

Broad-Band OFDM Radio Transmission for Multimedia Applications

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Invited Paper

Future multimedia services will require the transmission of very high data rates over broad-band radio channels. In order to provide these services to mobile users, an appropriate transmission technique has to cope with frequency-selective and time-variant radio channels. The computation complexity for an equalizer increases in a frequency-selective radio channel for high data-rate applications. Furthermore, the overhead for channel estimation procedures increases in time-variant channels.

To overcome these drawbacks orthogonal frequency division multiplexing (OFDM) has been considered for broad-band applications in many publications. The objective of this paper is to describe the large potential, the flexibility and adaptivity of the OFDM transmission technique in frequency-selective and time-variant radio channels. In this paper several technical aspects of OFDM transmission systems are discussed, especially the topics of differential modulation for which we compare different demodulation methods and channel coding with soft decision decoding. For higher level differential modulation, multilevel coding is taken into consideration.

Keywords—Better differential modulation, broad-band communication, convolutional codes, differential amplitude and phase shift keying, modulation, modulation coding, multimedia systems, orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

In this paper, we discuss transmission techniques for broad-band mobile multimedia applications in frequency-selective and time-variant radio channels. Literally, multimedia means that different media like video, graphics, speech, text, or other data are used within the same radio link as requested in order to support communication. This means that the transmission system may have to adapt itself to very different requirements of the individual services in terms of data rate, admissible BER [quality of service (QoS)], and maximum delay. Furthermore, bidirectional communication has to be supported. A typical system which supports multimedia communication in this sense is the

Internet, which is already provided to everyone via fixed cable networks for the time being.

High bit-rate services like digital video transmission are principally not a technical problem in the case of wired networks due to the channel behavior. But for future developments it can be expected that there will also be a huge demand for mobile or at least wireless access to multimedia services. In the case of radio transmission, however, high data rates lead to additional technical considerations. A broad-band radio channel is characterized both by time-variant behavior caused by a moving receiver or transmitter and by frequency-selective fading which is caused by multipath propagation. Therefore, if a conventional single carrier system is used for this purpose, the channel equalization can be very complex. In Europe, two digital terrestrial broadcasting services have already been standardized which therefore alternatively use the orthogonal frequency division multiplexing (OFDM) transmission technique. These systems are called digital audio broadcasting (DAB) and digital terrestrial video broadcasting (DTVB). Both are pure broadcasting services which transmit data over broad-band radio channels. DAB is suited for mobile receivers and the total data rate is 1.7 Mbit/s. DTVB has a net data rate of more than 20 Mbit/s and is designed for stationary reception. Currently, there is a discussion about extending DAB in order to enable higher data rates for multimedia services, e.g., [18].

Meanwhile, the application of the OFDM transmission technique is discussed for various broad-band communication systems and services. ETSI-BRAN has selected this transmission technique for HIPERLAN 2 (high performance local area networks), and OFDM has also been chosen for the extension of the IEEE 802.11 standard for the 5-GHz frequency range. In these systems data rates of, e.g., 25 Mbit/s are considered. OFDM proposals have been described for the universal mobile telecommunication system (UMTS) in [2]. Future mobile communication systems with data rates far beyond UMTS can be realized using OFDM as a suitable transmission technique. An important

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motivation for using OFDM is to reduce the computation complexity associated with high data-rate transmission over frequency-selective radio channels.

If an OFDM system [1], [31] is used, the radio channel is divided into many narrow-band subchannels which appear to be frequency nonselective. Thus, the task of channel equalization is reduced to estimating a single complex factor (channel transfer factor) for each subcarrier. Such a channel estimation can be realized by inserting so-called pilot symbols with known modulation into the transmitted signal. Based on these pilot symbols the receiver can measure the channel transfer factors for each subcarrier using interpolation techniques [14]. In this case, each subcarrier can be demodulated coherently.

Alternatively, differential modulation can be performed according to the well-known M -DPSK (differential phase shift keying) scheme, which is a pure phase modulation. This scheme has been extended for higher level modulation to a combined differential amplitude and phase modulation in order to achieve a higher bandwidth efficiency and still rather good performance. This modulation technique termed differential amplitude and phase shift keying (DAPSK) is explained and analyzed in [11] and [23]. It is interesting to note that the performance of OFDM systems with differential modulation compares quite well with systems using nondifferential modulation and coherent demodulation [25]. Furthermore, the computation complexity in the demodulation process is quite low for differential modulation.

Channel coding is an important topic for OFDM systems. Transmission over a frequency-selective channel implies that some of the subcarriers are strongly attenuated and cause errors even at a high average signal power. In this flat fading situation an efficient channel coding leads to a very high coding gain, especially if soft decision decoding is applied. For this reason, OFDM systems will always have to make use of channel coding. In the following we describe coded OFDM systems for which the performance of coherent and differential modulation are compared.

Apart from simplifying the channel equalization with respect to single carrier systems, OFDM also offers more possibilities to adapt the transmission system to the radio channel by varying the modulation of the subcarriers [13]. However, an optimal adaptation is limited to quasi-stationary radio channels.

In Section II the OFDM transmission technique is explained. In Section III modulation and demodulation are described with emphasis on higher level differential modulation which plays an important role for OFDM systems. Convolutional channel coding and multilevel coding are discussed in Section IV. Finally, in Section V the system performance is presented on the basis of simulation results.

II. THE OFDM TRANSMISSION TECHNIQUE

In this section we give a brief overview of the history of OFDM and an analytical description of this transmission technique. Finally, we show the technical structure of an OFDM transmission system.

A. History of OFDM

The basic principles of OFDM have already been proposed in several publications in the 1960's. However, these ideas could not be realized efficiently, since powerful semiconductor devices were not available at that time. Today, even relatively complex OFDM transmission systems with high data rates are technically feasible, and such systems can be taken advantage of in frequency-selective radio channels.

In a classical FDM system, narrow-band signals are generated independently, assigned to various frequency bands, parallel transmitted, and separated by filters at the receiver. The new aspect of OFDM is that the various signals are generated jointly by a fast Fourier transform (FFT) and that their spectra overlap. As a result, generating the signal is simplified and the bandwidth efficiency of the system is improved.

It is interesting to note that as soon as in 1961 a code multiplexing scheme has been proposed where sine and cosine functions were used as orthogonal signals [12]. The resulting signal can already be compared with an OFDM signal. However, the fact that this system was identical with a FDM was not important for the proposal and the benefits in frequency-selective channels were not recognized.

Since 1966 FDM systems with overlapping spectra were proposed in several publications [5], [26], [34]. The next step was the proposal to realize an FDM system with the discrete Fourier transform (DFT) [10]. Finally, in 1971 Weinstein and Ebert have proposed a complete OFDM system [31], which included generating the signal with an FFT and adding a guard interval in the case of multipath channels. This system will be referred to in the following.

In the further development, OFDM has been discussed for channels with both flat and frequency-selective fading in [7]. In [1] it has been proposed for broadcast applications and mobile reception. Meanwhile, the OFDM transmission technique has been standardized in the DAB and DTVB systems.

The most important advantage of the OFDM transmission technique as compared to single carrier systems is obtained in frequency-selective channels. The signal processing in the receiver is rather simple in this case, because after transmission over the radio channel the orthogonality of the OFDM subcarriers is maintained and the channel interference effect is reduced to a multiplication of each subcarrier by a complex transfer factor. Therefore, equalizing the signal is very simple, whereas equalization may not be feasible in the case of conventional single carrier transmission covering the same bandwidth. It must be mentioned, however, that in [28] a single carrier system with frequency-domain equalization has been proposed which also copes with large delays.

B. Mathematical Description

1) *Transmitter*: An OFDM signal consists of N subcarriers spaced by the frequency distance Δf . Thus, the total system bandwidth B is divided into N equidistant

subchannels. All subcarriers are mutually orthogonal within a time interval of length $T_S = 1/\Delta f$. The k th subcarrier signal is described analytically by the function $g_k(t)$, $k = 0, \dots, N-1$. For each subcarrier a rectangular pulse shaping is applied

$$g_k(t) = \begin{cases} e^{j2\pi k \Delta f t}, & \forall t \in [0, T_S] \\ 0, & \forall t \notin [0, T_S] \end{cases} \quad (1)$$

Since the system bandwidth B is subdivided into N narrow-band subchannels, the OFDM block duration T_S is N times as large as in the case of a single carrier transmission system covering the same bandwidth. This subcarrier signal $g_k(t)$ is extended by a cyclic prefix (called guard interval) with the length T_G yielding the following signal:

$$g_k(t) = \begin{cases} e^{j2\pi k \Delta f t}, & \forall t \in [-T_G, T_S] \\ 0, & \forall t \notin [-T_G, T_S] \end{cases} \quad (2)$$

The guard interval is added to the subcarrier signal in order to avoid intersymbol interferences (ISI) which occur in multipath channels. At each receiver the guard interval is removed and only the time interval $[0, T_S]$ is evaluated. From this point of view the guard interval is a pure system overhead. The total OFDM block duration is $T = T_S + T_G$.

It is an important advantage of the OFDM transmission technique that ISI which occur in all multipath channels can be reduced considerably. If the guard interval length T_G is larger than the maximal delay in the radio channel, no ISI occur at all and the orthogonality of the subcarriers is not affected.

Each subcarrier can be modulated independently with the complex modulation symbol $S_{n,k}$, where the subscript n refers to the time and k to the number of the subcarrier inside the considered OFDM block. Thus, within the symbol duration T the following signal of the n th OFDM block is formed:

$$s_n(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_{n,k} g_k(t - nT). \quad (3)$$

The total continuous-time signal consisting of all OFDM blocks is

$$s(t) = \frac{1}{\sqrt{N}} \sum_{n=0}^{\infty} \sum_{k=0}^{N-1} S_{n,k} g_k(t - nT). \quad (4)$$

Due to the rectangular pulse shaping of the signal, the spectra of the subcarriers are si-functions, e.g., for the k th subcarrier:

$$G_k(f) = T \text{si} [\pi T(f - k\Delta f)]. \quad (5)$$

The spectra of the subcarriers overlap, but the subcarrier signals are mutually orthogonal and the modulation symbols $S_{n,k}$ can be recovered by a correlation

$$\langle g_k, g_l \rangle = \int_0^{T_S} g_k(t) \overline{g_l(t)} dt = T_S \delta_{k,l} \quad (6)$$

$$S_{n,k} = \frac{\sqrt{N}}{T_S} \langle s_n(t), \overline{g_k(t - nT)} \rangle \quad (7)$$

where $\overline{g_k(t)}$ is the conjugate of $g_k(t)$. In a practical application the OFDM signal $s_n(t)$ is generated in a first step as a discrete-time signal in the digital signal processing part of the transmitter. As the bandwidth of an OFDM system is $B = N\Delta f$, the signal must be sampled with the sampling time $\Delta t = 1/B = 1/N\Delta f$. The samples of the signal are written as $s_{n,i}$, $i = 0, 1, \dots, N-1$ and can be calculated as

$$s_{n,i} = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_{n,k} e^{j2\pi i k / N}. \quad (8)$$

This equation describes exactly the inverse discrete Fourier transform (IDFT) which is typically realized by an inverse FFT (IFFT).

2) *Receiver*: The subcarrier orthogonality is not affected at the output of the radio channel if the guard interval length T_G is larger than the maximal multipath delay. Therefore, the received signal $r_n(t)$ can be separated into the orthogonal subcarrier signals by a correlation technique according to (7)

$$R_{n,k} = \frac{\sqrt{N}}{T_S} \langle r_n(t), \overline{g_k(t - nT)} \rangle. \quad (9)$$

Alternatively, the correlation at the receiver can be realized as a DFT or an FFT, respectively

$$R_{n,k} = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} r_{n,i} e^{-j2\pi i k / N}. \quad (10)$$

Here $r_{n,i}$ is the i th sample of the received signal $r_n(t)$ and $R_{n,k}$ is the received complex symbol of the k th subcarrier. The FFT and IFFT algorithms can be implemented very efficiently.

If the subcarrier spacing Δf is chosen much smaller than the coherence bandwidth and the symbol duration T much smaller than the coherence time of the channel, then the transfer function of the radio channel $H(f, t)$ can be considered constant within the bandwidth Δf of each subcarrier and the duration of each modulation symbol $S_{n,k}$. In this case, the effect of the radio channel is only a multiplication of each subcarrier signal $g_k(t)$ by a complex transfer factor $H_{n,k} = H(k\Delta f, nT)$. As a result, the received complex symbol $R_{n,k}$ after the FFT is

$$R_{n,k} = H_{n,k} S_{n,k} + N_{n,k} \quad (11)$$

where $N_{n,k}$ is additive noise of the channel. Equation (11) implies perfect synchronization at the receiver, which is an important topic in OFDM transmission systems, since time and frequency synchronization errors disturb the orthogonality of the subcarriers and therefore reduce the SNR considerably. Synchronization of OFDM systems has been analyzed, e.g., in [8], [21], and [29].

Another aspect which is not taken into account in (11) is the signal distortion due to amplifier nonlinearities. Unfortunately, an OFDM signal has a high peak-to-average power ratio so that the power amplifier usually has to be operated at a large input backoff, mainly in order to avoid out-of-band interference. If, for example, the interference

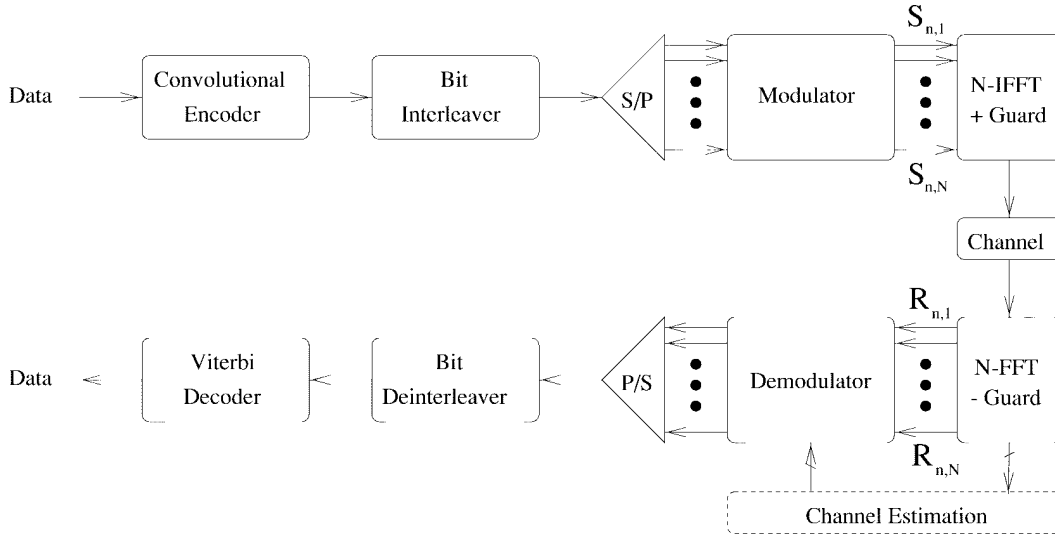


Fig. 1. Block diagram of an OFDM transmission system with convolutional encoding.

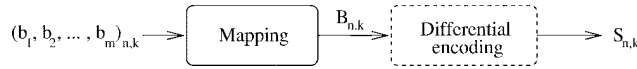


Fig. 2. Block diagram of linear subcarrier modulation.

in the adjacent frequency bands is to be lower than 40 dB below the power density in the OFDM band, then an input backoff of more than 7.5 dB is required. However, the amplitude range of OFDM signals can be reduced by means of coding or signal processing so that a lower input backoff is enabled. These topics have been analyzed, e.g., in [17], [22], and [20].

A general block diagram of the considered OFDM transmission system is depicted in Fig. 1. In the following the topics of modulation and channel coding are discussed more in detail.

III. MODULATION AND DEMODULATION

A. Modulation

As shown in (3), each subcarrier can be modulated independently by a complex modulation symbol $S_{n,k}$. Generally, this complex value is generated by mapping a sequence of m bits to a specific point in the constellation diagram followed by an optional differential encoding (see Fig. 2).

1) *Without Differential Encoding:* If no differential encoding is applied, the data bits to be transmitted are directly mapped to the modulation symbols $S_{n,k}$. Examples of this modulation technique are M -PSK and M -QAM.

2) *Differential Encoding:* Differential encoding for narrow-band channels implies that the transmitted information is contained in the quotient of two successive modulation symbols. Differential encoding means that the data bits to be transmitted are not directly mapped to the modulation symbols $S_{n,k}$, but rather to the quotient $B_{n,k}$

of two successively transmitted modulation symbols. This technique can be applied either in time or in frequency direction.

The operation of differential encoding in the time direction can be described analytically by the following multiplication:

$$S_{n,k} = S_{n-1,k} \cdot B_{n,k}. \quad (12)$$

In the case of M -DPSK for example, $B_{n,k} \in \{e^{j2\pi i/M} | i = 0, \dots, M-1\}$. However, M -DPSK has a poor performance if M is large. In this case a combined differential amplitude and phase modulation should be applied which is discussed in detail in Section III-C.

Differential encoding can alternatively be performed in frequency direction as described in (13)

$$S_{n,k} = S_{n,k-1} \cdot B_{n,k}. \quad (13)$$

In any case, the coherence time and the coherence bandwidth have to be large compared to the symbol duration and subcarrier spacing respectively, to assure that the corresponding channel transfer factors $H_{n,k}$ are approximately equal. Depending on the channel and on the design of the OFDM system, differential coding in time or frequency direction may be more suitable. Especially in communication systems differential modulation in frequency direction has some advantages with respect to the system overhead. For reasons of clarity, only differential encoding in time direction is considered in the following.

B. Demodulation

In this section it is described how the transmitted data bits can be recovered from the received symbol sequence $R_{n,k}$. This could more exactly be termed hard output demodulation. In case of channel coding, soft decision decoding may be preferable, which means that soft output demodulation is required. This topic is discussed in more detail in the channel coding section.

Depending on the type of modulation (differential or nondifferential), there are different ways to demodulate the received complex sequence which are discussed in detail in the following sections.

1) *Coherent Demodulation*: Coherent demodulation has to be applied if a nondifferential modulation is used in the transmitter. Originally, the term coherent means that the mixer in the receiver is synchronized in frequency and phase with the carrier frequency of the received signal. In case of an OFDM transmission system each subcarrier has to be synchronized or, equivalently, the phase offset has to be known. If any kind of amplitude modulation, like quadrature amplitude modulation (QAM), is applied then the attenuation of each subcarrier has to be known, too. To generate this information in the receiver, a channel estimation has to be performed which provides estimates $\hat{H}_{n,k}$ for the channel transfer factors. The decision is based on the quotient

$$D_{n,k}^c = \frac{R_{n,k}}{\hat{H}_{n,k}} = S_{n,k} + \frac{N_{n,k}}{\hat{H}_{n,k}} \Rightarrow \hat{S}_{n,k} = \text{dec} \{D_{n,k}^c\}. \quad (14)$$

For $D_{n,k}^c$ the receiver makes a decision according to given thresholds.

The channel estimation can be performed on the basis of known symbols (pilot symbols) which are included in the OFDM signal before transmission [14], [27]. In the first step, the receiver extracts the transfer factors at those times and frequencies at which pilot symbols have been transmitted. Then the transfer factors between the positions of pilot symbols can be interpolated by means of filtering. If the transfer function is filtered in time direction with a given density of pilot symbols, this requires the maximal Doppler frequency of the channel to be sufficiently small according to the sampling theorem. In the same way, filtering in frequency direction means that the maximal delay of the channel must be sufficiently small.

2) *Noncoherent Demodulation*: If differential encoding is used in the transmitter, the demodulation can be performed either noncoherently (nc) or quasi-coherently (qc). With noncoherent demodulation the decision is based on the quotient of two successive symbols

$$D_{n,k}^{\text{nc}} = \frac{R_{n,k}}{R_{n-1,k}} = \frac{S_{n-1,k} \cdot B_{n,k} \cdot H_{n,k} + N_{n,k}}{S_{n-1,k} \cdot H_{n-1,k} + N_{n-1,k}} \quad (15)$$

$$\hat{B}_{n,k} = \text{dec} \{D_{n,k}^{\text{nc}}\}. \quad (16)$$

In general, successive channel transfer factors are strongly correlated so that $H_{n,k} \approx H_{n-1,k}$ and therefore cancel down in (15) (if the noise influence is neglected). Unfortunately, the $D_{n,k}^{\text{nc}}$ is affected by twice the noise power than $D_{n,k}^c$ leading to a higher BER than coherent demodulation (with perfect channel state information) does. Note that with noncoherent demodulation, no channel estimation has to be performed. Thus, the computation complexity in the receiver is relatively low.

3) *Quasi-Coherent Demodulation*: Quasi-coherent demodulation is another way to demodulate differential modulation. Similar to coherent demodulation, the channel

influence is first removed before differential decoding takes place. Due to differential encoding/decoding, there is no need to determine the channel phase exactly but only up to an ambiguity of $2\pi/N_p$ rad (N_p = number of phases). In this case, no pilot symbols are required to estimate the unknown channel transfer factors [30].

Apart from differential decoding the processing in a quasi-coherent receiver is similar to that in a coherent one, i.e.,

$$\begin{aligned} \hat{B}_{n,k} &= \frac{\hat{S}_{n,k}}{\hat{S}_{n-1,k}} = \frac{\text{dec}(R_{n,k}/\hat{H}_{n,k})}{\text{dec}(R_{n-1,k}/\hat{H}_{n-1,k})} \\ &= \frac{\text{dec}(D_{n,k}^c)}{\text{dec}(D_{n-1,k}^c)}. \end{aligned} \quad (17)$$

In case of an incorrect decision this error influences two successive symbols due to differential encoding. Therefore, the error probability with quasi-coherent demodulation is approximately twice as high as with coherent detection, provided that the error rate is relatively small ($<10\%$). But doubling the error rate with respect to coherent demodulation results in a smaller SNR loss than noncoherent demodulation does (at least in case of perfect channel estimation).

C. Multilevel Differential Modulation (DAPSK)

In order to increase the bandwidth efficiency, the well-known M -DPSK can be extended to a differential amplitude and phase modulation (M -DAPSK), which shows a substantial performance improvement over M -DPSK for $M \geq 16$.

1) *Modulation*: DAPSK can be described as a differentially encoded APSK the signal-space diagram of which is defined by the signal set

$$\begin{aligned} \Psi &= \{a^A \cdot e^{j\Delta\varphi \cdot P} | A \in \{0, \dots, N_a - 1\}, \\ &P \in \{0, \dots, N_p - 1\}\}. \end{aligned} \quad (18)$$

As an example, a 64-APSK signal space diagram with $N_a = 4$ and $N_p = 16$ is depicted in Fig. 3. Note that the amplitudes are spaced by a factor a . The mapping of the m input bits is done separately for amplitude and phase using m_a and m_p bits, respectively. To minimize the BER, Gray coding should be used.

The number of amplitude circles $N_a = 2^{m_a}$, the number of phases per amplitude circle $N_p = 2^{m_p}$, and the amplitude ratio a are free parameters which have to be optimized in dependency of the number of signal states $M = N_a \cdot N_p = 2^m$ and the demodulation method (quasi-coherent or noncoherent). The optimized values for different number of bits per symbol are given in Table 1.

Differential encoding of the APSK symbols $Q_{n,k}$ can be performed as follows:

$$S_{n,k} = Q_{n,k} \odot S_{n-1,k} \quad (19)$$

$$\begin{aligned} &= a^{[A(Q_{n,k}) + A(S_{n-1,k})] \bmod N_a} \\ &\quad \cdot e^{j(2\pi/N_p)[P(Q_{n,k}) + P(S_{n-1,k})]}. \end{aligned} \quad (20)$$

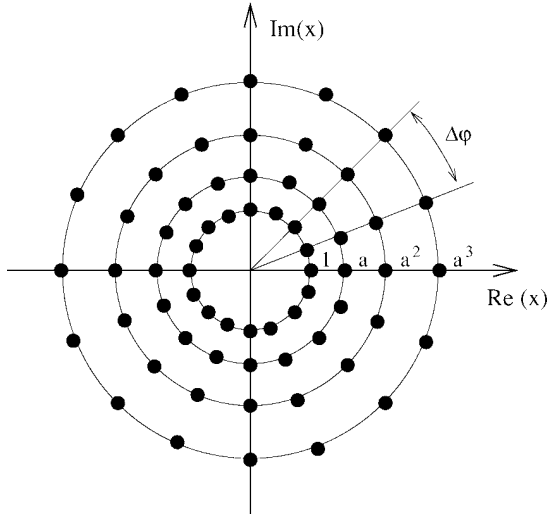


Fig. 3. A 64-(D)APSK signal space diagram.

Table 1
Optimal Modulation Parameters for Quasi-Coherent
and Noncoherent Demodulation

	N_a	N_p	a (inc. dem.)	a (quasi-coh. dem.)
$M < 16$	1	M	-	-
$M = 16$	2	8	2.0	1.8
$M = 32$	2	16	1.6	1.45
$M = 64$	4	16	1.4	1.38
$M = 128$	4	32	1.3	1.21

To assure that $S_{n,k} \in \Psi$, the amplitude exponents $A(Q_{n,k})$ and $A(S_{n-1,k})$ are added modulo the number N_a of amplitude circles. This is not explicitly necessary for the phase due to its inherent periodicity. Alternatively, the modulo operation can be integrated into the amplitude mapping (see Fig. 4) so that differential encoding can again be described by a multiplication according to (12)

$$S_{n,k} = B_{n,k} \cdot S_{n-1,k} \quad (21)$$

where

$$B_{n,k} \in \Psi' = \left\{ a^{A'} \cdot e^{j\Delta\varphi P} \mid \begin{array}{l} A' \in \{-N_a+1, \dots, N_a-1\} \\ P \in \{0, \dots, N_p-1\} \end{array} \right\}. \quad (22)$$

As an example, this modified mapping is depicted in Table 2 for $m_a = 2$.

2) *Demodulation*: Hard output demodulation of DAPSK signals requires thresholds to be defined for the decision variable $D_{n,k}^{\text{nc}}$ (noncoherent demodulation) and $D_{n,k}^{\text{qc}}$ (quasi-coherent demodulation). Since the bits are independently mapped to amplitude and phase, it is sufficient to define corresponding amplitude and phase thresholds. In case of quasi-coherent demodulation with independent identically distributed symbols, it is well known that the optimum threshold is given by half the distance between two adjacent

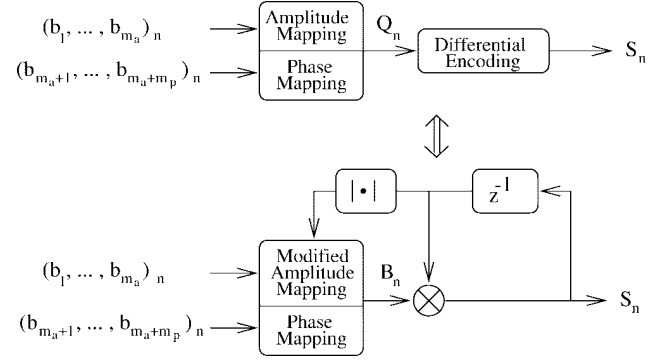


Fig. 4. DAPSK modulation.

Table 2
Modified Amplitude Mapping in Case of Four Amplitude Circles

$ B_{n,k} $	$(b_1 b_2)_{n,k}$			
	00	01	11	10
$ S_{n-1,k} $	1	a	a^2	a^3
	a^1	a	a^2	$1/a$
	a^2	a	$1/a^2$	$1/a$
	a^3	$1/a^3$	$1/a^2$	$1/a$

valid phases or amplitudes

$$T_p^c = \Delta\varphi/2 + i\Delta\varphi \quad i = 0, \dots, N_p - 1 \quad (23)$$

$$T_a^c = a^j \cdot \frac{1+a}{2} \quad j = 0, \dots, N_a - 2. \quad (24)$$

In case of noncoherent demodulation, the same thresholds are valid for the phase, i.e., $T_p^{\text{nc}} = T_p^c$ but not for the amplitude. Due to the division of two successively received noise symbols [see (15)], the exact thresholds are hard to calculate. But there exists a good approximation which is given by the geometric average of two adjacent valid amplitudes [11]

$$T_a^{\text{nc}} \approx a^j \cdot \sqrt{a} \quad j = 0, \dots, N_a - 2 \quad (25)$$

IV. CHANNEL CODING

As it has already been explained, the radio channel attenuates each OFDM subcarrier by a complex transfer factor $H_{n,k}$. If the channel is a multipath channel with many propagation paths and without a line-of-sight (LOS) path, then the amplitude of the transfer factors is Rayleigh distributed according to the central limit theorem.

This means that even at a very large average SNR there are always some subcarriers which are strongly attenuated and have a rather low SNR. Consequently, many bit errors occur on these subcarriers. The situation is similar to a narrow-band fading channel which is not surprising, since an OFDM transmission system consists of many narrowband fading channels.

For this reason, channel coding is a very important topic for OFDM systems. Without channel coding the typical flat fading curve is obtained for the BER. With channel coding, large SNR gains are achieved. The code provides diversity

in this situation. The additional gain which can be realized with soft decision instead of hard decision decoding is very large.

Therefore, convolutional codes are a reasonable choice, because soft decision decoding can easily be performed by the Viterbi algorithm which is usually applied in order to decode convolutional codes. Furthermore, the code rate of convolutional codes can be adjusted in a very flexible way by puncturing the code. This can be applied in order to adapt the error correcting capability of the code to the channel state.

A. Soft Output Demodulation

In order to apply soft-decision decoding, the decoder needs metric increments for the received modulation symbols instead of ready decisions. This means the demodulator has to provide soft outputs $\lambda(S_{n,k}^\nu)$ for each symbol $R_{n,k}$ where the superscript ν enumerates the (possibly transmitted) modulation symbols in the set Ψ .

1) *Coherent*: As it has already been explained, the received modulation symbol of the k th OFDM subcarrier in the n th time interval is

$$R_{n,k} = H_{n,k}S_{n,k} + N_{n,k} \quad (26)$$

where $S_{n,k}$ is the transmitted symbol, $H_{n,k}$ is the complex channel transfer factor of the considered subcarrier, and $N_{n,k}$ is additive white Gaussian noise with the noise power $2\sigma^2$. The probability density function (pdf) of $R_{n,k}$ is a complex Gaussian distribution

$$p(R_{n,k}|S_{n,k}) = \frac{1}{2\pi\sigma^2} e^{-(|R_{n,k} - H_{n,k}S_{n,k}|^2/2\sigma^2)}. \quad (27)$$

A maximum likelihood sequence estimator has to determine an estimate $\langle \widehat{S}_{n,k} \rangle$ out of all possibly transmitted symbol sequences $\langle S_{n,k} \rangle^{(\mu)}$ according to the following criterion:

$$\langle \widehat{S}_{n,k} \rangle = \arg \max_{\mu} P(\langle R_{n,k} \rangle | \langle S_{n,k} \rangle^{(\mu)}) \quad (28)$$

$$= \arg \min_{\mu} \sum_k |R_{n,k} - H_{n,k}S_{n,k}(\mu)|^2. \quad (29)$$

The summands in (29) are calculated by the demodulator for each possibly transmitted modulation symbol $S_{n,k}^\nu$ with $\nu = 1, \dots, M$ and passed to the decoder as metric increments

$$\lambda(S_{n,k}) = |R_{n,k} - H_{n,k}S_{n,k}|^2 \quad (30)$$

for further evaluation and final decision. The decoder can estimate the symbol sequence or the corresponding bit sequence on the basis of the given metric increments according to (29).

This metric consisting of squared Euclidean distances as metric increments can be used with all coherent demodulation techniques because it is only based on the assumptions that the transmitted modulation states are statistically independent and that the received modulation states are Gaussian distributed due to the AWGN.

2) *Noncoherent*: In the case of differential modulation techniques soft outputs are calculated in a very similar way. Decision variables are chosen having a pdf which can be approximated by a Gaussian distribution and determine the variance of this distribution. Then the demodulator generates a metric consisting of squared distances with these appropriate decision variables weighted according to the determined variance.

For higher level M -DPSK ($M \geq 8$) the pdf of the phase difference $\varphi_{n,k} = \arg D_{n,k}^{nc}$ of successive symbols can already be approximated by a Gaussian distribution

$$p(\varphi_{n,k}|\psi_{n,k}) \approx \frac{1}{\sqrt{2\pi\sigma_\varphi^2}} e^{-(|\varphi_{n,k} - \psi_{n,k}|^2/2\sigma_\varphi^2)} \quad (31)$$

where $\psi_{n,k}$ is the transmitted and $\varphi_{n,k}$ the received phase difference. This results in the following metric increments for M -DPSK symbols [19]:

$$\lambda(\psi_{n,k}) = \arg \min_{\mu} \sum_k |H_{n,k}|^2 |\varphi_{n,k} - \psi_{n,k}|^2. \quad (32)$$

DPSK consists of a DPSK and a differential amplitude modulation. The amplitude information is contained in the quotient $B_{n,k}$ of the amplitudes of two successive modulation symbols which is the usual decision variable. For the purpose of generating a metric for the Viterbi algorithm, new decision variable $W_{n,k}$ is used instead of $B_{n,k}$. $W_{n,k}$ is approximately Gaussian distributed again

$$W_{n,k} = \ln \left| \frac{R_{n,k}}{R_{n-1,k}} \right|, \quad V_{n,k} = \ln \left| \frac{S_{n,k}}{S_{n-1,k}} \right| \quad (33)$$

$$p(W_{n,k}|V_{n,k}) \approx \frac{1}{\sqrt{2\pi\sigma_w^2}} e^{-(|W_{n,k} - V_{n,k}|^2/2\sigma_w^2)} \quad (34)$$

$$\sigma_w^2 = \sigma_\varphi^2 = \frac{\sigma^2}{|H_{n,k}S_{n,k}|^2} + \frac{\sigma^2}{|H_{n-1,k}S_{n-1,k}|^2}. \quad (35)$$

Using the decision variables $W_{n,k}$ and $\varphi_{n,k}$ a metric with squared distances can also be given for DAPSK modulation [19]. The transmitted symbols $B_{n,k}$ are described by $V_{n,k}$ and $\psi_{n,k}$

$$\lambda(B_{n,k}) = \arg \min_{\mu} \sum_k d_{n,k}^2 RI_{n,k}^2 \quad (36)$$

$$d_{n,k} = (W_{n,k} - V_{n,k})^2 + (\varphi_{n,k} - \psi_{n,k})^2 \quad (37)$$

$$RI_{n,k}^2 = \frac{1}{|1/R_{n,k}|^2 + |1/R_{n-1,k}|^2}. \quad (38)$$

By the term $d_{n,k}^\nu$ the position of the received symbols in the constellation diagram is considered. $RI_{n,k}$ is an information about the reliability of the subcarrier (reliability information).

3) *Quasi-Coherent*: In the case of quasi-coherent detection, the two successively received modulation symbols are evaluated separately in order to calculate the metric increments. The Viterbi decoder uses the following metric:

$$\lambda(B_{n,k}) = \min_{(S^\kappa/S^\rho)=B_{n,k}} [|R_{n,k} - H_{n,k}S^\kappa|^2 + |R_{n-1,k}H_{n-1,k}S^\rho|^2]. \quad (39)$$

Again, all possible metric increments $\lambda_{n,k}^\nu$ can be calculated by the demodulator and passed to the decoder.

B. Bit Metric

As it has been described above, metric increments are calculated for each modulation symbol $S_{n,k}^\nu$ or $B_{n,k}^\nu$ by the demodulator. The task of the decoder is to decide for a bit sequence $\langle b_{n,k,j} \rangle$ which leads to the minimal metric. The subscript $j = 1, \dots, m$ enumerates the bits which are transmitted by a modulation symbol. If the Viterbi algorithm is to be used for decoding, metric increments $\lambda_{n,k,j}$ for each bit are required instead of metric increments for each symbol in order to make the following decision:

$$\langle \widehat{b_{n,k,j}} \rangle = \arg \max_{\mu} P(\langle R_{n,k} \rangle | \langle b_{n,k,j} \rangle^{(\mu)}) \quad (40)$$

$$\approx \arg \min_{\mu} \sum_{k,j} \lambda_{n,k,j}(\mu). \quad (41)$$

A metric increment $\lambda(b_{n,k,j})$ for the bit $b_{n,k,j} \in \{-1, 1\}$ can be obtained by

$$\lambda(b_{n,k,j}) = \min_{S_{n,k} | b_{n,k,j}} \lambda(S_{n,k}). \quad (42)$$

In the case of differential modulation, $B_{n,k}^\nu$ replaces $S_{n,k}^\nu$ in (42). The Viterbi decoder chooses the bit sequence $\langle b_{n,k,j} \rangle$ which minimizes the sum in (41). For all bits $b_{n,k,j}$ of a subcarrier k , the corresponding metric increments are determined on the basis of the modulation symbol with the highest conditional probability $P(R_{n,k} | S_{n,k})$ and by which the assumed bit (“−1” or “1”) is transferred.

This way of calculating and using a bit metric for the Viterbi algorithm is an approximation. For an optimal decision the metric increments for all bits in the bit sequence which are associated with the same subcarrier would have to refer to the same modulation symbol.

C. Multilevel Coding

In the discussion above, modulation and channel coding have been considered separately. A joint consideration of coding and modulation can achieve a better performance by taking advantage of the fact that different error events occur with very different probabilities. A very general approach of combining coding and modulation is multilevel coding [16].

If a modulation scheme with m bits/symbol is used, the data bits are encoded into m sequences of l bits by m different encoders. These bit sequences are mapped to l modulation symbols.

The set of all modulation symbols is partitioned into two subsets such that the error probability within each subset is minimized. In the case of nondifferential modulation this means that the Euclidean distance of the modulation symbols in each subset must be maximized. Subsequently, each of the subsets is recursively partitioned again until the subsets contain only a single modulation symbol. An example for this set partitioning is given in Fig. 5 for 16-DAPSK.

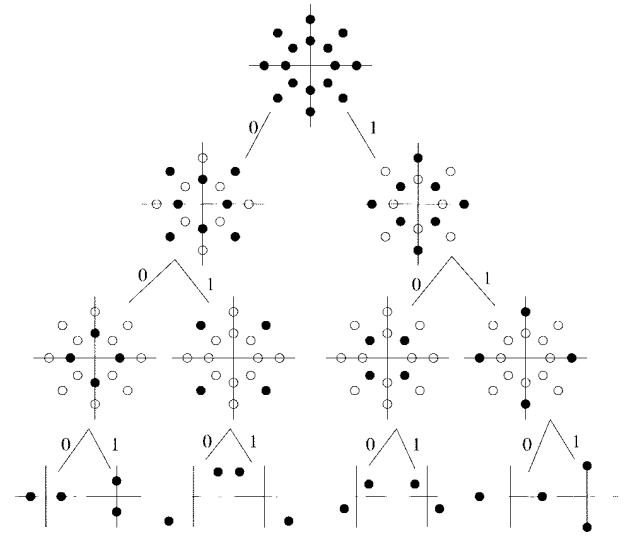


Fig. 5. Set partitioning for 16-DAPSK.

Based on this set partitioning, the data bits are mapped to modulation symbols in the following way. One data bit is taken from each of the m coded sequences. The first data bit selects one of the two subsets of the first partitioning level. Each of the following data bits selects one of the two subsets of the next partitioning level until the modulation symbol to be transmitted is defined after m selections.

The receiver evaluates the transmitted modulation symbols and subsequently determines the coded bit sequences. Provided that the receiver is able to decode the first bit sequence which corresponds to the first partitioning level without errors, this information can be used in order to evaluate the second partitioning level and so on. As the error probability within the subsets decreases from level to level, the codes need less redundancy in order to correct errors.

For differential modulation, the set of all possible quotients $B_{n,k} = S_{n,k}/S_{n-1,k}$ has to be partitioned so that the error probability within each subset is maximized. In the case of M -DPSK this leads to the same set partitioning which is known for M -PSK. For M -DAPSK the set partitioning is discussed in the following.

We assume that the M -DAPSK symbol B_1 with logarithm of amplitude V_1 and phase difference ψ_1 is transmitted, that the symbol B_2 with V_2 and ψ_2 belongs to the same subset as B_1 , and that the decisions variables which the receiver calculates from the received symbol B' are $W = V_1 + \Delta V$ and $\varphi = \psi_1 + \Delta\psi$.

The probability density functions of W and φ are approximately Gaussian as it has been shown above. The approximated joint pdf of W and φ is

$$\begin{aligned} p(W, \varphi | V_1, \psi_1) &\approx \frac{1}{2\pi\sigma_w^2} e^{-(1/2\sigma_w^2)(|W-V_1|^2 + |\varphi-\psi_1|^2)} \\ &= \frac{1}{2\pi\sigma_w^2} e^{-(1/2\pi\sigma_w^2)(|\Delta V|^2 + |\Delta\psi|^2)}. \end{aligned} \quad (43)$$

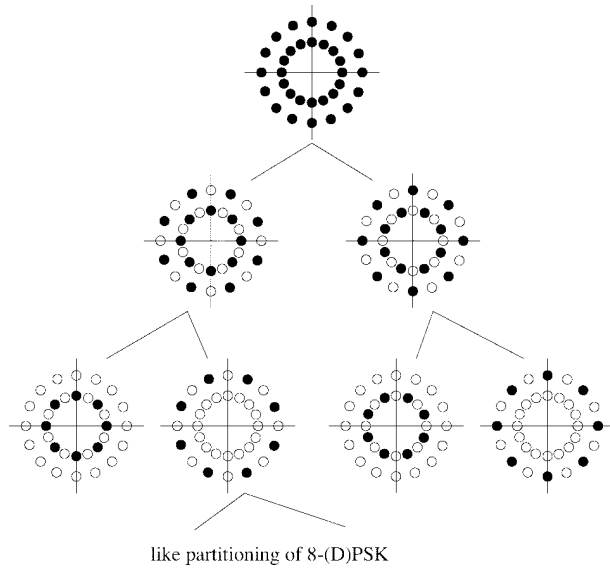


Fig. 6. Set partitioning of 32-DAPSK.

The receiver decides incorrectly, if

$$p(W, \varphi | V_2, \psi_2) > p(W, \varphi | V_1, \psi_1) \quad (44)$$

$$\Rightarrow (W - V_2)^2 + (\varphi - \psi_2)^2 < (W - V_1)^2 + (\varphi - \psi_1)^2. \quad (45)$$

The decision variables W and φ make the situation very similar to the partitioning of a non-differential modulation scheme like M -QAM, where the Euclidean distance within the subsets is maximized. Here, the error probability within a subset is minimized if the smallest distance $d_{i,j}$ between two symbols B_i and B_j in the subset is maximized

$$d_{i,j}^2 = (V_i - V_j)^2 + (\psi_i - \psi_j)^2. \quad (46)$$

The distance measure $d_{i,j}$ is nothing else but a Euclidean distance referring to the coordinates V and ψ .

The optimal set partitioning is given for 16, 32, and 64-DAPSK in Figs. 5–7. The partitioning diagrams refer to the Q symbols (see Section III-C) which represent the information to be transmitted.

V. SYSTEM PERFORMANCE

A. Simulation

The baseband part of a general OFDM transmission system with equalization and channel coding is depicted in Fig. 1. In this paper we consider OFDM systems with 256 subcarriers, a total bandwidth of 7.16 MHz, and a guard interval of 7 μ s. The net data rates are 18 and 27 Mbit/s for 16-QAM/DAPSK and 64-QAM/DAPSK, respectively.

For each OFDM block the sequence of data bits to be transmitted is encoded by a convolutional code with memory 6 from [33]. The coded bits are interleaved by a block interleaver in order to avoid burst errors which would affect the performance of a convolutional code.

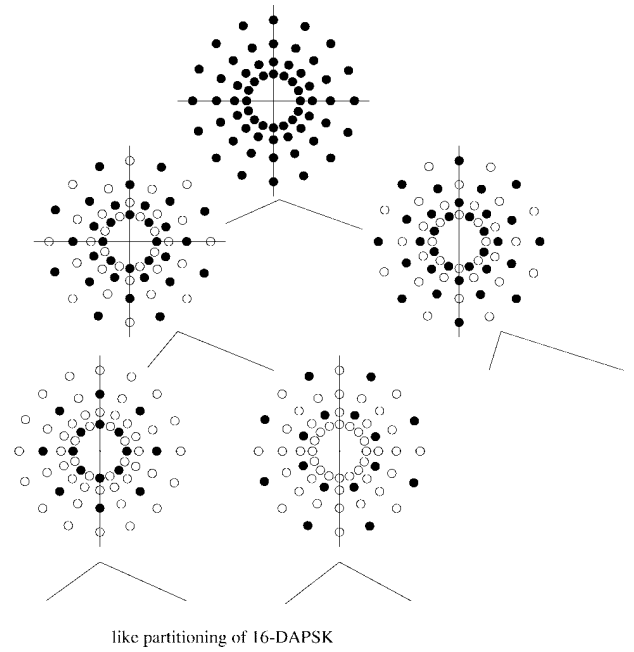


Fig. 7. Set partitioning of 64-DAPSK.

The modulation takes place in frequency direction as a mapping of the coded and interleaved data bits to a sequence of modulation symbols with or without differential encoding. In this paper, differential modulation is always performed in time direction so that each modulation symbol $S_{n,k}$ refers to the previously transmitted symbol $S_{n-1,k}$ of the same OFDM subcarrier. In the next step the IFFT is performed in order to obtain the time signal and the guard interval is added as a cyclic prefix.

This signal is transmitted over the frequency-selective radio channel. The channel has been simulated according to the WSSUS channel model described in [15] with the power delay profile “Typical Urban” from [9] (maximum delay $\tau = 7 \mu$ s), not time varying.

At the receiver the guard interval is removed and the signal is transformed back into the frequency domain using an FFT. If coherent modulation has been used, an ideal channel estimation is provided by the simulation program. Practically, a channel estimation based on pilot symbols would have to be performed. The demodulator detects the received modulation states $R_{n,k}$ of the subcarriers and generates metric increments for the soft decision Viterbi decoder. The metric increments delivered by the demodulator are deinterleaved. Finally the Viterbi decoder determines the original data bits.

If a nondifferential modulation scheme (PSK or QAM) is used, a practical transmission system would have to use some of the subcarriers for the transmission of pilot symbols for the channel estimation. This causes both an additional overhead and a system degradation compared to the ideal channel estimation. In literature [14], [27], it has been shown that the loss due to an imperfect channel estimation is approximately 1.5 dB.

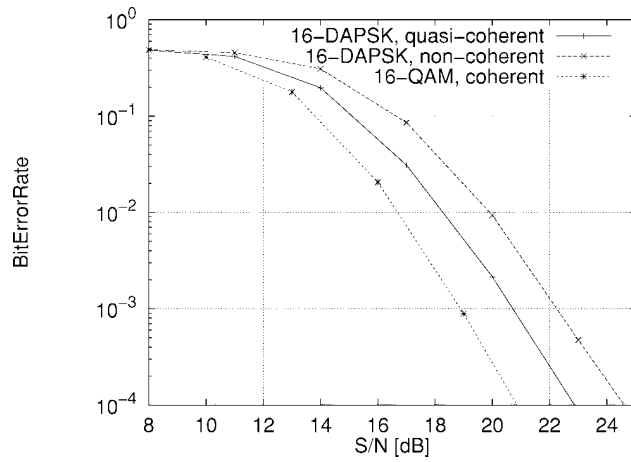


Fig. 8. Comparison of 16-DAPSK and 16-QAM.

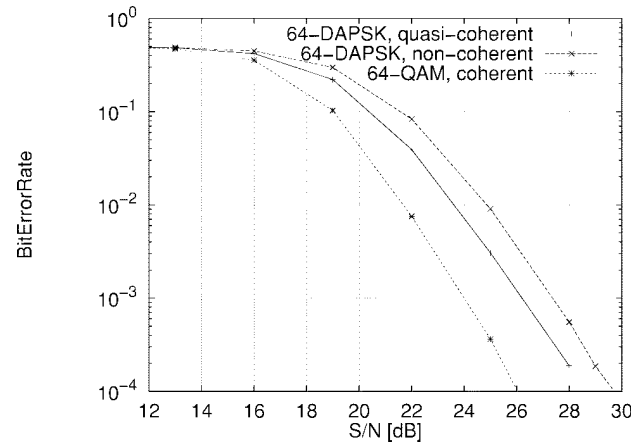


Fig. 9. Comparison of 64-DAPSK and 64-QAM.

If differential modulation is considered, pilot signals are not necessary so that less redundancy has to be transmitted in this case. In order to make the situation comparable with systems that employ nondifferential modulation, we use a convolutional code with increased redundancy instead of the pilot symbols. The code rates are chosen in such a way that the same user data rate is transmitted in the systems which are compared. Such a comparison with a fixed data rate seems to be reasonable. If, for example, 6% of the subcarriers are needed for the transmission of pilot signals in the case of M -QAM as a nondifferential modulation scheme, the code rate $R = 4/5$ can be used for M -QAM and the code rate $R = 3/4$ for M -DAPSK. The performance of these systems can be compared reasonably.

B. Results

Simulation results are shown in Fig. 8 for 16-DAPSK, code rate $R = 3/4$, with noncoherent and quasi-coherent demodulation and for 16-QAM, $R = 4/5$. The curve for 16-QAM has been obtained with a perfect channel estimation. In Fig. 9 results for 64-DAPSK and 64-QAM are given.

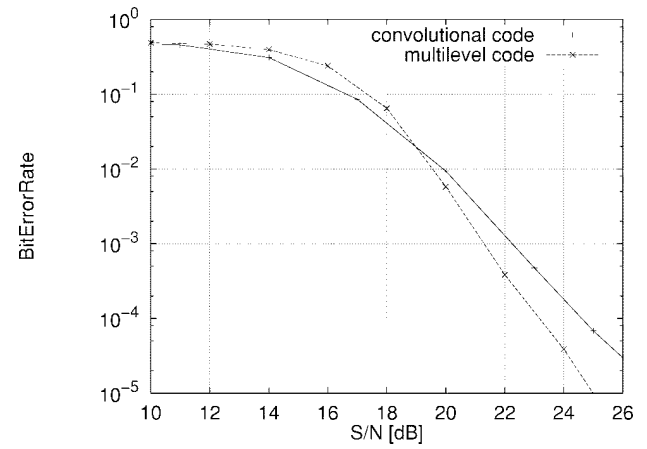


Fig. 10. 16-DAPSK with convolutional code $R = 3/4$ and with multilevel code with rates $1/2$, $2/3$, $5/6$, $11/12$.

With multilevel coding, an additional coding gain can be obtained. An example for 16-DAPSK is depicted in Fig. 10, where a convolutional code with code rate $R = 3/4$ is compared with a multilevel coding schemes based on convolutional codes with the code rates $1/2$, $2/3$, $5/6$, and $11/12$ for the four bit levels. The overall code rate of the multilevel code is 0.73.

VI. CONCLUSION

In this paper convolutionally coded OFDM transmission systems with coherent and noncoherent detection have been analyzed which are suited for the transmission of high data rates for multimedia applications over a frequency-selective radio channel.

The main difference between coherent and noncoherent systems is the channel estimation which is required for the coherent system. This channel estimation is an additional effort which is made in order to obtain a performance gain over the differential systems. However, the channel estimation requires the transmission of pilot symbols which can be replaced by additional code redundancy in the case of differential modulation. Additionally, a loss of 1.5 dB due to a real channel estimation has to be taken into account. Based on these assumptions the remaining performance gain of the coherent system is rather small.

The decision as to which modulation scheme is to be preferred depends on the properties of the considered radio channel. If only 6% pilot symbols are sufficient for adaptation to the frequency-selective and time-variant channel, the performance of 64-QAM compared with 64-DAPSK as an example may justify the additional effort for channel estimation. But if more pilot symbols are needed in order to achieve a stable transmission over a time-varying channel, like in [4], it is preferable to decide for differential modulation.

The BER performance of the transmission systems can be further improved by multilevel coding. The set partitioning for DAPSK which is required for multilevel coding has

been given in this paper. The improvement compared with a simple convolutional code is about 1.5 dB.

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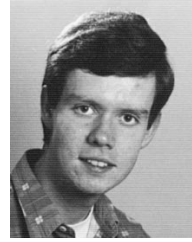
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