_n = a_n \cdot e^{-j\phi_n}, \text{ with} \quad (22.3)$$

$$\phi_n = \angle \hat{H}_n \text{ and} \quad (22.4)$$

$$a_n = \begin{cases} |\hat{H}_n| & \text{for } |\hat{H}_n| \leq l \\ l & \text{otherwise.} \end{cases} \quad (22.5)$$

Limiting the values of  $E_n$  to the value of  $l$  does not affect the phase of the channel pre-equalisation. Depending on the modulation mode employed for transmission, reception of the symbols affected by the amplitude limitation is still possible, perhaps for phase shift keying. Multilevel modulation modes exploiting the received symbol's amplitude will be affected by the imperfect pre-equalisation. The associated mean OFDM symbol power histogram is shown in Figure 22.25. Given - for example - a maximum allowed amplification factor of 6 and 12 dB, the normalised transmitted OFDM symbol power in the figure should be limited to 4 and 16, respectively. However, in practice even higher values may be observed, which is due to the OFDM symbol's energy fluctuation as a function of the specific data sequence, if multilevel modulation

modes are used. In order to circumvent the above peak-to-mean envelope fluctuation problems, it is in practical terms more attractive to combine pre-equalisation with sub-band blocking, which is the topic of the next section.

### 22.4.2 Pre-equalisation with sub-band blocking

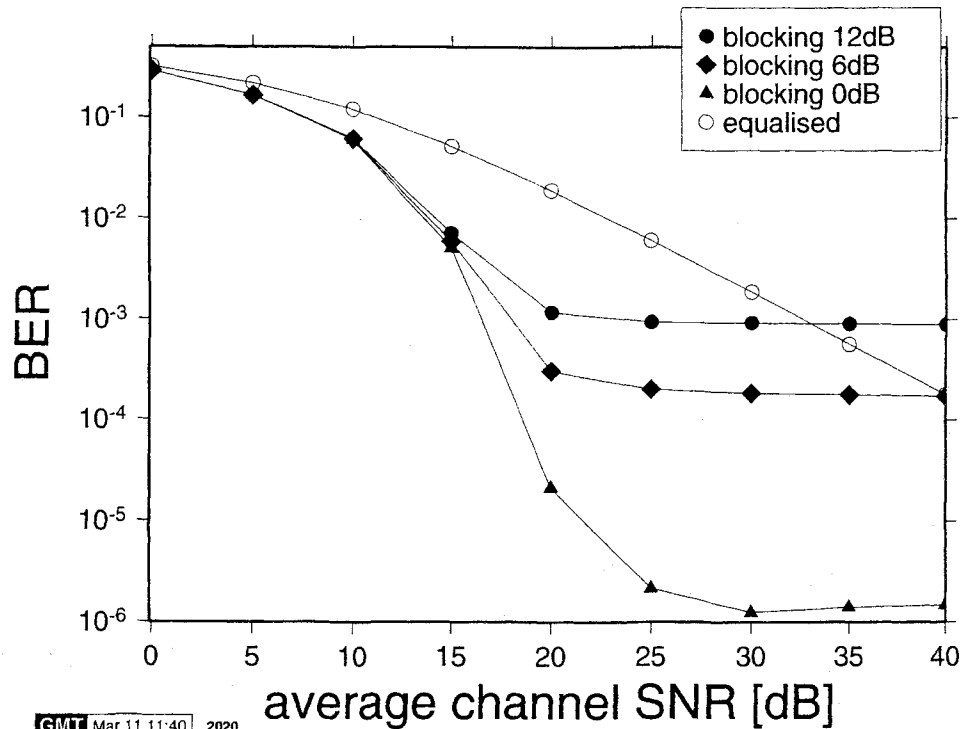
Limiting the maximal amplification of the subcarriers leads to a reduced BER performance compared to the full channel inversion, but the system performance can be improved by identifying the subcarriers that cannot be fully pre-equalised and by disabling subsequent transmission in these subcarriers. This “blocking” of the transmission in certain subcarriers can be seen as adaptive modulation with two modulation modes, which introduces the problem of modulation mode signalling. As has been discussed in the context of Figure 22.1 for the adaptive modulation modems above, this signalling task can be solved in different ways, namely by blind detection of blocked subcarriers, or by transmitting explicit signalling information contained in the data block. We have seen above that employing sub-band adaptivity - rather than subcarrier-by-subcarrier adaptivity - simplifies both the modem mode detection as well as the mode signalling, at the expense of a lower system throughput. In order to keep the system’s complexity low and to allow for simple modem mode signalling or blind detection, we will assume a 16 subband adaptive scheme here.

Analogously to the adaptive modulation schemes above, the transmitter decides for all subcarriers in each sub-band, whether to transmit data or not. If pre-equalisation is possible under the power constraints, then the subcarriers are modulated with the pre-equalised data symbols. The information on whether or not a sub-band is used for transmission, this is signalled to the receiver.

Since no attempt is made to transmit in the sub-bands that cannot be power-efficiently pre-equalised, the power conserved in the blank subcarriers can be used for “boosting” the data-bearing sub-bands. This scheme allows for a more flexible pre-equalisation algorithm than the fixed threshold method described above; here is a summary:

- (1) Calculate the necessary transmit power  $p_n$  for each sub-band  $n$ , assuming perfect pre-equalisation.
- (2) Sort sub-bands according to their required transmit power  $p_n$ .
- (3) Select sub-band  $n$  with the lowest power  $p_n$ , and add  $p_n$  to the total transmit power. Repeat this procedure with the next lowest power, until no further sub-bands can be added without the total power  $\sum p_j$  exceeding power limit  $l$ .

Figure 22.26 depicts the 16-QAM BER performance over the short WATM channel of Figure 20.3. The BER floor stems from the channel’s time-variant nature, since there is a delay between the channel estimation instant and the instant of transmission. The average throughput figures for the 6 dB and 12 dB symbol energy limits are 3.54 and 3.92 bits per data symbol, respectively. It can be noted that the BER floor is lower for  $l = 6$  dB than for  $l = 12$  dB. This is because the effects of the channel variation due to the delay between the instants of channel estimation and reception in the faded subcarriers on the equalisation function are more dramatic, than in the higher

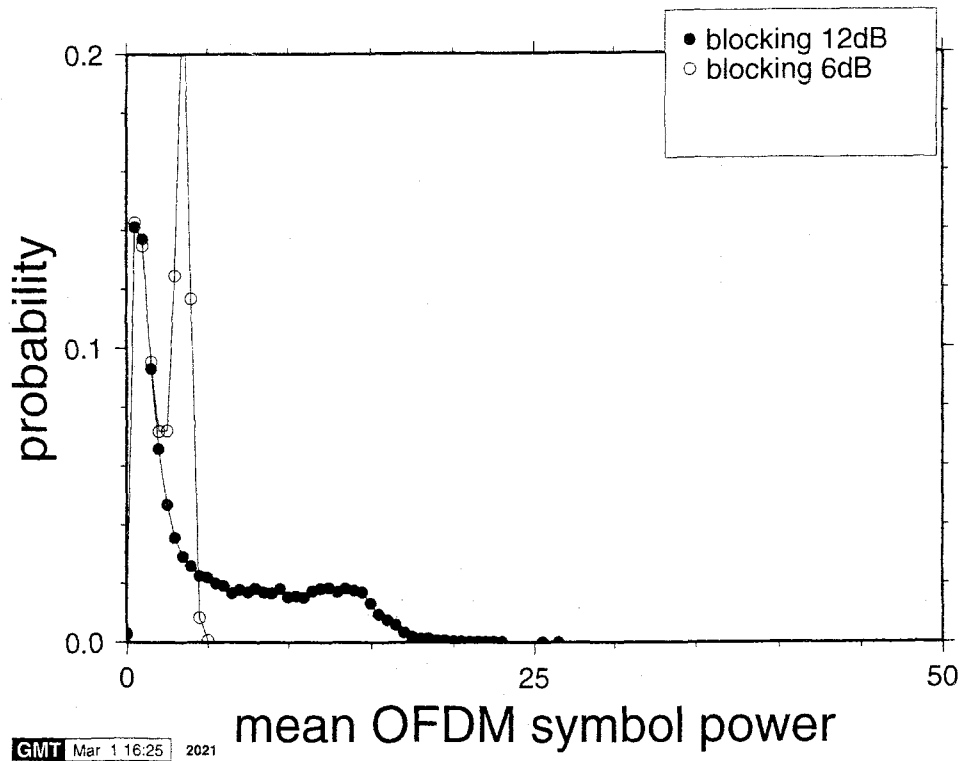


**Figure 22.26:** BER performance of the 512-subcarrier 16-QAM OFDM modem over the fading short WATM channel of Figure 20.3 employing 16 sub-band pre-equalisation with blocking and a delay of 1 time slot between the instants of perfect channel estimation and reception. Note that the transmit power is not shown in this figure

quality subcarriers. The lower the total symbol energy limit  $l$ , the smaller the number of low quality subcarriers used for transmission. If the symbol energy is limited to 0 dB, then the BER floor drops to  $1.5 \cdot 10^{-6}$  at the expense of the throughput, which attains 2.5 BPS. Figure 22.27 depicts the mean OFDM symbol energy histogram for this scenario. It can be seen that, compared with the limited channel inversion scheme of Figure 22.25, the allowable symbol energy is more efficiently allocated, with a higher probability of high energy OFDM symbols. This is the result of the flexible reallocation of energy from blocked sub-bands, instead of limiting the output power on a subcarrier-by-subcarrier basis.

### 22.4.3 Adaptive modulation with spectral predistortion

The pre-equalisation algorithms discussed above invert the channel's anticipated transfer function, in order to transform the resulting channel into a Gaussian-like non-fading channel, whose SNR is dependent only on the path loss. Sub-band blocking has been introduced above, in order to limit the transmitter's output power, while maintaining the near-constant SNR across the used subcarriers. The pre-equalisation algorithms discussed above do not cancel out the channel's path loss, but rely on the receiver's gain control algorithm to automatically account for the channel's average



**Figure 22.27:** OFDM symbol energy histogram for 512 subcarrier, 16 sub-band pre-equalisation with blocking over the short WATM channel of Figure 20.3 using 16-QAM. The corresponding BER curves are given in Figure 22.26

path loss.

We have already that maintaining Gaussian channel characteristics is not the most efficient way of exploiting the channel's time variant capacity. If maintaining a constant data throughput is not required by the rest of the communications system, then a fixed BER scheme in conjunction with error correction coding can assist in maximising the system's throughput. The results presented for the target BER adaptive modulation scheme in Figure 22.14(b) showed that, for the particular turbo coding scheme used, an uncoded BER of 1% resulted in error-free channel coded data transmission, and that for an uncoded target BER of 10% the turbo decoded data BER was below  $10^{-5}$ . We have seen that it is impossible to exactly reach the anticipated uncoded target BER with the adaptive modulation algorithm, since the adaptation algorithm operates in discrete steps between modulation modes.

Combining the target BER adaptive modulation scheme and spectral pre-distortion allows the transmitter to react to the channel's time and frequency variant nature, in order to fine-tune the behaviour of the adaptive modem in fading channels. It also allows the transmitter to invest the energy that is not used in "no transmission" sub-bands into the other sub-bands without affecting the equalisation at the receiver.

The combined algorithm for adaptive modulation with spectral pre-distortion described here does not intend to invert the channel's transfer function across the OFDM

target BER	$10^{-4}$	1%	10%
SNR(BPSK)[dB]	8.4	4.33	-0.85
SNR(QPSK)[dB]	11.42	7.34	2.16
SNR(16QAM)[dB]	18.23	13.91	7.91

**Table 22.3:** Required target SNR levels for 1% and 10% target BER for the different modulation schemes over an AWGN channel

symbol's range of subcarriers, it is therefore not a pure pre-equalisation algorithm. Instead, the aim is to transmit a sub-band's data symbols at a power level which ensures a given target SNR at the receiver, that is constant for all subcarriers in the sub-band, which in turn results in the required BER. Clearly, the receiver has to anticipate the different relative power levels for the different modulation modes, so that error-free demodulation of the multilevel modulation modes employed can be ensured.

The joint adaptation algorithm requires the estimates of the noise floor level at the receiver as well as the channel transfer function, which includes the path loss. On the basis of these values, the necessary amplitude of  $E_n$  required to transmit a data symbol over the subcarrier  $n$  for a given received SNR of  $\gamma_n$  can be calculated as follows:

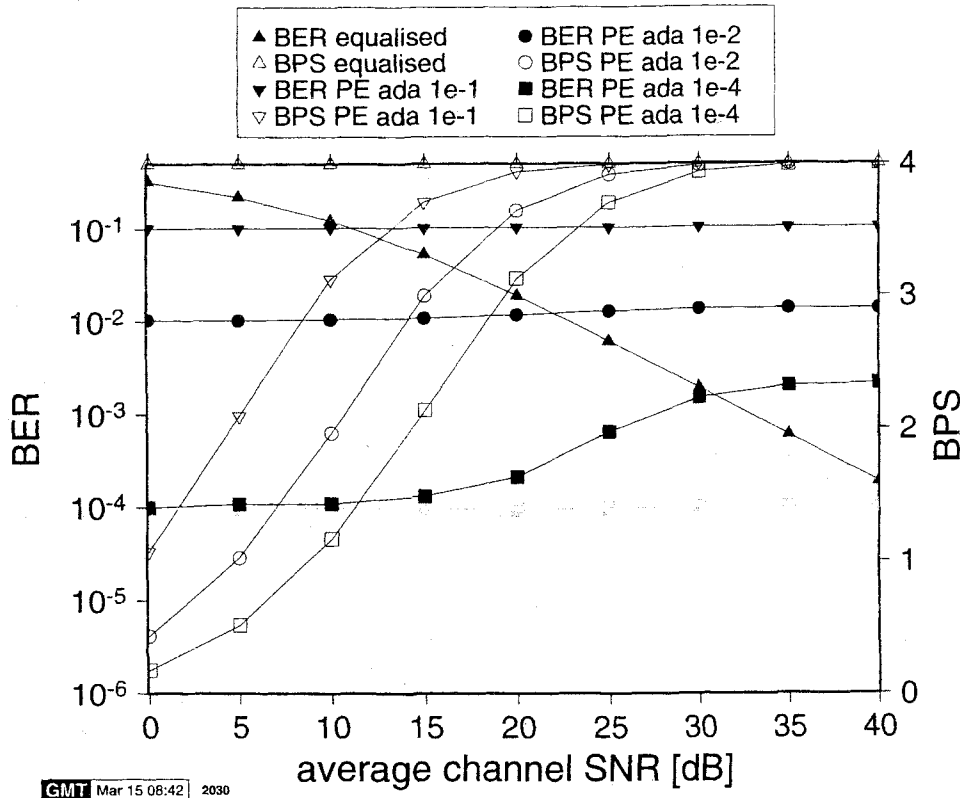
$$|E_n| = \frac{\sqrt{N_0 \cdot \gamma_n}}{|\hat{H}_n|},$$

where  $N_0$  is the noise floor at the receiver. The phase of  $E_n$  is used for the pre-equalisation, and hence:

$$\angle E_n = -\angle \hat{H}_n.$$

The target SNR, of subcarrier  $n$ ,  $\gamma_n$ , is dependent on the modulation mode that is signalled over the subcarrier, and determines the system's target BER. We have identified three sets of target SNR values for the modulation modes, with uncoded target BER values of 1% and 10% for use in conjunction with channel coders, as well as  $10^{-4}$  for transmission without channel coding. Table 22.3 gives an overview of these levels, which have been read from the BER performance curves of the different modulation modes in a Gaussian channel.

Figure 22.28 shows the performance of the joint predistortion and adaptive modulation algorithm over the fading time-dispersive short WATM channel of Figure 20.3 for the set of different target BER values of Table 22.3, as well as the comparison curves of the perfectly equalised 16-QAM modem under the same channel conditions. It can be seen that the BER achieved by the system is close to the BER targets. Specifically, for a target BER of 10%, no perceptible deviation from the target has been recorded, while for the lower BER targets the deviations increase for higher channel SNRs. For a target BER of 1%, the highest measured deviation is at the SNR of 40 dB, where the recorded BER is 1.36%. For the target BER of  $10^{-4}$ , the BER deviation is small at 0 dB SNR, but at an SNR of 40 dB the experimental BER is  $2.2 \cdot 10^{-3}$ . This increase of the BER with increasing SNR is due to the rapid channel



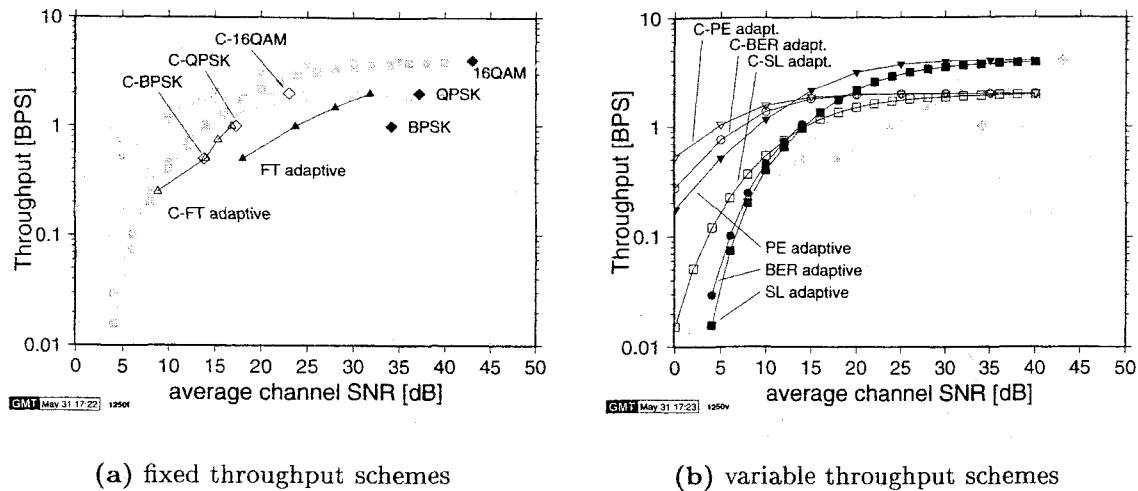
**Figure 22.28:** BER performance and BPS throughput of the 512 subcarrier, 16 sub-band adaptive OFDM modem with spectral pre-distortion over the Rayleigh fading time dispersive short WATM channel of Figure 20.3, and that of the perfectly equalised 16-QAM modem. The light grey BER curve gives the performance of the adaptive modem for a target BER of  $10^{-4}$  with no delay between channel estimation and transmission, while the other results assume 1 time slot delay between uplink and downlink.

variations in the deeply faded subcarriers, which are increasingly used at higher SNR values. The light grey curve in the figure denotes the system's performance if no delay is present between the channel estimation and the transmission. In this case, the simulated BER shows only very little deviation from the target BER value. This is consistent with the behaviour of the full channel inversion pre-equalising modem.

## 22.5 Comparison of the Adaptive Techniques

Figure 22.29 compares the different adaptive modulation schemes discussed in this chapter. The comparison graph is split into two sets of curves, depicting the achievable data throughput for a data BER of  $10^{-4}$  highlighted for the fixed throughput systems in Figure 22.29(a), and for the time-variant -throughput systems in Figure 22.29(b).

The fixed throughput systems, highlighted in black in Figure 22.29(a), comprise the non-adaptive BPSK, QPSK and 16-QAM modems, as well as the fixed-throughput adaptive scheme, both for coded and uncoded applications. The non-



**Figure 22.29:** BPS throughput versus average channel SNR for non-adaptive and adaptive modulation as well as for pre-equalised adaptive techniques, for a data bit error rate of  $10^{-4}$ . Note that for the coded schemes the achieved BER values are lower than  $10^{-4}$ . (a) Fixed-throughput systems: coded (C-) and uncoded BPSK, QPSK, 16-QAM, and fixed-throughput (FT) adaptive modulation. (b) Variable-throughput systems: coded (C-) and uncoded switching level adaptive (SL), target BER adaptive (BER) and pre-equalised adaptive (PE) systems. Note that the separately plotted variable-throughput graph also shows the light grey benchmark curves of the complementary fixed-rate schemes and vice versa

adaptive modems' performance is marked on the graph as diamonds, and it can be seen that the uncoded fixed schemes require the highest channel SNR of all examined transmission methods to achieve a data BER of  $10^{-4}$ . Channel coding employing the advocated turbo coding schemes dramatically improves the SNR requirements, at the expense of half the data throughput. The uncoded fixed-throughput (FT) adaptive scheme, marked by filled triangles, yields consistently worse data throughput than the coded (C-) fixed modulation schemes C-BPSK, C-QPSK and C-16QAM, with its throughput being about half the coded fixed scheme's at the same SNR values. The coded FT (C-FT) adaptive system, however, delivers very similar throughput to the C-BPSK and C-QPSK transmission, and can deliver a BER of  $10^{-4}$  for SNR values down to about 9 dB.

The variable throughput schemes, highlighted in Figure 22.29(b), outperform the comparable fixed throughput algorithms. For high SNR values, all uncoded schemes' performance curves converge to a throughput of 4 bits/symbol, which is equivalent to 16-QAM transmission. The coded schemes reach a maximal throughput of 2 bits/symbol. Of the uncoded schemes, the "data" switching level (SL) and target BER adaptive modems deliver a very similar BPS performance, with the target-BER scheme exhibiting slightly better throughput than the SL adaptive modem. The adaptive modem employing pre-equalisation (PE) significantly outperforms the other uncoded adaptive schemes and offers a throughput of 0.18 BPS at an SNR of 0 dB.



The coded transmission schemes suffer from limited throughput at high SNR values, since the half-rate channel coding limits the data throughput to 2 BPS. For low SNR values, however, the coded schemes offer better performance than the uncoded schemes, with the exception of the “speech” SL adaptive coded scheme, which is outperformed by the uncoded PE adaptive modem. The poor performance of the coded SL scheme can be explained by the lower uncoded target BER of the “speech” scenario, which was 1%, in contrast to the 10% uncoded target BER for the coded BER and PE adaptive schemes. The coded PE adaptive modem outperforms the target-BER adaptive scheme, thanks to its more accurate control of the uncoded BER, leading to a higher throughput for low SNR values.

It is interesting to observe that for the given set of four modulation modes the uncoded PE adaptive scheme is close in performance to the coded adaptive schemes, and that for SNR values of more than 14 dB it outperforms all other studied schemes. It is clear, however, that the coded schemes would benefit from higher-order modulation modes, which would allow these modems to increase the data throughput further when the channel conditions allow. Before concluding this chapter, in the next section let us consider the generic problem of optimum power and bit allocation in the context of uncoded OFDM systems.

## 22.6 A Fast Algorithm for Near-Optimum Power and Bit Allocation in OFDM Systems [439]

### 22.6.1 State-of-the-art

In this section the problem of efficient OFDM symbol-by-symbol based power and bit allocation is analysed in the context of highly dispersive time-variant channels. A range of solutions published in the literature is reviewed briefly and Piazza’s [439] computationally efficient algorithm is exposed in somewhat more detail.

When OFDM is invoked over highly frequency selective channels, each subcarrier can be allocated a different transmit power and a different modulation mode. This OFDM symbol-by-symbol based “resource” allocation can be optimised with the aid of an algorithm which, if the channel is time variant, has to be repeated on an OFDM symbol-by-symbol basis. Some of the existing algorithms [83, 440] are mainly of theoretical interest due to their high complexity. Amongst the practical algorithms [426, 427, 439, 441, 442] the Hughes-Hartog algorithm (HHA) [442–444] is perhaps best known, but its complexity is somewhat high, especially for real-time OFDM symbol-by-symbol based applications at high bit rates. Hence the HHA has stimulated extensive research for computationally more efficient algorithms [426, 427, 439, 441, 442]. The most efficient appears to be that of Lai *et al.* [442], which is a fast version of the HHA and that of Piazza [439].

### 22.6.2 Problem description

Piazza [439] considered an OFDM system using  $N$  subcarriers, each employing a potentially different modulation mode and transmit power. Below we follow the

notation and approach proposed by Piazzo [439]. The different modes use different modum constellations and thus carry a different number of bits per subcarrier, ranging from 1 to  $I$  bits per subcarrier, corresponding to BPSK and  $2^I$ -ary QAM. We denote the transmit power and the number of bits allocated to subcarrier  $k$  ( $k = 0, \dots, N-1$ ) by  $p_k$  and  $b_k$ , respectively. If  $b_k = 0$ , subcarrier  $k$  is allocated no power and no bits, hence it is disabled. The total transmit power is  $P = \frac{1}{N} \sum_{k=0}^{N-1} p_k$  and the number of transmitted bits per OFDM symbol is  $B = \sum_{k=0}^{N-1} b_k$ . The  $i$ -bit modulation mode is characterised by the function  $R_i(S)$ , denoting the SNR required at the input of the detector, in order to achieve a target bit error rate (BER) equal to  $S$ . Finally, we denote the channel's power attenuation at subcarrier  $k$  by  $a_k$ , and the power of the Gaussian noise by  $P_N$ , so that the SNR of subcarrier  $k$  is  $r_k = p_k / (a_k \cdot P_N)$ .

We consider the problem of minimising the transmit power for a fixed target BER of  $S$  and for a fixed number of transmitted bits  $B$  per OFDM symbol. We impose an additional constraint, namely that the BER of every carrier has to be equal to  $S$ . This constraint simplifies the problem, while producing a system close to the unconstrained optimum system [427, 442, 444], while [440] considers an unconstrained system. Furthermore, from an important practical point of view, it produces a near-constant BER at the input of the channel decoder, if FEC is used, which maximises the achievable coding gain, since the channel does not become overwhelmed by the plethora of transmission errors, which would be the case for a more bursty error statistics without this constraint. In order to satisfy this constraint, the power transmitted on subcarrier  $k$  has to be  $p_k = P_N a_k R_{b_k}(S)$ , and the total power to be minimised is given by the sum of the  $N$  subcarriers' powers across the OFDM symbol:

$$P = \frac{P_N}{N} \sum_{k=0}^{N-1} a_k \cdot R_{b_k}(S). \quad (22.6)$$

We now state a property of the optimum system. Namely, in the optimum system if a subcarrier has a lower attenuation than another one - i.e. it exhibits a higher frequency domain transfer function value and hence experiences a higher received SNR - then it must carry at least as many bits as the lower SNR subcarrier. More explicitly:

$$a_k < a_h \Rightarrow b_k \geq b_h. \quad (22.7)$$

The above property in Equation 22.7 can be readily proven. Let us briefly consider a system, which does not satisfy Equation 22.7, where for subcarriers  $k$  and  $h$  the above condition is violated and hence we have  $a_k < a_h$  and  $b_k = i_1 < b_h = i_2$ . In other words, although the attenuation  $a_k$  is lower than  $a_h$ ,  $i_1 < i_2$ . Consider now a second system, where the lower attenuation subcarrier was assigned the higher number of bits, i.e.  $b_k = i_2$  and  $b_h = i_1$ . Since the required SNR for maintaining the target BER of  $S$  is lower for a lower number of bits, - i.e. we have  $R_{i_1}(S) < R_{i_2}(S)$ , upon substituting these SNR values in Equation 22.6 we can infer that the second system requires a lower total power  $P$  per OFDM symbol for maintaining the target BER  $S$ . Thus the first system is not optimum in this sense.

Equation 22.7 states a necessary condition of optimality, which was also exploited

From now on, we consider the channel's transfer function or attenuation vector sorted in the order of  $a_0 \leq a_1 \leq a_2 \dots$ , which simplifies our forthcoming discussions.

### 22.6.3 Power and bit allocation algorithm

Piazzo's algorithm [439] solves the above resource allocation problem for the general system by repeatedly solving the problem for a simpler system. Explicitly, the simpler system employs only two modulation modes, those carrying  $J$  and  $J - 1$  bits. This system can be termed the twin-mode system (TMS). On the basis of Equation 22.7 and since the channel's frequency-domain attenuation vector was sorted in the order of  $a_0 \leq a_1 \leq a_2 \dots$ , for the optimum TMS (OTMS) the OFDM subcarriers will be partitioned in three groups:

- (1) Group  $J$  comprises the first or lowest attenuation OFDM subcarriers using a  $J$ -bit modulation mode.
- (2) Group 0 is constituted by the last or highest attenuation OFDM subcarriers transmitting zero bits.
- (3) Group  $(J - 1)$  hosts the remaining OFDM subcarriers using a  $(J - 1)$ -bit modem mode.

In order to find the OTMS - minimising the required transmit power of the OFDM symbol for a fixed target BER of  $S$  and for a fixed number of transmitted bits  $B$  per OFDM symbol - we initially assign all the  $B$  bits of the OFDM symbol to the highest quality, i.e. lowest attenuation, group  $J$ . This of course would be a sub-optimum scheme, leaving the medium quality subcarriers of group  $J - 1$  unused, since even the highest quality subcarriers would require an excessive SNR, i.e. transmit power, for maintaining the target BER, when transmitting  $B$  bits per OFDM symbol. We note furthermore that the above bit allocation may require padding of the OFDM symbol with dummy bits if  $J$  is not an integer divisor of  $B$ .

Following the above initial bit allocation, Piazzo suggested performing a series of *bit reallocations, reducing the transmit power upon each reallocation*. Specifically, in each power and bit reallocation step we move the  $J \cdot (J - 1)$  bits allocated to the last, i.e. highest attenuation or lowest quality,  $(J - 1)$  OFDM subcarriers of group  $J$  to group  $(J - 1)$ . For example, if 1 bit/symbol BPSK and 2 bit/symbol 4-QAM are used, then we move  $2 \cdot 1 = 2$  bits, which were allocated to the highest attenuation 4-QAM subcarrier to two BPSK modulated subcarriers. The associated trade-off is that while previously the lowest quality subcarrier of class  $J$  had to carry 2 bits, it will now be conveying only 1 bit and additionally the highest quality and previously unused subcarrier has to be assigned 1 bit. This reallocation was motivated by the fact that before reallocation the lowest quality subcarrier of class  $J$  would have required a higher power for meeting the target BER requirement of  $S$  upon carrying 2 bits, than the regime generated by the reallocation step.

In general, for the sake of performing this power-reducing bit reallocation we have to add  $J$  subcarriers to group  $(J - 1)$ . Hence we assign the last, i.e. lowest quality,  $(J - 1)$  OFDM subcarriers of group  $J$  and the first, i.e. highest quality, unused subcarrier to group  $(J - 1)$ . Based on Equation 22.6 and upon denoting the index of

the last, i.e. lowest quality, subcarrier of group  $J$  before reallocation by  $M_J$  and the index of the first, i.e. highest quality, unused subcarrier before reallocation by  $M_0$ , the condition of successful power reduction after the tentative bit reallocation can be formulated. Specifically, the bit reallocation results in a system using less power, if the sum of the subcarriers' attenuations carrying  $J$  bits weighted by their SNR  $R_J(S)$  required for the  $J$ -bit modem mode for maintaining the target BER of  $S$  is higher than that of the corresponding constellation after the above bit reallocation process, when an extra previously unused subcarrier was invoked for transmission. This can be expressed in a more compact form as:

$$R_J(S) \sum_{k=0}^{J-2} a_{M_J-k} > R_{J-1}(S) (a_{M_0} + \sum_{k=0}^{J-2} a_{M_J-k}). \quad (22.8)$$

If Equation 22.8 is satisfied, the reallocation is performed and another tentative reallocation step is attempted. Otherwise the process is terminated, since the optimum twin-mode power and bit allocation scheme has been found.

According to Piazza's proposition [439] the above procedure can be further accelerated. Since the attenuation vector was sorted, we have  $a_{M_J-k} \approx a_{M_J}$  in Equation 22.8. Upon replacing  $a_{M_J-k}$  by  $a_{M_J}$ , after some manipulations we can reformulate Equation 22.8, i.e. the condition for the modem mode allocation after the bit reallocation to become more efficient as:

$$K_J(S) a_{M_J} - a_{M_0} > 0, \quad (22.9)$$

where  $K_J(S) = (J-1) \left( \frac{R_J(S)}{R_{J-1}(S)} - 1 \right)$  and  $K_J(S) > 0$  holds, since  $R_J(S) > R_{J-1}(S)$ . Piazza denoted the values of  $M_J$  and  $M_0$  after  $m$  reallocation steps by  $M_J(m)$  and  $M_0(m)$ . Since initially all the bits were allocated to group  $J$ , we have for the index of the last subcarrier of group  $J$  at the commencement of the bit reallocation steps  $M_J(0) = \lfloor B/J \rfloor - 1$ , while for the index of the first unused subcarrier is  $M_0(0) = \lfloor B/J \rfloor$ , where  $\lfloor x \rfloor$  is the smallest integer greater than or equal to  $x$ . Furthermore, since in each bit reallocation step the last  $(J-1)$  OFDM subcarriers of group  $J$  and the first subcarrier of group 0 are moved to group  $(J-1)$ , after  $m$  reallocations we have  $M_J(m) = \lfloor B/J \rfloor - 1 - (J-1)m$  and  $M_0(m) = \lfloor B/J \rfloor + m$ . Upon substituting these values in Equation 22.9 the left-hand side becomes a function of  $m$ , namely  $f(m) = K_J(S) a_{M_J(m)} - a_{M_0(m)}$ . Because the frequency domain channel transfer function's attenuation vector was ordered and since  $K_J(S) > 0$ , hence it is readily seen that  $f(m)$  is a monotonically decreasing function of the reallocation index  $m$ . Therefore the method presented above essentially attempts to find the specific value of the reallocation index  $m$ , for which we have  $f(m) > 0$  and  $f(m+1) < 0$ . In other words, when we have  $f(m+1) < 0$ , the last reallocation step resulted in a power increment, not a decrement, hence the reallocation procedure is completed.

The search commences from  $m = 0$  and increases  $m$  by one at each bit reallocation step. In order to accelerate the search procedure, Piazza replaced the linearly incremented search by a logarithmic search. This is possible, since the  $f(m)$  function is monotonically decreasing upon increasing the reallocation index  $m$ . Piazza [439] stipulated the search range by commencing from the minimum value of the reallocation

index  $m$ , namely from  $m_0 = 0$ .

The maximum value, denoted by  $m_1$ , is determined by the number of OFDM subcarriers  $N$  or by the number of bits  $B$  to be transmitted per OFDM symbol, as will be argued below. There are two limitations, which determine the maximum possible number of reallocation steps. Namely, the reallocation steps have to be curtailed when there are no more bits left in the group of subcarriers associated with the  $J$ -bit modem mode group or when there are no more unused carriers left after iteratively invoking the best unused carrier from the group of disabled carriers. These limiting factors, which determine the maximum possible number of bit reallocation steps, are augmented further below.

Recall that at the commencement of the algorithm all the bits were assigned to the subcarrier group associated with the  $J$ -bit modem mode and hence there were  $\lfloor B/J \rfloor$  subcarriers in group  $J$ . Upon reallocating the  $J \cdot (J - 1)$  bits allocated to the last, i.e. highest attenuation or lowest quality,  $(J - 1)$  OFDM subcarriers of group  $J$  to group  $(J - 1)$  until no more bits were left in the subcarrier group associated with the  $J$ -bit modem mode naturally constitutes an upper limit for the maximum number of reallocation steps  $m_1$ , which is given by  $\lfloor B/J \rfloor / (J - 1)$ . Again, the other limiting factor of the maximum number of bit reallocation steps is the number of originally unused carriers, which was  $N - \lfloor B/J \rfloor$ . Hence the maximum possible number of reallocations is given by  $m_1 = \min(\lfloor \lfloor B/J \rfloor / (J - 1) \rfloor; N - \lfloor B/J \rfloor)$ , where  $\lfloor x \rfloor$  is the highest integer smaller than or equal to  $x$ .

The accelerated logarithmic search proposed by Piazzo [439] halves the maximum possible range at each bit reallocation step, by testing the value of  $f(m)$  at the centre of the range and by updating the range accordingly. Piazzo's proposed algorithm can be summarised in a compact form as follows [439]:

**Algorithm 1**  $OTMS(B, S, J, N, a_k)$

- (1) Initialise  $m_0 = 0$ ,  $m_1 = \min(\lfloor \lfloor B/J \rfloor / (J - 1) \rfloor, N - \lfloor B/J \rfloor)$ .
- (2) Compute  $m_x = m_0 + \lfloor m_1 - m_0 \rfloor / 2$ .
- (3) If  $f(m_x) \geq 0$  let  $m_0 = m_x + 1$ ; else let  $m_1 = m_x$ .
- (4) If  $m_1 = m_0 + 1$  goto step 5); else goto step 2).
- (5) Stop. The number of carriers in group  $J$  is  $N_J = \lfloor B/J \rfloor - m_0 \cdot (J - 1)$ .

When the algorithm is completed, the value  $N_J$  becomes known and this specifies the number of OFDM subcarriers in the group  $J$  associated with the  $J$ -bit modem mode.

Having generated the optimum twin-mode system, Piazzo also considered the problem of finding the optimum general system (OGS) employing OFDM subcarrier modulation modes carrying  $1, \dots, I$  bits. The procedure proposed initially invoked Algorithm 1 in order to find the optimum twin-mode system carrying a total of  $B$  bits per OFDM symbol using the  $I$ -bit and the  $(I - 1)$ -bit per subcarrier modulation modes. At the completion of Algorithm 1 we know  $N_I$ , the number of OFDM subcarriers carrying  $I$  bits. These subcarriers are now confirmed. These OFDM subcarriers as well as the associated  $I \cdot N_I$  bits can now be eliminated from the resource allocation problem, and the optimum system transmitting the remaining  $B - I \cdot N_I$  bits of the remaining  $(N - N_I)$  subcarriers can be sought, using subcarrier modulation modes transmitting  $(I - 1), (I - 2), \dots$  bits.

Again, Algorithm 1 can be applied to this new system, now using the modulation modes with  $(I - 1)$  and  $(I - 2)$  bits per subcarrier and repeating the procedure. After each application of Algorithm 1 a new group of subcarrier is confirmed. Piazza's general algorithm can be summarised in a compact form as follows [439]:

**Algorithm 2**  $OGS(B, S, I, N, a_k)$

- (1) Initialise  $\hat{B} = B$ ,  $\hat{N} = N$ ,  $\hat{a}_k = a_k$ ,  $J = I$ .
- (2) Perform  $OTMS(\hat{B}, S, J, \hat{N}, \hat{a}_k)$  to compute  $N_J$ .
- (3) If  $J = 2$ , let  $N_1 = \hat{B} - 2 \cdot N_2$  and stop.
- (4) Remove the first  $N_J$  carriers from  $\hat{a}_k$ , let  $\hat{B} = \hat{B} - J \cdot N_J$ ,  $\hat{N} = \hat{N} - N_J$ ,  $J = J - 1$  and goto step 2).

When the algorithm is completed, the values  $N_i$  specifying the number of OFDM subcarriers conveying  $i$  bits, become known for all the legitimate modes carrying  $i = I, (I - 1), \dots, 1$  bits per subcarrier. Hence we know the number of bits allocated to subcarrier  $k$  ( $k = 0, \dots, N - 1$ ) expressed in terms of the  $b_k$  values as well as the associated minimum power requirements. Hence the system is specified in terms of  $p_k = P_N a_k R_{b_k}(S)$ . In closing it is worthwhile noting that the algorithm can be readily modified to handle the case where the two modes of the twin-mode system carry  $J$  and  $K < J - 1$  bits.

Having considered the above near-optimum power- and bit allocation algorithm, in the next section we will consider a variety of OFDM systems supporting multiple users and invoking adaptive beam-steering.

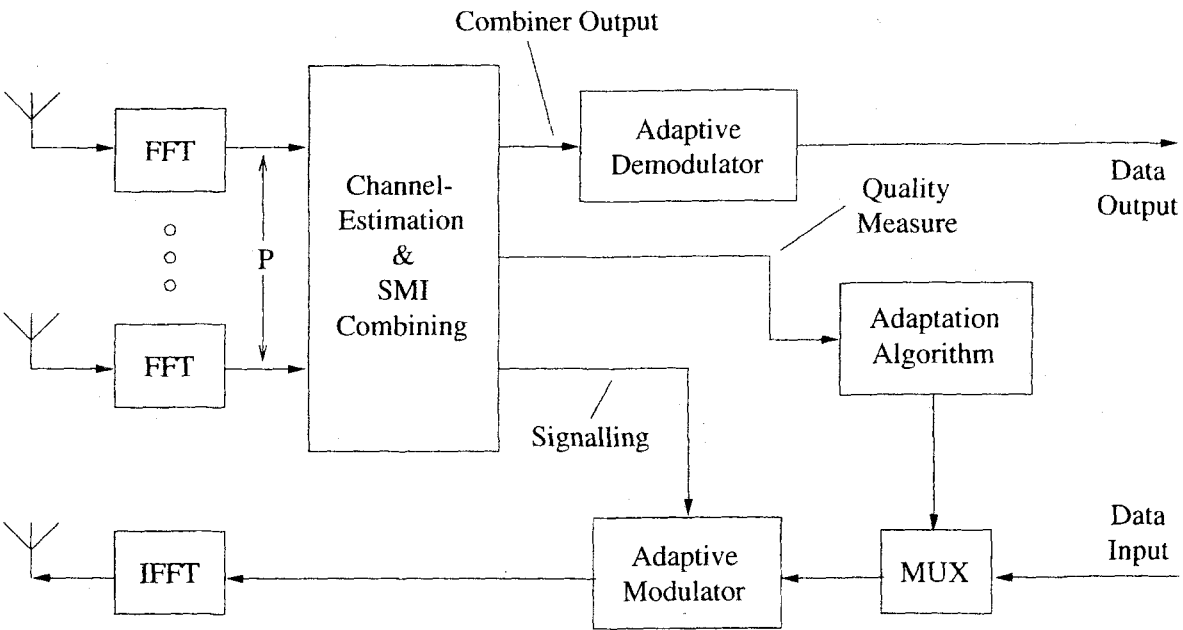
## 22.7 Multi-User AOFDM

M. Münster, T. Keller and L. Hanzo<sup>3</sup>

### 22.7.1 Introduction

Signal fading as well as co-channel interference are known to have a severe impact on the system performance in multi-cellular mobile environments. *Adaptive modulation* as a method of matching the system to fading induced variations of the channel quality has originally been proposed for single carrier transmission, but its potential was also soon discovered in the context of multicarrier transmissions, with the aim of concentrating the throughput on subcarriers least affected by frequency selective fading [437]. On the other hand, adaptive antenna array techniques have been shown to be effective in reducing co-channel interference at the receiver side [445, 446]. One of the most prominent schemes for performing the combining operation is the *Sample Matrix Inversion* (SMI) technique, which has recently drawn wide interest [132, 133, 447]. We commence our discussions in the next section with a description of a system amalgamating adaptive modulation and co-channel interference suppression. Initial performance results will be presented in Subsection 22.7.3 assuming perfect knowledge

<sup>3</sup>This section is based on M. Münster, T. Keller and L. Hanzo; Co-Channel Interference Suppression Assisted Adaptive OFDM In Interference Limited Environments, ©IEEE, VTC'99, Amsterdam, NL, 17-19 Sept. 1999



**Figure 22.30:** Schematic structure of the adaptive transceiver with interference suppression at the receiver

of all channel parameters, whilst in Subsection 22.7.4 the problem of channel parameter estimation will be addressed by means of orthogonal pilot sequences, leading to our conclusions.

### 22.7.2 Adaptive transceiver architecture

**An Overview** - The transceiver schematic is shown in Figure 22.30, where the receiver employs a multiple-antenna assisted front end. The signal received by each individual antenna element is fed to an FFT block, and the resulting parallel received OFDM symbols are combined on a subcarrier-by-subcarrier basis. The combining is accomplished on the basis of the weight vector, which has been obtained by solving the Wiener equation, constituting the core of the sample matrix inversion algorithm [445, 446]. After combining the signal is fed into the adaptive demodulator of Figure 22.30, which delivers the output bits in the form of soft-decision information to an optional channel decoder. The demodulator operates in one of a set of four modes, namely 'no transmission', BPSK, QPSK and 16-QAM. Since an interfered channel cannot be considered to constitute a reciprocal system, the modem mode adaptation operates in a closed-loop fashion, where each of the receivers instructs the remote transmitter as to the required set of modulation modes for the next AOFDM symbol, which is necessary for maintaining a given target Bits per Symbol (BPS) performance. On reception of a packet the adaptation algorithm computes the set of modulation modes to be employed by the remote transmitter for the next transmitted AOFDM symbol on the basis of a channel quality measure, namely the subcarrier SNR, which can be estimated by the interference suppression algorithm. The set of requested modulation modes is signalled to the remote receiver along with the next transmitted AOFDM

symbol, which is then used by the remote transmitter in its next transmission.

**The Signal Model** - The  $P \times 1$  vector of complex signals,  $\mathbf{x}[n, k]$ , received by the antenna array in the  $k$ -th subcarrier of the  $n$ -th OFDM symbol is constituted by a superposition of the independently faded signals associated with the *desired* user and the  $L$  *undesired* users plus the Gaussian noise at the array elements:

$$\begin{aligned} \mathbf{x}[n, k] &= \mathbf{d}[n, k] + \mathbf{u}[n, k] + \mathbf{n}[n, k], \quad \text{with} \\ \mathbf{d}[n, k] &= \mathbf{H}^{(0)}[n, k]s_0[k] \\ \mathbf{u}[n, k] &= \sum_{l=1}^L \mathbf{H}^{(l)}[n, k]s_l[k], \end{aligned} \quad (22.10)$$

where  $\mathbf{H}^{(l)}[n, k]$  for  $l = 0, \dots, L$  denotes the  $P \times 1$  vector of complex channel coefficients between the  $l$ -th user and the  $P$  antenna array elements. We assume that the vector components  $H_m^{(l)}[n, k]$  for different array elements  $m$  or users  $l$  are independent, stationary, complex Gaussian distributed processes with zero-mean and different variance  $\sigma_l^2$ ,  $l = 0, \dots, L$ . The variable  $s_l[n, k]$  - which is assumed to have zero-mean and unit variance - represents the complex data of the  $l$ -th user and  $\mathbf{n}[n, k]$  denotes the aforementioned  $P \times 1$  vector of additive white Gaussian noise contributions with zero mean and variance  $\sigma^2$  [132].

**The SMI Algorithm** - The idea behind minimum mean-square error (MMSE) beamforming [446] is to adjust the antenna weights, such that the power of the differential signal between the combiner output and a reference signal - which is characteristic of the desired user - is minimised. The solution to this problem is given by the well-known Wiener equation, which can be directly solved by means of the SMI algorithm in order to yield the optimum weight vector  $\mathbf{w}[n, k]$  of dimension  $P \times 1$ . Once the instantaneous correlation between the received signals, which is represented by the  $P \times P$  matrix  $\mathbf{R}[n, k]$  - and the channel vector  $\mathbf{H}^{(0)}[n, k]$  of the desired user become known, the weights are given by [132, 446, 447]:

$$\mathbf{w}[n, k] = (\mathbf{R}[n, k] + \gamma \mathbf{I})^{-1} \mathbf{H}^{(0)}[n, k], \quad (22.11)$$

where  $\gamma$  represents the so-called diagonal augmentation factor [132]. Assuming knowledge of all channel parameters and the noise variance  $\sigma^2$ , the correlation matrix can be determined by:

$$\begin{aligned} \mathbf{R}[n, k] &\triangleq E_c\{\mathbf{x}[n, k]\mathbf{x}^H[n, k]\} \\ &= \mathbf{R}_d[n, k] + \mathbf{R}_u[n, k] + \mathbf{R}_n[n, k], \quad \text{with} \\ \mathbf{R}_d[n, k] &= \mathbf{H}^{(0)}[n, k]\mathbf{H}^{(0)H}[n, k] \\ \mathbf{R}_u[n, k] &= \sum_{l=1}^L \mathbf{H}^{(l)}[n, k]\mathbf{H}^{(l)H}[n, k] \\ \mathbf{R}_n[n, k] &= \sigma^2 \mathbf{I}, \end{aligned} \quad (22.12)$$

which is a superposition of the correlation matrices  $\mathbf{R}_d[n, k]$ ,  $\mathbf{R}_u[n, k]$  and  $\mathbf{R}_n[n, k]$  of the desired and undesired users as well as of the array element noise, respectively.



The combiner output can now be inferred from the array output vector  $\mathbf{x}[n, k]$  by means of:

$$y[n, k] = \mathbf{w}^H[n, k]\mathbf{x}[n, k]. \quad (22.13)$$

The Signal-to-Noise Ratio (SNR) at the combiner output - which is of vital importance for the modulation mode adaptation - is given by [446]:

$$\begin{aligned} SNR &= \frac{E\{|\mathbf{w}^H[n, k]\mathbf{d}[n, k]|^2\}}{E\{|\mathbf{w}^H[n, k]\mathbf{n}[n, k]|^2\}} \\ &= \frac{\mathbf{w}^H[n, k]\mathbf{R}_d[n, k]\mathbf{w}[n, k]}{\mathbf{w}^H[n, k]\mathbf{R}_n[n, k]\mathbf{w}[n, k]} \end{aligned} \quad (22.14)$$

and correspondingly the Signal-to-Interference+Noise Ratio (SINR) is given by [446]:

$$SINR = \frac{\mathbf{w}^H[n, k]\mathbf{R}_d[n, k]\mathbf{w}[n, k]}{\mathbf{w}^H[n, k](\mathbf{R}_u[n, k] + \mathbf{R}_n[n, k])\mathbf{w}[n, k]}. \quad (22.15)$$

Equation 22.12 is the basis for our initial simulations, where *perfect channel knowledge* has been assumed. Since in a real environment the receiver does not have perfect knowledge of the channel, its parameters have to be estimated, a problem which we will address in Section 22.7.4 on an OFDM symbol-by-symbol basis by means of orthogonal pilot sequences.

**The Adaptive Bit-assignment Algorithm** - The adaptation performed by the modem is based on the choice between a set of four modulation modes, namely 4, 2, 1 and 0 bit/subcarrier, where the latter corresponds to 'no transmission'. The modulation mode could be assigned on a subcarrier-by-subcarrier basis, but the signalling overhead of such a system would be prohibitive. Hence, we have grouped adjacent subcarriers into 'sub-bands' and assign the same modulation mode to all subcarriers in a sub-band. Note that the frequency domain channel transfer function is typically not constant across the subcarriers of a sub-band, hence the modem mode adaptation will be sub-optimal for some of the subcarriers. The Signal-to-Noise Ratio (SNR) of the subcarriers will be shown to be in most cases an effective measure for controlling the modulation assignment. The modem mode adaptation is hence achieved by calculating in the first step for each sub-band and for all four modulation modes the expected overall sub-band bit error rate (BER) by means of averaging the estimated individual subcarrier BERs. Throughout the second step of the algorithm - commencing with the lowest modulation mode in all sub-bands - in each iteration the number of bits/subcarrier of that sub-band is increased, which provides the best compromise in terms of increasing the number of expected bit errors compared to the number of additional data bits accommodated, until the target number of bits is reached.

**The Channel Models** - Simulations have been conducted for the indoor Wireless Asynchronous Transfer Mode (WATM) channel impulse response (CIR) of Section 20.1 [437]. This three-path impulse response exhibits a maximal dispersion of 11 time-domain OFDM samples, with each path faded according to a Rayleigh distribution of a normalised maximal Doppler frequency of  $f'_d = 1.235 \cdot 10^{-5}$ , where the normalisation interval was the OFDM symbol duration. This model corresponds to the channel experienced by a mobile transmitting at a carrier frequency of 60 GHz

with a sampling rate of 225 MHz and travelling at a vehicular velocity of 50 km/h. An alternative channel model, which we considered in our simulations is a Wireless Local Area Network (WLAN) model associated with a seven-path impulse response having a maximal dispersion of 32 samples. However, for this more dispersive and higher Doppler frequency channel adaptive modulation has turned out to be less effective due to its significantly increased normalised Doppler frequency of  $f'_d = 3.935 \cdot 10^{-5}$  corresponding to a carrier frequency of 17 GHz, sampling rate of 20 MHz and vehicular velocity of 50 km/h.

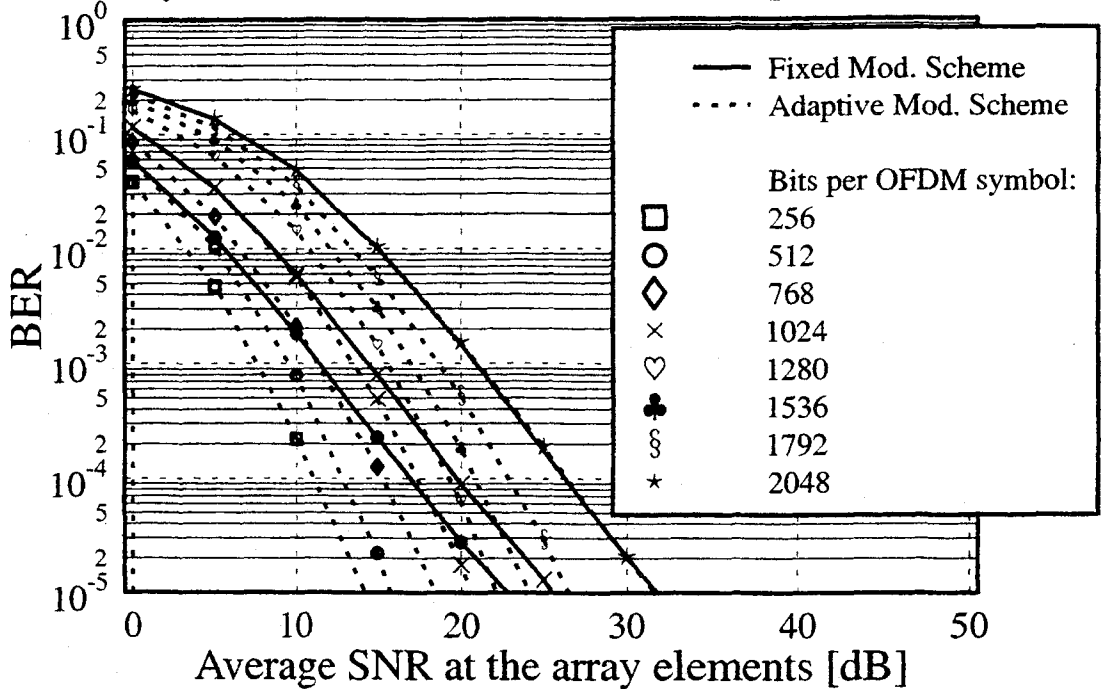
### 22.7.3 Simulation results - perfect channel knowledge

**General Remarks** - In our initial simulations we assumed that the receiver had perfect knowledge of all channel parameters, which enabled the estimation of the correlation matrix required by the SMI algorithm upon using Equation 22.12. Furthermore, we initially assumed that the receiver was capable of signalling the modulation modes to the transmitter without any additional delay. Throughout our discussions we will gradually remove the above idealistic assumptions. In all simulations we assumed a partitioning of the 512-subcarrier OFDM symbol's total bandwidth into 16 equal-sized 32-subcarrier sub-bands. This has been shown to provide a reasonable compromise between signalling overhead and performance degradation compared to a subcarrier-by-subcarrier based modulation mode assignment.

**Two-Branch Maximum-Ratio Combining** - Initial simulations were conducted in the absence of co-channel interference. In this scenario the SMI equations take the form of the MMSE maximum-ratio combiner, resulting in a high diversity gain even with the minimal configuration of only two reception elements. Adaptive modulation was performed on the basis of the estimated SNR of each subcarrier, which is given by Equation 22.14. Since due to diversity reception the dramatic fades of the channel frequency response have been mitigated, the performance advantage of adaptive modulation is more modest, as illustrated by Figure 22.31 for the 'transmission frame-invariant' WATM channel, for which the fading profile is kept constant for the OFDM symbol duration, in order to avoid inter-subcarrier interference. For the equivalent simulations in the 'transmission frame-invariant' WLAN channel environment we observed a more distinct performance gain due to adaptive modulation, which is justified by the higher degree of frequency selectivity introduced by the WLAN channel's seven-path impulse response.

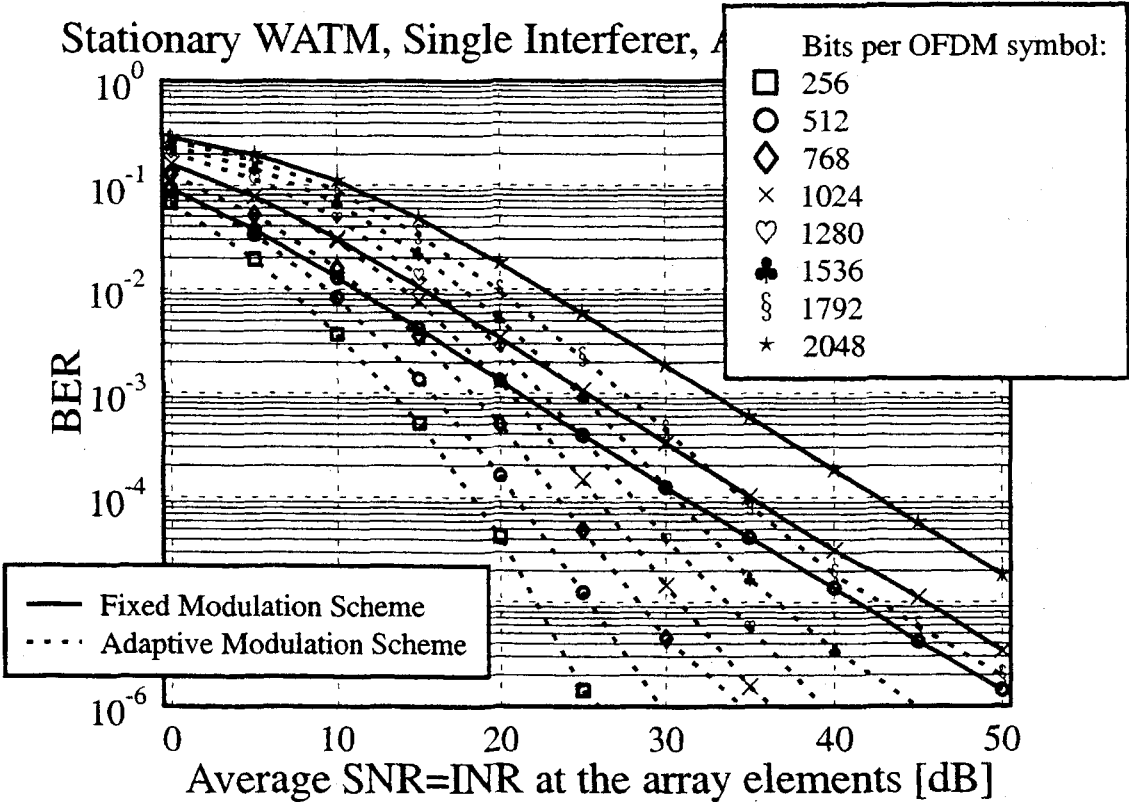
**SMI Co-Channel Interference Suppression** - In these simulations we considered first of all the case of a single dominant co-channel interferer of the same signal strength as the desired user. It is well-known that if the total number of users - whose signals arrive at the antenna array - is less or equal to the number of array elements, the unwanted users are suppressed quite effectively. Hence, for our modulation adaptation requirements we can assume that  $SNR \approx SINR$ , which enables us to use the algorithm described in Section 22.7.2, on the basis of the SNR estimated with the aid of Equation 22.14. Figure 22.32 illustrates the impact of adaptive modulation in the WATM channel environment under the outlined conditions. At a given SNR the performance gain due to adaptive modulation decreases with an increasing bitrate, since the higher bitrate imposes a more stringent constraint on the modulation

Stationary WATM, Two-Branch MRC, Adaptive Mod., 16 SB



**Figure 22.31:** BER of 16-sub-band AOFDM modem with *two-branch maximum-ratio combining* in a 'frame-invariant' indoor WATM environment, assuming perfect channel knowledge and zero-delay signalling of the modulation modes

mode assignment, invoking a higher number of low-SNR subcarriers. Upon comparing Figure 22.32 and 22.33 we observe that AOFDM attains a significantly higher SNR gain in the presence of co-channel interference, than without interference. As alluded to in the previous Section this is because under co-channel interference the SMI scheme exploits most of its diversity information extracted from the antenna array for suppressing the unwanted signal components, rather than mitigating the frequency domain channel fades experienced by the wanted user. For decreasing values of the Interference-to-Noise Ratio (INR) at the antenna array output, the system performance will gradually approach the performance observed for the MRC system. In order to render our investigations more realistic in our next experiment we allow a continuous, i.e. 'frame-variant' fading across the OFDM symbol duration. The system performance corresponding to this scenario is illustrated in Figure 22.33. At low SNRs the observed performance is identical to that recorded in Figure 22.32 for the 'frame-invariant' channel model, whereas at high SNRs we experience a residual BER due to inter-subcarrier interference. Again, for a low number of bits per OFDM symbol the adaptive scheme is capable of reducing the 'loading' of subcarriers with low SNR values, which are particularly impaired by inter-subcarrier interference. Hence AOFDM exhibits a BER improvement in excess of an order of magnitude. So far we



**Figure 22.32:** BER of 16-sub-band AOFDM modem with *two-branch SMI* and 2 users in a 'frame-invariant' indoor WATM environment, assuming perfect channel knowledge and zero-delay signalling of the modulation modes

have assumed that the receiver is capable of instantaneously signalling the required modulation modes for the next OFDM symbol to the transmitter. This assumption cannot be maintained in practice. Here we assume a time division duplexing (TDD) system with identical transceivers at both ends of the link, which communicate with each other using adjacent up-link and down-link slots. Hence we have to account for this by incorporating an additional delay of at least one OFDM symbol, while neglecting the finite signal processing delay. Simulation results for this scenario are depicted in Figure 22.34. We observe that the performance gain attained by adaptive modulation is reduced compared to that associated with the zero-delay assumption in Figure 22.33. This could partly be compensated for by channel prediction. When employing a higher number of array elements, the performance gain achievable by adaptive modulation will mainly depend on the number of users and their signal strength. If the number of users is lower, than the number of array elements, or if the interferers are predominantly weak, the remaining degrees of freedom for influencing the array response are dedicated by the SMI scheme to providing diversity for the reception of the wanted user and hence adaptive modulation proves less effective. If the

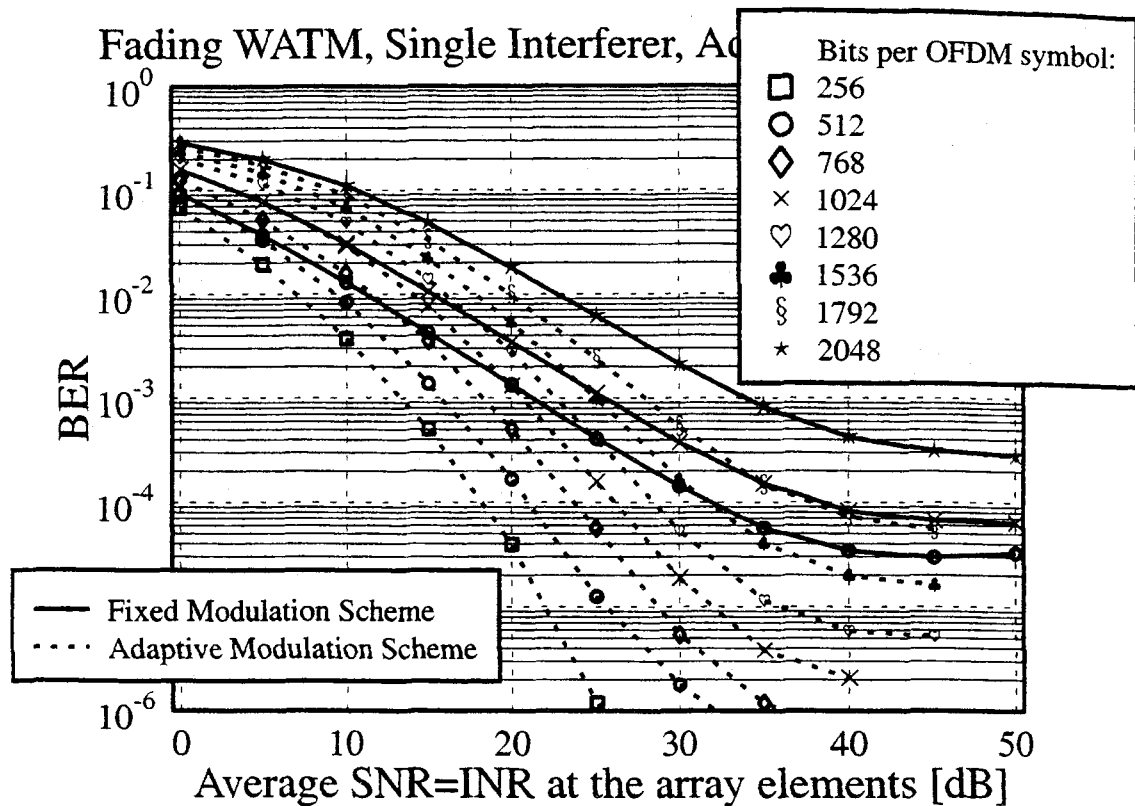


Figure 22.33: Performance results of Figure 22.32 repeated for a *fading indoor WATM* environment

number of users exceeds the number of array elements, the system becomes incapable of suppressing the undesired users effectively, resulting in a residual BER at high SNRs due to the residual co-channel interference. Since for a relatively high number of users the residual interference exhibits Gaussian-noise like characteristics, the SINR given by Equation 22.15 could be a suitable measure for performing the modulation mode assignment. By contrast, for a low number of interferers it is difficult to predict the impact on the system performance analytically. A possible approach would be to use the instantaneous number of errors in each sub-band (e.g. at the output of a turbo decoder) as a basis for the modulation assignment, which constitutes our future work. Let us now consider the issues of channel parameter estimation.

22.7.4 Pilot-based channel parameter estimation

**System Description** - Vook and Baum [447] have proposed SMI parameter estimation for OFDM by means of orthogonal reference sequences carrying pilot slots, which are transmitted over several OFDM symbol durations. This principle can also be applied on an OFDM symbol-by-symbol basis, as required for adaptive modulation. Upon invoking the idea of pilot based channel estimation by means of sampling

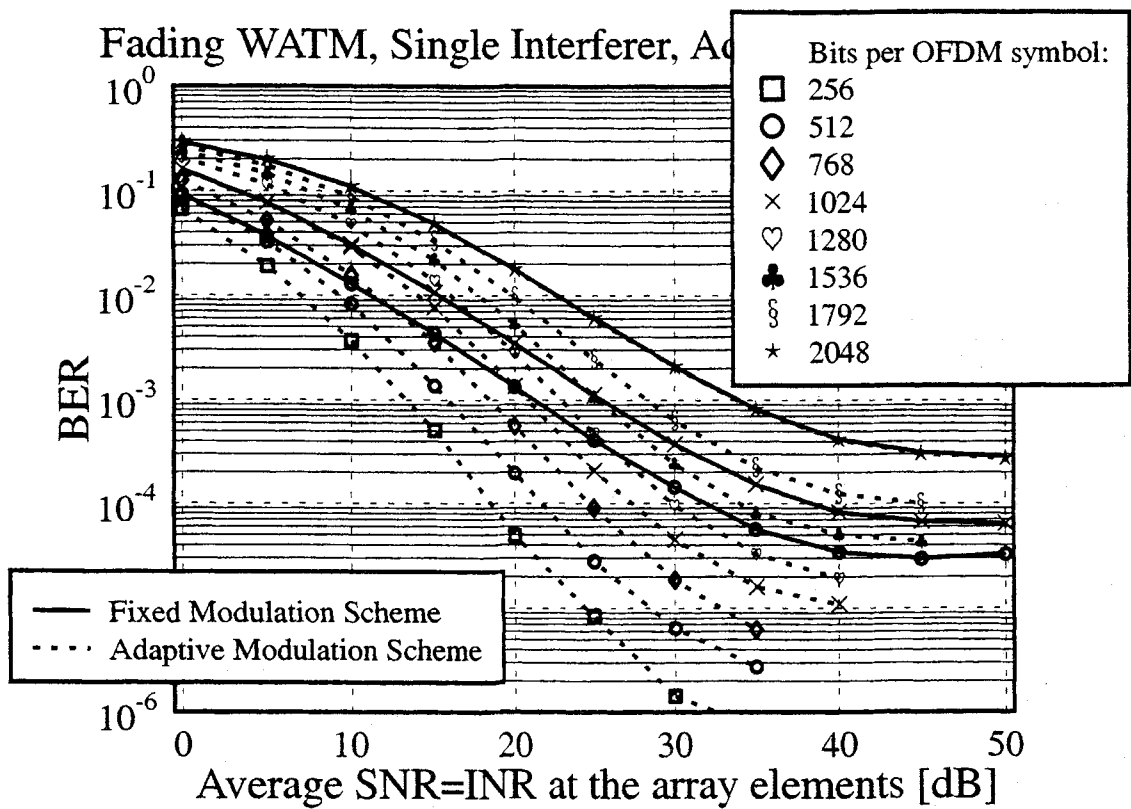


Figure 22.34: Performance results of Figure 22.33 repeated for *one OFDM symbol delayed signalling*

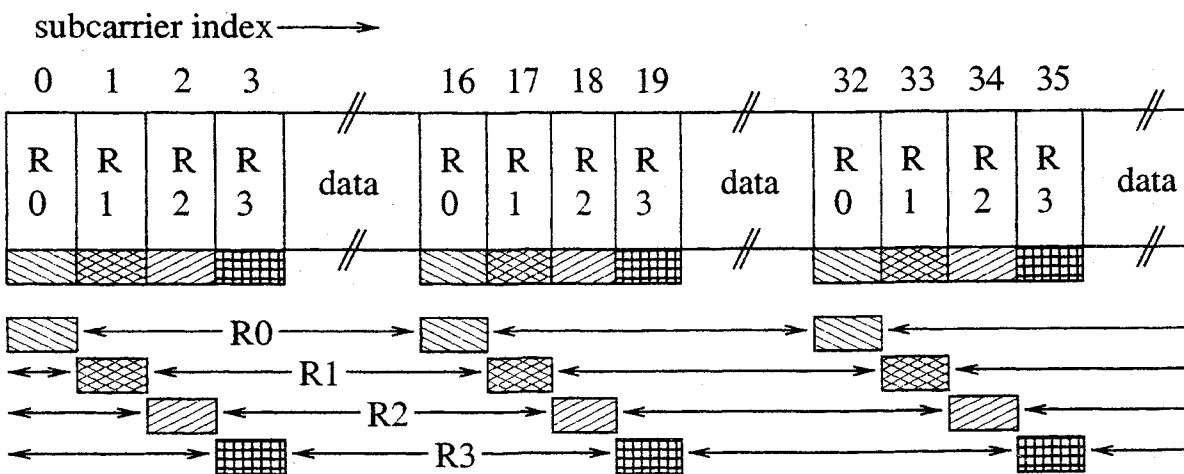


Figure 22.35: Pilot arrangement in each OFDM symbol for a *reference length of 4 bit* and a *group distance of 16 subcarriers*; interpolation is performed between pilots associated with the same bit position within the reference sequence

code/bit	0	1	2	3
0	1	1	1	1
1	1	1	-1	-1
2	1	-1	-1	1
3	1	-1	1	-1

**Table 22.4:** Orthogonal Walsh codes with a length of 4 bit

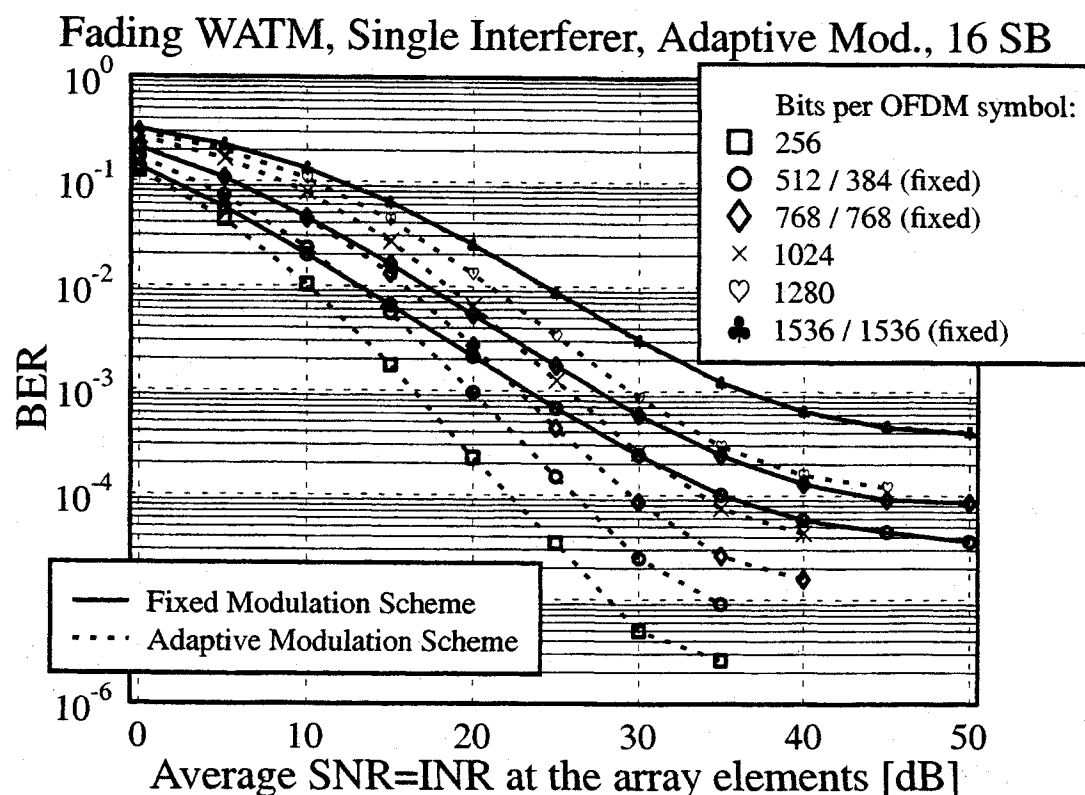
and low-pass interpolating the channel transfer function, we replace each single pilot subcarrier by a group of pilots, which carries a replica of the user's unique reference sequence. This is illustrated in Figure 22.35 for a reference sequence having a length of 4 bit, and for a pilot group distance of 16 subcarriers, which corresponds to the frequency required for sampling the WATM channel's transfer function. The corresponding 4-bit orthogonal Walsh code based reference sequences are listed in Table 22.4. Each of these 4 bits is assigned using BPSK to one of the 4 pilots in a pilot-group. The complex signal received by the  $m$ -th antenna in a pilot subcarrier at absolute index  $k$  and local index  $i$  within the reference sequence is constituted by a contribution of all users, each of which consists of the product of the Walsh code value associated with the user at bit position  $i$  of the reference sequence of Table 22.4 and the complex channel coefficient between the transmitter and the  $m$ -th antenna. MMSE lowpass interpolation is performed between all pilot symbols of the same relative index  $i$  within the  $k$ -spaced pilot blocks - as seen in Figure 22.35 - in order to generate an interpolated estimate of the reference for each subcarrier. An estimate of the channel vector  $\hat{\mathbf{H}}^{(0)}[n, k]$  and the correlation matrix  $\hat{\mathbf{R}}[n, k]$  for the  $k$ -th subcarrier of the  $n$ -th OFDM symbol is then given by [445–447]:

$$\hat{\mathbf{H}}^{(0)}[n, k] = \frac{1}{N} \sum_{i=0}^{N-1} r^{(0)*}(i) \mathbf{x}_{LP}[n, k](i) \quad (22.16)$$

$$\hat{\mathbf{R}}[n, k] = \frac{1}{N} \sum_{i=0}^{N-1} \mathbf{x}_{LP}[n, k](i) \mathbf{x}_{LP}^H[n, k](i), \quad (22.17)$$

where  $r^{(0)}(i)$  denotes the  $i$ -th value of the reference sequence associated with the desired user,  $\mathbf{x}_{LP}[n, k](i)$  represents the low-pass interpolated received signal at sequence position  $i$  and  $N$  denotes the total reference length.

**Simulation Results** - The performance of this scheme is characterised by the simulation results presented in Figure 22.36. Compared to the results presented in Figure 22.34 for 'perfect channel knowledge', in Figure 22.36 we observe that besides the reduced range of supported bitrates there is an additional performance degradation, which is closely related to the choice of the reference length. Specifically, there is a reduction in the number of useful data subcarriers due to the pilot overhead, which reduces the 'adaptively' exploitable diversity potential. Secondly, a relatively short reference sequence results in a limited accuracy of the estimated channel parameters - an effect which can be partly compensated for by a technique referred to as diagonal



**Figure 22.36:** BER of 16-sub-band adaptive OFDM modem with *two-branch SMI* and *2 users* in a *fading indoor WATM* environment, with *pilot based channel parameter estimation* and *one OFDM symbol delayed signalling* of the modulation assignment using a *diagonal loading* of  $\gamma = 1.0$

loading [132]. However, the effect of short reference sequences becomes obvious for a higher number of antenna elements, since more signal samples are required, in order to yield a reliable estimate of the correlation matrix. Hence our scheme proposed here is attractive for a scenario having 2-3 reception elements, where the interference is due to 1-2 dominant interferers and an additional Gaussian noise like contribution of background interferers, which renders the SINR of Equation 22.15 to be an effective measure of channel quality. **In conclusion** the proposed adaptive array assisted AOFDM scheme resulted typically in an order of magnitude BER reduction due to employing adaptive modulation.

## 22.8 Summary

A range of adaptive modulation and spectral predistortion techniques have been presented in this chapter, all of which aim to react to the time and frequency dependent channel transfer function experienced by OFDM modems in fading time dispersive



channels. It has been demonstrated that by exploiting the knowledge of the channel transfer function at the transmitter, the overall system performance can be increased substantially over the non-adaptive case. It has been pointed out that the prediction of the channel transfer function for the next transmission time slot and the signalling of the parameters are the main practical problems in the context of employing adaptive techniques in duplex communications.

The channel prediction accuracy is dependent on the quality of the channel estimation at the receiver, as well as on the temporal correlation of the channel transfer function between the uplink and downlink time slots. Two-dimensional channel estimation techniques [133,448] can be invoked in order to improve the channel prediction at the receivers.

It has been demonstrated that sub-band adaptivity instead of subcarrier-by-subcarrier adaptivity can significantly decrease the necessary signalling overhead, with a loss of system performance that is dependent on the channel's coherence bandwidth. We have seen that sub-band adaptivity allows the employment of blind detection techniques in order to minimise the signalling overhead. Further work into blind detection algorithms as well as new signalling techniques is needed for improving the overall bandwidth efficiency of adaptive OFDM systems.

Pre-equalisation or spectral predistortion techniques have been demonstrated to significantly improve an OFDM system's performance in time dispersive channels, while not increasing the system's output power. It has been shown that spectral predistortion can integrate well with adaptive modulation techniques, improving the system's performance significantly. We have seen in Figure 22.1 that a data throughput of 0.5 bits/symbol has been achieved at 0 dB average channel SNR with a BER of below  $10^{-4}$ .

An adaptive power and bit allocation algorithm was also highlighted and the performance of AOFDM in a beam-forming assisted multi-user scenario was characterised.



# Glossary

<b>8-DPSK</b>	8-Phase Differential Phase Shift Keying
<b>ACF</b>	autocorrelation function
<b>ADC</b>	Analog-to-Digital Converter
<b>ADM</b>	adaptive delta modulation
<b>ADPCM</b>	Adaptive Differential Pulse Coded Modulation.
<b>AGC</b>	Automatic Gain Control
<b>AM-PM</b>	amplitude modulation and phase modulation
<b>AWGN</b>	Additive White Gaussian Noise
<b>BCH</b>	Bose-Chaudhuri-Hocquenghem, A class of forward error correcting codes (FEC)
<b>BCM</b>	block code modulation
<b>BER</b>	Bit error rate, the fraction of the bits received incorrectly
<b>BPF</b>	Bandpass Filter
<b>BPSK</b>	Binary Phase Shift Keying
<b>BS</b>	A common abbreviation for Base Station
<b>C</b>	Channel Capacity
<b>CCITT</b>	Now ITU, standardisation group
<b>CD</b>	Code Division, a multiplexing technique where signals are coded and then combined, in such a way that they can be separated using the assigned user signature codes at a later stage.

CISI	controlled inter-symbol interference
CT	cordless telecommunications
DAB	Digital Audio Broadcasting
DECT	A Pan-European digital cordless telephone standard.
DFT	Discrete Fourier Transform
DTX	discontinuous transmission
EQ	Equaliser
FD	Frequency Division, a multiplexing technique, where different frequencies are used for each communications link.
FDD	Frequency-Division Duplex, a multiplexing technique, where the forward and reverse links use a different carrier frequency.
FDM	Frequency Division Multiplexing
FEC	Forward Error Correction
FFT	Fast Fourier Transform
FIR	Finite Impulse Response
FSK	Frequency Shift Keying
GF	Galois field
GMSK	Gaussian Mean Shift Keying, a modulation scheme used by the Pan-European GSM standard by virtue of its spectral compactness.
GSM	A Pan-European digital mobile radio standard, operating at 900MHz.
ICI	Inter-Channel Interference
IFFT	Inverse Fast Fourier Transform
IMD	Intermodulation Distortion
ISI	Inter Symbol Interference, Inter Subcarrier Interference
LOS	Line-Of-Sight
LPF	low pass filter
LSB	least significant bit
LSF	Least Squares Fitting

<b>LTP</b>	long term predictor
<b>MDI</b>	multi-dimensional interference
<b>MLSE</b>	maximum likelihood sequence estimator
<b>MOS</b>	mean opinion score
<b>MS</b>	A common abbreviation for Mobile Station
<b>MSK</b>	minimum shift keying
<b>NLA</b>	non-linear amplification
<b>NLF</b>	non-linear filtering
<b>OFDM</b>	Orthogonal Frequency Division Multiplexing
<b>OOB</b>	out of band
<b>OQAM</b>	offset quadrature amplitude modulation
<b>OQPSK</b>	offset quadrature phase shift keying
<b>OSWE</b>	one-symbol window equaliser
<b>PAM</b>	pulse amplitude modulation
<b>PCM</b>	pulse code modulation
<b>PCN</b>	Personal Communications Network
<b>PD</b>	phase detector
<b>PDF</b>	Probability Density Function
<b>PLL</b>	phase locked loop
<b>PR</b>	PseudoRandom
<b>PRMA</b>	Packet Reservation Multiple Access, a statistical multiplexing arrangement contrived to improve the efficiency of conventional TDMA systems, by detecting inactive speech segments using a voice activity detector, surrendering them and allocating them to subscribers contending to transmit an active speech packet.
<b>PSAM</b>	Pilot symbol assisted modulation, a technique where known symbols (pilots) are transmitted regularly. The effect of channel fading on all symbols can then be estimated by interpolating between the pilots
<b>PSD</b>	Power Spectral Density

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<b>PSK</b>	Phase Shift Keying
<b>QAM</b>	Quadrature Amplitude Modulation
<b>QPSK</b>	Quaternary Phase Shift Keying
<b>RF</b>	radio frequency
<b>RPE</b>	regular pulse excited
<b>RSSI</b>	Received Signal Strength Indicator, commonly used as an indicator of channel quality in a mobile radio network.
<b>SEGSNR</b>	segmental signal-to-noise ratio
<b>SIR</b>	Signal to Interference ratio
<b>SNR</b>	Signal to Noise Ratio, noise energy compared to the signal energy
<b>TC</b>	Trellis Coded
<b>TCM</b>	trellis code modulation
<b>TDD</b>	Time-Division Duplex, a technique where the forward and reverse links are multiplexed in time.
<b>TDMA</b>	Time Division Multiple Access
<b>TTIB</b>	transparent tone in band
<b>TWT</b>	travelling wave tube
<b>UHF</b>	ultra high frequency
<b>VAD</b>	voice activity detection
<b>VCO</b>	voltage controlled oscillator
<b>WN</b>	white noise

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