INVITED PAPER

Near-Instantaneously Adaptive HSDPA-Style OFDM Versus MC-CDMA Transceivers for WIFI, WIMAX, and Next-Generation Cellular Systems

Various techniques that are designed to allow future wireless systems to adapt very rapidly to changing conditions are studied and compared.

By Lajos Hanzo, *Fellow IEEE*, and Byoung-Jo Choi

ABSTRACT | Burts-by-burst (BbB) adaptive high-speed downlink packet access (HSDPA) style multicarrier systems are reviewed, identifying their most critical design aspects. These systems exhibit numerous attractive features, rendering them eminently eligible for employment in next-generation wireless systems. It is argued that BbB-adaptive or symbol-by-symbol adaptive orthogonal frequency division multiplex (OFDM) modems counteract the near instantaneous channel quality variations and hence attain an increased throughput or robustness in comparison to their fixed-mode counterparts. Although they act quite differently, various diversity techniques, such as Rake receivers and space-time block coding (STBC) are also capable of mitigating the channel quality variations in their effort to reduce the bit error ratio (BER), provided that the individual antenna elements experience independent fading. By contrast, in the presence of correlated fading imposed by shadowing or time-variant multiuser interference, the benefits of space-time coding erode and it is unrealistic to expect that a fixed-mode space-time coded system remains capable of maintaining a near-constant BER. Following a basic portrayal of adaptive OFDM, the paper investigates a combined system constituted by a constant-power adaptive modem employing space-time coded diversity techniques in the context of both OFDM and multicarrier code division multiple access (MC-CDMA). The combined system is configured to produce a constant uncoded BER and exhibits virtually error-free performance when a turbo convolutional code is concatenated with a space-time block code. Again, it was found that the advantage of rendering the multiple-input multiple-output (MIMO)-aided modem BbB-adaptive erodes, when the affordable system complexity facilitates the implementation of multiple transmitters and receivers, if it can be ensured that no shadowing and no multiuser interference fluctuation is experienced. By contrast, in the presence of these impairments, only BbB-adaptive transceivers can provide a near-constant BER.

KEYWORDS | Adaptive modulation; adaptive multicarrier code division multiple access (MC-CDMA); adaptive orthogonal frequency division multiplexing (OFDM); high-speed data packet access (HSDPA)

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L. Hanzo is with the School of Electrical and Computer Science (ECS), University of Southampton, SO17 1BJ Southampton, U.K. (e-mail: lh@ecs.soton.ac.uk; http://www-mobile.ecs.soton.ac.uk).
B.-J. Choi is with the Multimedia Systems Engineering Department, University of Incheon, Incheon 402-749, Korea.

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I. MOTIVATION

While the rollout of the third-generation (3G) systems [1] continues, the academic and industrial research community turned their attention to the investigation of next-generation multicarrier transceiver techniques [2]–[6]. The next-generation wireless systems are expected to support both variable-rate as well as extremely high-bit-rate services in a wide range of different propagation environments. Under these propagation conditions it is unrealistic to expect that conventional fixed-mode transceivers might be capable of supporting a time-invariant wireline-like quality of service and hence near-instantaneously adaptive or burst-by-burst (BbB) adaptive transceiver techniques [4]–[8] have found their way into the high-speed data packet access (HSDPA) mode of the 3G systems. Multistandard operation is also a salient requirement. As it was argued in the context of the generic future-proof system design framework outlined in [4, Ch. 1], multicarrier (MC) transmission techniques [4], [5], [9], such as orthogonal frequency division multiplexing (OFDM) [5] and its frequency-domain spreading aided version, namely MC code division multiple access (MC-CDMA) exhibit a high number of reconfigurable parameters. Another attractive relative of this MC transceiver family is constituted by direct sequence (DS) MC-CDMA as well as its space-time-spreading assisted counterparts [4], [10], [11]. Both time- and frequency-domain spread MC-CDMA was detailed in [12], while the choice of interference-mitigating spreading code design was discussed in [13] and that of bandlimited waveforms in [15]. Since DS MC-CDMA was documented in [4], [10], and [11], following a rudimentary introduction to the subject and a brief tour of the MC transceiver history, this overview article will consider a number of design aspects pertaining to OFDM and frequency-domain spreading aided MC communications.

The brief outline of the paper is as follows. In Section II the basic OFDM principles are introduced, followed by a conceptual introduction to near-instantaneously adaptive OFDM in Section III. Following this rudimentary introduction, in Section IV a more detailed, but mainly conceptual literature-based historical perspective of the associated research problems and their evolution is provided. Finally, the most detailed technical discussions are provided in Section V, where multiple-input multiple-output (MIMO)-aided adaptive OFDM and MC-CDMA are studied comparatively.

II. OFDM BASICS

In this introductory section, we examine OFDM as a means of counteracting the channel-induced linear distortions encountered when transmitting over a dispersive radio channel. The fundamental principle of orthogonal multiplexing originates from Chang [16], and over the years a number of researchers have investigated this technique [17]–[28]. Despite its conceptual elegance, until the 1990s its employment has been mostly limited to military applications due to implementational difficulties. However, it has been adopted as the digital audio broadcast (DAB) as well as digital video broadcast (DVB) standard and also for various wireless local area network (WLAN) standards. Finally, it is also advocated by the 3G Partnership Project (3GPP) for the long-term evolution of the 3G systems. These consumer electronics applications underline its significance [29]–[32].

In the OFDM scheme of Fig. 1 the serial data stream of a channel is passed through a serial-to-parallel convertor, which splits the data into a number of parallel channels. The data in each channel is applied to a modulator, so that for N channels there are N modulators whose carrier frequencies are $f_0, f_1, \ldots, f_{N-1}$. The frequency difference between adjacent channels is $\Delta f = f_0$ and the overall bandwidth W of the N modulated subcarriers is $N\Delta f = N \cdot f_0$.

These N modulated carriers are then combined, in order to generate the OFDM signal. We may view the serial-to-parallel (SP) convertor as applying every Nth symbol to a modulator. This has the effect of interleaving the symbols entered into each modulator, e.g., symbols $S_0, S_N, S_{2N}, \ldots$ are applied to the modulator whose carrier frequency is f_1. At the receiver, the received OFDM signal is demultiplexed into N frequency bands, and the N modulated signals are demodulated. The baseband signals are then recombined using a parallel-to-serial (PS) convertor. *The role of the SP convertor may also be viewed as that of expanding the duration of each subcarrier symbol by a factor corresponding to the number of subcarriers N. The benefit of this symbol duration expansion is that at a given maximum channel-induced dispersion, only a short fraction of the extended-length subcarrier symbol is affected by the dispersive channel when it is convolved with the channel's impulse response (CIR).*

In the more conventional serial transmission approach [6], the data is applied directly to the modulator transmitting at a carrier frequency positioned at the center of the transmission band. The modulated signal occupies the entire bandwidth W. When the data is transmitted serially, a deep fade imposed by a mobile channel inflicts a burst of transmission errors since the fades typically extend over the duration of several bits. By contrast, during an N-symbol transmission burst of the conventional serial system, each of the N number of OFDM subchannel modulators carries only one symbol, each of which has an N times longer duration. Hence, again, a fixed-duration channel fade would only affect a fraction of the duration of each of the extended-length subcarrier symbols transmitted in parallel. Therefore, the OFDM system may be able to recover all of the partially faded, extended-duration N subcarrier symbols without transmission errors. *Thus, in a somewhat simplistic, but appealingly plausible approach one might argue that while a serial system exhibits an error burst in case of a deep time-domain (TD) fade, no errors or only a few errors may occur using the OFDM approach since the subcarrier symbols are N times longer. The resultant less*

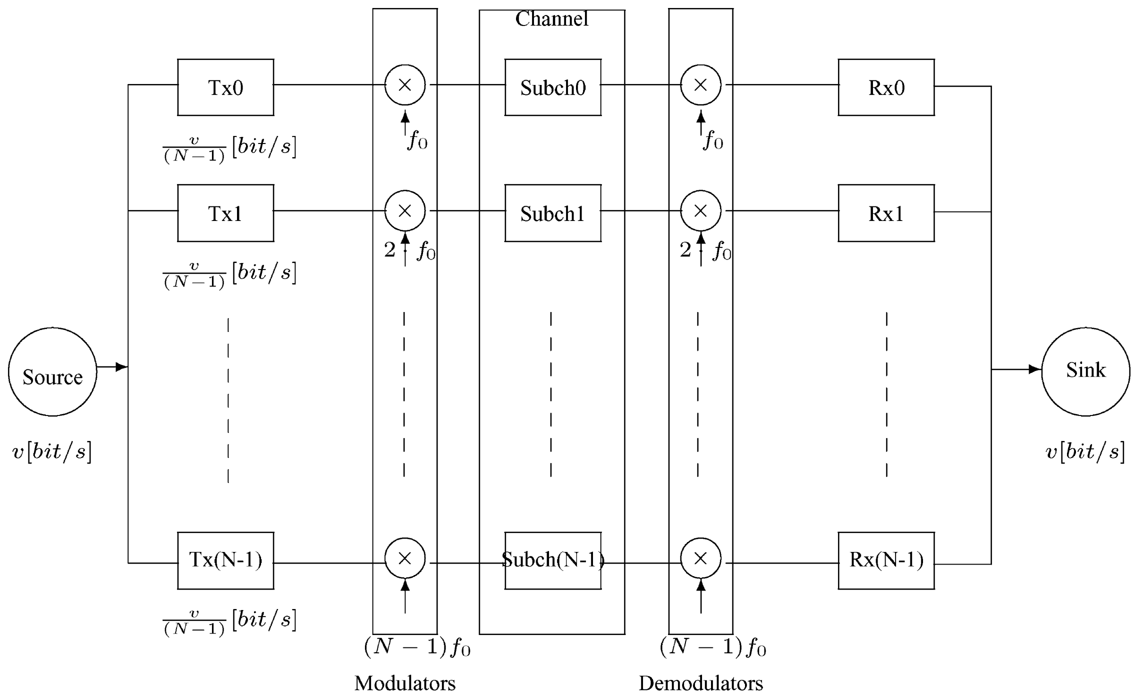


Fig. 1. ***Simplified block diagram of the orthogonal parallel modem.***

bursty and/or reduced number of errors experienced by the OFDM modem may also result in a further improved performance in conjunction with forward error correction (FEC) coding. In Sections V-C and V-D we will elaborate on this issue a little further in the context of turbo coded OFDM and MC-CDMA systems, noting that the employment of frequency-domain (FD) spreading across the subcarriers has the ability of even further randomizing the effects of transmission errors for the sake of improving the attainable FEC performance. This phenomenon may be interpreted as follows. When considering now frequency-selective fading, a deep FD fade may wipe out a specific subcarrier and hence corresponding symbol is obliterated. By contrast, in the contaxt of FD-spread MC-CDMA only a single chip may be wiped out and hence the resultant spreading code as well as the corresponding symbol may still be recovered.

A further advantage of OFDM is that because the symbol period has been increased, the channel's delay spread becomes a significantly shorter fraction of an OFDM symbol period than in the serial system, potentially rendering the system less sensitive to channel-induced dispersion, than the conventional serial system.

A disadvantage of the OFDM approach portrayed in Fig. 1 is its seemingly increased complexity in comparison to a conventional serial modem, which is a consequence of employing N number of subcarrier modulators and transmit filters at the transmitter as well as N demodulators and receive filters at the receiver.

However, as it was shown in Chapter 2 of [5], the associated complexity can be substantially reduced by employing the discrete Fourier transform (DFT) for modulating all subcarriers in a single step. From a tangible physical perspective this may be explained by arguing that all the OFDM subcarriers are orthogonal complex-valued exponential functions, which have a frequency that is an integer multiple of the basis frequency f_0, exactly as in case of the complex-valued exponential basis functions of the DFT. Hence, instead of multiplying each subcarrier individually for example by ± 1 as in binary phase-shift keying (BPSK) modulation, the modulation process implies "transforming" a block of N BPSK modulated subcarrier symbols in a single step from the subcarriers' frequency-domain to the modulated time-domain signal using the inverse DFT. This yields a block of N modulated samples, as illustrated mathematically in [5]. When the number of subcarriers is high, the system's complexity may be further reduced by implementing the DFT with the aid of the fast Fourier transform (FFT), again, as it was shown mathematically in [5].

III. ADAPTIVE OFDM BASICS

As mentioned above, a particularly attractive feature of OFDM systems is that they are capable of operating without a classic channel equalizer when communicating over dispersive transmission media, such as wireless channels, while conveniently accommodating the time- and frequency-domain channel quality fluctuations of the wireless channel.

Explicitly, the channel SNR variation versus both time and frequency of an indoor wireless channel is shown in a

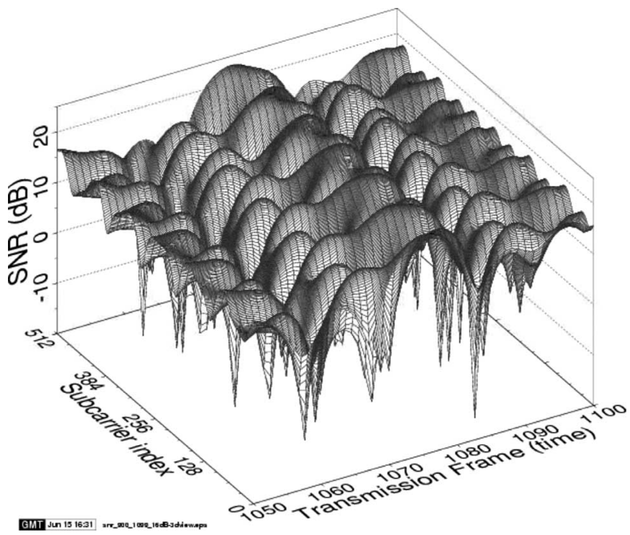


Fig. 2. ***Instantaneous channel SNR for the 512 OFDM subcarriers versus time, for an average channel SNR of 16 dB over the channel characterized by the CIR of Fig. 3 [6].***

three-dimensional form in Fig. 2 versus both time and frequency, which suggests that OFDM constitutes a convenient framework for accommodating the channel quality fluctuations of the wireless channel, as will be briefly augmented below. This frequency-domain channel transfer function (FDCHTF) was recorded for the CIR of Fig. 3, *by simply transforming the CIR to the frequency domain at regular time intervals while the CIR taps obey the Rayleigh distribution*. These channel quality fluctuations may be readily accommodated with the aid of subband-adaptive modulation as follows. Such an adaptive OFDM (AOFDM) modem is characterized by Fig. 4, portraying at the top a contour plot of the above-mentioned wireless channel's signal-to-noise ratio (SNR) fluctuation versus both time and frequency for each OFDM subcarrier. We note at this early stage that these channel quality fluctuations may be mitigated with the aid of frequency-domain channel equalization, as it was illustrated both graphically as well as mathematically in [5]. We will briefly revisit this topic also in this contribution at a later stage. More specifically, as seen in Figs. 2 and 4 when the channel is of high quality—as for example in the vicinity of the OFDM symbol index of 1080—a higher throughput may be achieved, than during the periods of lower channel quality. The average throughput of each OFDM symbol constituted by a column of 16 vertically stacked 32-subcarrier subbands of Fig. 4 was 1 bit per symbol (BPS) in this example, as in conventional BPSK.

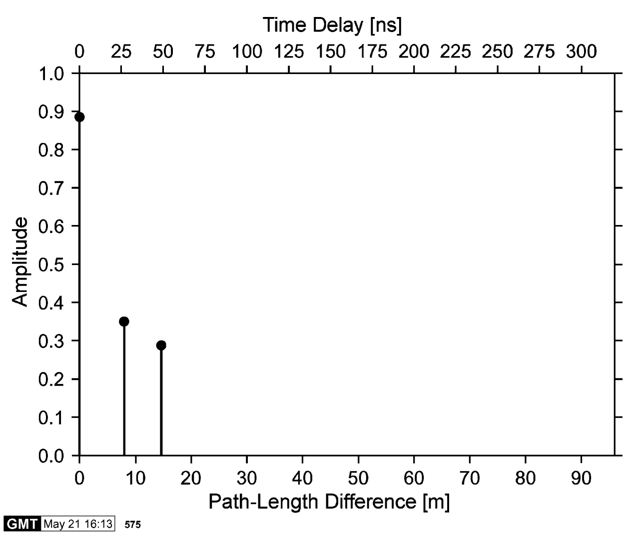


Fig. 3. ***Indoor three-path wireless asynchronous transfer mode (WATM) CIR, where the path-delays are shown both in terms of time expressed in nanoseconds and in the corresponding distance in meters.***

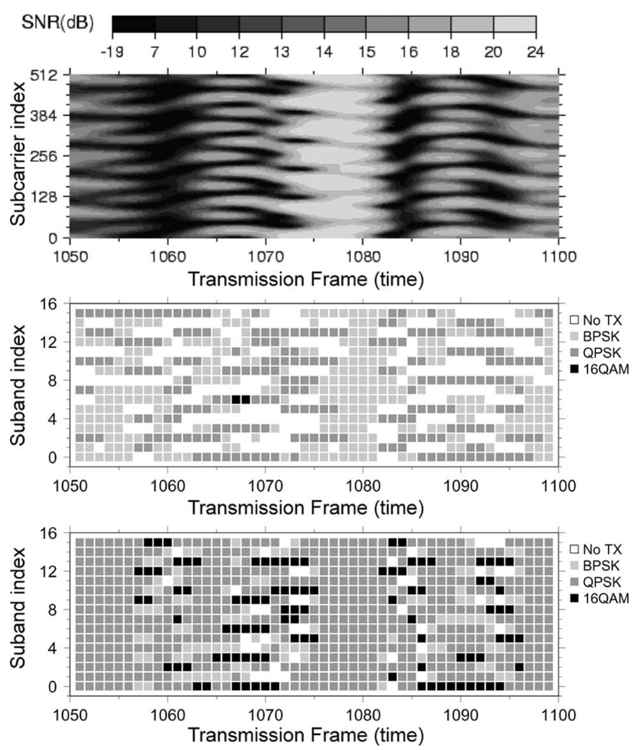


Fig. 4. ***Bit allocation of the AOFDM modem communicating over the channel having the CIR of Fig. 3 and the FDCHTF of Fig. 2. The top graph is a contour plot of the channel SNR for all 512 subcarriers versus time. The bottom two graphs show the modulation modes chosen for all 16 32-subcarrier subbands for the same period of time. The middle graph shows the performance of the 3.4 Mb/s subband-adaptive modem, which operates at the same bit rate as a fixed BPSK modem. By contrast, the bottom graph represents the 7.0 Mb/s subband-adaptive modem, which operated at the same average bit rate as a fixed QPSK modem. The average channel SNR was 16 dB. We will demonstrate in the context of Fig. 5 that naturally, at a given SNR this doubled bit rate can only be maintained at the cost of a higher BER*** **[5], [33].**

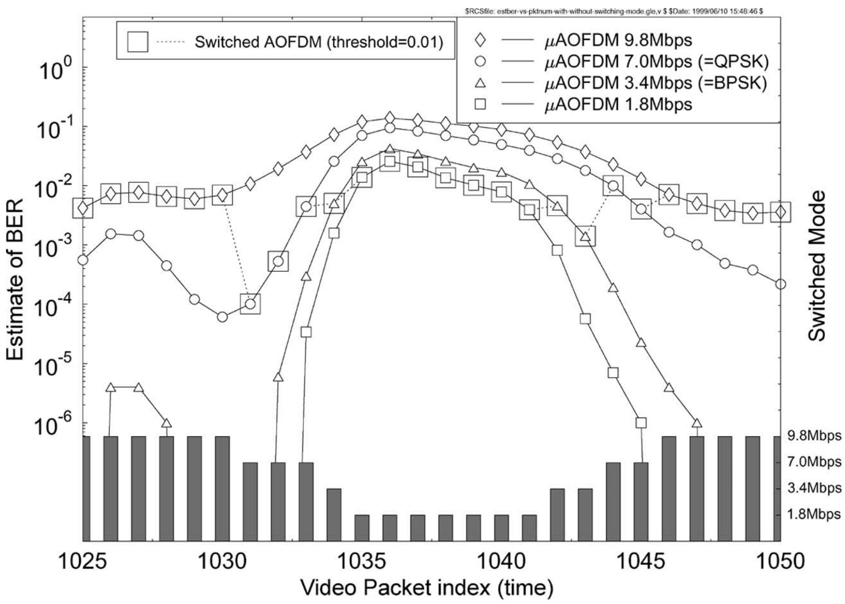


Fig. 5. ***Illustration of mode switching in the TVTBR-AOFDM modem, which we also refer to as the microadaptive OFDM (μAOFDM) scheme. The figure shows the estimate of the BER for the four possible modes. The large square and the dotted line indicate the modem mode chosen for each time interval by the mode switching algorithm. At the bottom of the graph, the bar chart specifies the bit rate of the switched subband adaptive modem on the right-hand axis versus time when using the channel model of Fig. 3 at a normalized Doppler frequency of $F_D = 7.41 \times 10^{-2}$ [35].***

More explicitly, in the center and bottom subfigures of Fig. 4 the modulation mode chosen for each 32-subcarrier subband is shown versus time for two different high-speed wireless modems communicating at either 3.4 or 7.0 Mb/s, respectively, again, corresponding to an average throughput of either 1 or 2 BPS. *We will demonstrate in the context of Fig. 5 that naturally, at a given SNR this doubled bit rate can only be maintained at the cost of a higher average BER and the achievable throughput is controlled with the aid of appropriately adjusting the modulation mode switching thresholds. Again, this basic interplay of the achievable BER, throughput and the required transmit power or SNR will become more explicit in Fig. 5.*

A. Time-Variant Target Bit-Rate AOFDM

Recall from Fig. 4 that in our previous discussions we assumed that each AOFDM symbol conveyed the same total number of bits, although it becomes explicit in Fig. 2 that the channel quality of each subcarrier fluctuates both as a function of frequency and time. This observation motivates the employment of a time variant target bit-rate (TVTBR) AOFDM scheme. In addition to the multipath-induced time-variant channel quality fluctuations the effects of shadow-fading, power control errors and multiuser interference (MUI) fluctuations—as portrayed for example in [4, p. 934, Fig. 23.2]—further motivate the employment of HSDPA-style TVTBR-AOFDM schemes.

A feasible principle is to invoke an estimate of the BER for mode switching, as follows. Since the noise energy in each subcarrier is independent of the channel's FDCHTF H_n, the SNR of subcarrier n can be expressed as $\gamma_n = |H_n|^2 \cdot \gamma$, where γ is the SNR. If no signal degradation is imposed by the inter-subcarrier interference (ICI) or MUI, then the value of γ_n determines the BER for the transmission of data symbols over the subcarrier n. Given γ_j across the N_s number of subcarriers in the jth subband, the expected overall BER of the modulation scheme M_j of subband j can be estimated, which is denoted by $\bar{p}_e(j) = 1/N_s \sum_n p_e(\gamma_j, M_j)$. For each subband, the scheme with the highest throughput, whose estimated BER is lower than a given threshold, is then chosen.

To elaborate a little further, the FDCHTF about to be experienced by the next transmitted AOFDM symbol may be assumed to be similar for the uplink (UL) and downlink (DL) of the time-division duplex (TDD) mode of the 3G systems [1] for example, since they are transmitted on the same frequency and are only slightly displaced in time. Otherwise, the FDCHTF may be estimated at the remote receiver during the previous AOFDM symbol and signaled back to the transmitter with a slight delay using high-compression vector-quantization. The impact of this delay may be eliminated by predicting the FDCHTF about to be experienced based on its past values using long-range channel prediction [34], as also discussed in this issue by Duel-Hallen.

Table 1 System Parameters for the Fixed QPSK and BPSK Transceivers, as Well as for the Corresponding Subband-Adaptive OFDM (AOFDM) Transceivers for Wireless Local Area Networks (WLANs)

	BPSK mode	QPSK mode
Packet rate	4687.5 packets/s	
FFT length	512	
OFDM symbols/packet	3	
OFDM symbol duration	2.6667μs	
OFDM time frame	80 timeslots = 213 μs	
Normalized Doppler frequency, f'_d	1.235×10^{-4}	
OFDM symbol normalised Doppler frequency, F_D	7.41×10^{-2}	
FEC coded bits/packet	1536	3072
FEC-coded video bit rate	7.2 Mbps	14.4 Mbps
Unprotected bits/packet	766	1534
Unprotected bit rate	3.6 Mbps	7.2 Mbps
Error detection CRC (bits)	16	16
Feedback error flag bits	9	9
Packet header bits/packet	11	12
Effective video bits/packet	730	1497
Effective video bit rate	3.4 Mbps	7.0 Mbps

A quadruple-mode TVTBR-AOFDM scheme using four different target bit rates is employed here for demonstrating the underlying principle, when using the systems parameters of Table 1. With the aid of the FDCHTF predictor we can then estimate the expected BER of the four possible AOFDM modes and opt for that specific AOFDM mode, whose estimated BER is below the target BER. This target BER may be varied in order to tune the behavior of the TVTBR-AOFDM scheme to the specific BER and throughput requirements. Fig. 5 demonstrates how the switching algorithm operates when configured for maintaining a 1% target BER. Specifically, the figure portrays the estimate of the BER for the four possible AOFDM modes versus time. The large square and the dotted line indicate the mode chosen for each time interval by the mode switching algorithm. The algorithm attempts to use the highest bit-rate mode, whose BER estimate is less than the target threshold, namely, 1% in this case. However, if all the four modes' estimate of the BER is above the 1% threshold, then the lowest BER mode is chosen since this will be the most robust to channel errors. An example of this is shown around AOFDM symbols or transmission frames 1035–1040. At the bottom of the graph a bar chart specifies the bit rate of the switched AOFDM modem versus time in order to emphasize when the switching occurs.

An example of the algorithm is shown in Fig. 6 when switching among the TVTBRs of 1.8, 3.4, 7, and 9.8 Mb/s. The upper part of the figure portrays the contour plot of the channel SNR for each subcarrier versus time. The lower part of the figure displays the modulation mode chosen for each 32-subcarrier subband versus time for the TVTBR-AOFDM modem. It can be seen at transmission frames 1051–1055 that all the subbands employ QPSK modulation. Therefore, the TVTBR-AOFDM modem has an instantaneous target bit rate of 7 Mb/s. As the channel degrades around frame 1060, the modem has switched to the more robust 1.8 Mb/s mode. When the channel quality is high around transmission frames 1074–1081, the highest bit-rate 9.8 Mb/s mode is used.

As the other papers in this special issue demonstrate, these adaptive transceiver principles are not limited to employment in OFDM-based multicarrier transmissions. The achievable channel capacity of adaptive transmissions employing maximal ratio combining for transmission over correlated Rayleigh fading channels was quantified in [37], while a joint data and voice transmission scheme was

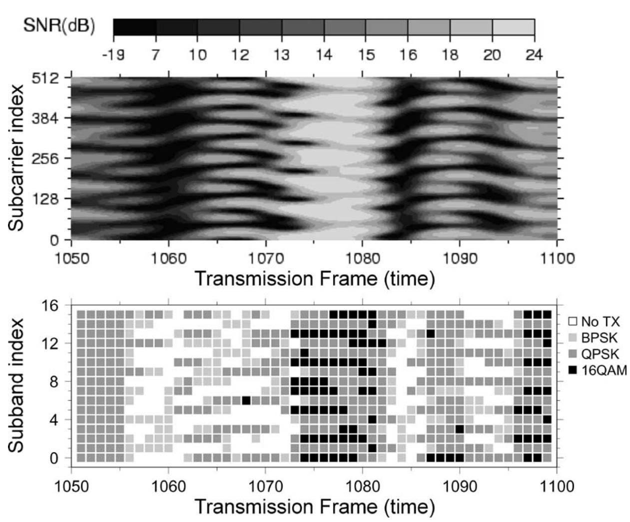


Fig. 6. ***Bit allocation of the TVTBR-AOFDM modem. The top graph is a contour plot of the channel SNR for all 512 subcarriers versus time. The bottom graph shows the modulation mode chosen for all 16 subbands for the same period of time. Each subband is composed of 32 subcarriers. The TVTBR AOFDM modem switches between target bit rates of 2, 3.4, 7, and 9.8 Mb/s, while attempting to maintain an estimated BER of 0.1% before channel coding. Average channel SNR is 16 dB over the channel of Fig. 3 at a normalized Doppler frequency of $F_D = 7.41 \times 10^{-2}$.***

proposed in [38]. Suffice to mention here a few examples of classic single-carrier contributions by Sampei, Komaki, Otsuki, and Morinaga [39]–[41] or the sophisticated joint adaptive joint coding and modulation schemes studied by Hole *et al.* [42] as well as by Duong, Oien, and Hole. These adaptive concepts were embedded into a rate-compatible punctured convolutionally coded CDMA system by Frenger *et al.* [43], while in [44] the expected channel quality estimation was improved by prediction. All these advances have led to the concept of HSDPA-style multimode transceivers in a variety of wireless systems [1], [4]–[7], [33], [45]–[48]. The range of various existing solutions that have found favor in already operational standard systems has been summarized in the excellent overview by Nanda *et al.* [47]. *The aim of these adaptive transceivers is to provide mobile users with the best possible compromise among a number of contradicting design factors, such as the power consumption of the MS, robustness against transmission errors, spectral efficiency, teletraffic capacity, audio/video quality, and so forth [46].*

B. AOFDM Versus STBC Basics

Another design alternative applicable in the context of OFDM systems is that the channel quality fluctuations observed, for example, in Fig. 2 are averaged out with the aid of time-, frequency-, or spatial-domain spreading codes,[1] *which potentially leads to the concept of three-dimensional spreading aided MC-CDMA [5]. This system may be expected to be capable of handling both TD and FD fading, where the duration of TD fades depends on the normalized Doppler frequency, while the width of the FD fades is determined by the height of the CIR. Suffice to say that in this scenario we may argue in a somewhat simplistic, but conceptually appealing manner that—depending on the normalized Doppler frequency and/or the length of the CIR—typically only a few chips of the spreading code may be obliterated by the time- or frequency-selective fading. Hence, there is a chance that the partially corrupted TD or FD spreading code itself may still be recoverable, which implies that the corresponding data symbol may also be recovered.*

The advantage of this approach is that in contrast to AOFDM-based communications, in MC-CDMA no channel quality estimation and signaling are required. Therefore, based on the more detailed exposures in [5], OFDM and MC-CDMA will be comparatively studied in Sections V-B–V-D of this contribution. We will also consider the employment of Walsh-Hadamard code-based spreading of each subcarrier's signal across the entire OFDM bandwidth, which was found to be an efficient frequency-domain fading countermeasure capable of operating without the employment of adaptive modulation [7].

A further techique, which is potentially capable of mitigating the channel quality fluctuations of wireless channels is constituted by space-time coding [8], [49]–[51], which will also be considered as an attractive antifading design option capable of attaining a high diversity gain. Space-time coding employs several transmit and receive antennas for the sake of achieving diversity gain and hence results in an improved performance, provided that the antenna elements experience independent fading.

It is a natural ambition to combine these adaptive transceivers with diversity aided MIMO space-time coded diversity systems in a further effort towards mitigating the effects of fading and hence rendering the channel more Gaussian-like. **A vital question in this context is, do adaptive transceivers retain their performance advantages in conjunction with MIMOs?**

Fig. 7 intuitively answers this question, where the instantaneous channel SNR experienced by an OFDM modem [7], [8], [52] versus both time as well frequency has been plotted. The OFDM modem has 512 subcarriers. More explicitly, Fig. 7 demonstrates for a conventional single-transmitter, single-receiver scheme as well as for Alamouti's STBC $\mathbf{G}_2$ [36] using one, two, and six receivers when communicating over a low-dispersion indoor channel. The average channel SNR is 10 dB. We can see in Fig. 7 that the variation of the instantaneous channel SNR for the single-transmitter and single-receiver scheme is severe. The instantaneous channel SNR may become as low as 4 dB due to deep fades of the channel. On the other hand, we can see that for the space-time block code $\mathbf{G}_2$ using one receiver the variation in the instantaneous channel SNR is slower and less severe.

As we increase the number of receivers, i.e., the diversity order, we observe that the variation of the channel becomes less dramatic, again, provided that the antenna elements experience **independent fading**. Effectively, by employing higher order diversity, the **independent fading channels** have been converted to an AWGN-like MIMO channel, as evidenced by the scenario employing the space-time block code $\mathbf{G}_2$ using six receivers. Hence, in a somewhat simplistic manner, we might be tempted to argue that the employment of adaptive modulation might become unnecessary when the combined transmit/receive diversity order is sufficiently high. More precisely, indeed, adaptive modulation may be viewed as a lower complexity fading countermeasure in comparison to space-time coding, since only a single transmitter and receiver is required.

These informal conclusions are confirmed by quantitative evidence in [7] and [8], suggesting that in the absence of shadow-fading no significant joint benefits accrue by combining HSDPA-style BbB-adaptive systems with high-order space-time coded transmit and receive diversity schemes, since both of these regimes aim for mitigating the effects of fading. More explicitly, *once the effects of fading have been sufficiently mitigated for the received signal's envelope to become near-constant, no further*

[1] Spatial-domain spreading [5] refers to spreading the signal to be transmitted to multiple transmit antennas for the sake of exploiting the independent fading of the antenna elements.

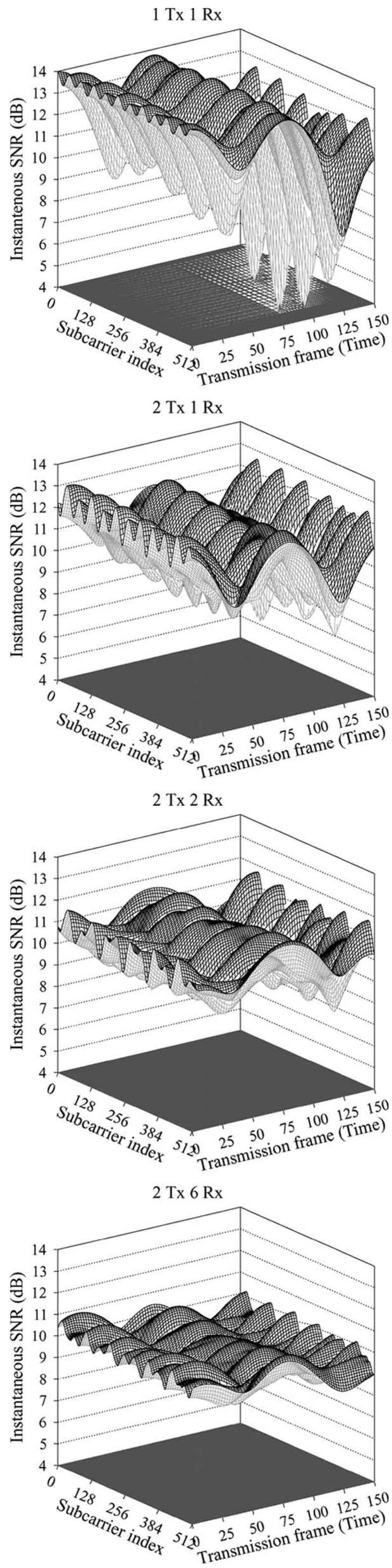


Fig. 7. ***Instantaneous channel SNR versus time and frequency for a 512-subcarrier OFDM modem in the context of a single-transmitter single-receiver as well as for the space-time block code*** G_2 ***[36] using one, two, and six receivers when communicating over an indoor wireless channel. The average channel SNR is 10 dB [8].***

fading countermeasures are necessary in the absence of shadow fading and MUI.

In conclusion of the above arguments, by employing multiple transmit antennas—as shown in Fig. 7—the effect of the channels' deep fades may be significantly reduced, provided that the individual antennas are spaced sufficiently far apart for the sake of experiencing independent fading. However, the antenna elements aften experience correlated fading, when their signals fade simultaneously owing to shadow fading. In this shadow-faded scenario the diversity gain of space-time coding erodes and only HSDPA-style TVTBR-AOFDM has the potential of mitigating its effects. Similarly, only TVTBR-AOFDM has the ability to combat the effects of time-variant MUI.

As a design alternative, in Part 2 of [5] MIMOs were employed at the base station (BS) for a different reason, namely for the sake of supporting multiple users, rather than for achieving transmit diversity gain. This is referred to as spatial division multiple access (SDMA) and it is feasible, since the users' CIR or FDCHTFs are accurately estimated and hence they may be viewed as unique user-specific signature sequences. These sufficiently accurately estimated FDCHTFs or user signatures allow us to both recognize and to demultiplex the simultaneous transmissions of the individual users, in a similar fashion to the unique user-specific spreading codes employed in CDMA systems.[2] We note, however, that this technique is only capable of reliably separating the users communicating within the same bandwidth, if their CIRs are sufficiently different. This assumption is typically valid for the uplink of spatially dispersed users, although it may have a limited validity, when the BS receives from mobile stations (MSs) having a similar CIR in its immediate vicinity. By contrast, different techniques have to be invoked for downlink multiuser transmissions. For reasons of space economy, here we restrict our discourse on SDMA-OFDM schemes to a rudimentary discussion of the related research aspects in Sections IV-F and IV-H. *A challenging SDMA-AOFDM problem is to design powerful nonlinear receivers, such as sphere decoders (SD) or genetic algorithm (GA) aided multiuser detectors (MUDs) [4]. These nonlinear detectors are particularly beneficial in so-called high-throughput rank-deficient scenarios, when the receiver is equipped with less antennas than the number of transmitters. This scenario was discussed in [53, Ch. 10, pp. 253–301], where the channel matrix becomes rank-deficient and hence cannot be inverted. In contrast to multiuser SDMA, spatial division multiplexing (SDM) [53] differentiates a number of antenna-specific*

[2]It is worth noting that although the CIRs are not orthogonal to each other, beneficially, there is an infinite number of them for differentiating the users. Naturally, their potentially poor correlations require a powerful SDMA MUD [5]. Another important point to note is that in a four-user up-link scenario, for example, where the BS typically has four receiver antennas, the estimation of $4 \times 4 = 16$ CIR has to be carried out, for example with the aid of the sophisticated decision-directed and interference-cancellation aided techniques of [5].

signals of a single user with the aid of his/her antenna-specific CIRs for the sake of increasing his/her throughput, which is typically more challenging than SDMA detection owing to the lower spatial separation and higher correlation of the antenna elements of SDM.

Having reviewed some of the basic OFDM-related transceiver functions, let us now provide a historical perspective of specific system design aspects in the next section.

IV. AOFDM RESEARCH HISTORY

A. Early Classic Contributions

The concept of parallel transmissions is likely to have emerged in 1957 in [227]. The first OFDM scheme was proposed by Chang in 1966 [16] for inter-symbol interference-free (ISI) transmissions over dispersive fading channels. The early OFDM research advances may be attributed to Weinstein, Peled, Ruiz, Hirosaki, Kolb, Cimini, Schüssler, Preuss, Rückriem, Kalet *et al.* [16], [17], [21]–[28], [31], [32], [54]. As a benefit of its attractive features, OFDM was standardized as the DAB as well as DVB scheme and it is the favorite 3GPP long-term evolution candidate system. It has also found its way into numerous WLAN standards.

These OFDM systems have been employed in military applications since the 1960s, for example by Bello [55], Zimmermann [17], Powers and Zimmerman [18], Chang and Gibby [20], and others. Saltzberg [19] studied a multicarrier system employing orthogonal time-staggered quadrature amplitude modulation (O-QAM) of the carriers.

The employment of the DFT for replacing the banks of sinusoidal modulators and demodulators was suggested by Weinstein and Ebert [21] in 1971, which significantly reduced the implementation complexity of OFDM modems. In 1980, Hirosaki [54] suggested an equalization algorithm in order to suppress both intersymbol and intersubcarrier interference caused by the CIR or by timing and frequency errors. Simplified OFDM modem implementations were studied by Peled [22] in 1980, while Hirosaki [23] introduced the DFT-based implementation of Saltzberg's O-QAM OFDM system. From Erlangen University, Kolb [24], Schüssler [25], Preuss [27], and Rückriem [28] conducted further research into the application of OFDM. Cimini [26] and Kalet [32] published analytical and early seminal experimental results on the performance of OFDM modems in mobile channels. Important further references are the books by van Nee and Prasad [56] as well as by Vandenameele *et al.* [57].

While OFDM transmission over mobile channels can alleviate the problem of multipath propagation, it is by no means flawless. Hence, recent research efforts have been focused on solving a set of inherent difficulties regarding OFDM, namely the peak-to-mean power ratio, time- and frequency-domain synchronization and on mitigating the effects of the frequency selective fading channel. These issues are addressed below with reference to the literature, while a more in-depth treatment is given throughout the book [5].

B. Peak-to-Mean Power Ratio

It is plausible that the AOFDM signal is constituted by the superposition of a high number of adaptively modulated subchannel signals. The resultant sum of numerous terms may exhibit a high instantaneous signal peak with respect to the average signal level. Particularly high out-of-band (OOB) harmonic distortion power emissions may be encountered, when the time domain signal traverses from a waveform having a low instantaneous power to a high-power AOFDM waveform. The resultant OOB emissions may only be mitigated, if the transmitter's power amplifier exhibits an extremely high linearity across the entire signal magnitude range. This OOB emission potentially contaminates the neighboring channels with adjacent channel interference (ACI). 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 to saturation. In order to avoid saturation and the resultant OOB emissions, the amplifiers are typically operated with a certain so-called back-off, creating a certain "head room" for the signal peaks. Two different families of solutions have been suggested in the literature, in order to mitigate these problems. Specifically, either the peak-to-mean power ratio has to be reduced or the amplifier linearity must be improved. In the context of AOFDM, all transceiver components, including the power amplifier has to satisfy the most stringent specifications of the highest-throughput QAM mode, such as 64QAM, for example.

More explicitly, Shepherd [58], Jones [59], and Wulich [60] have suggested the employment of techniques which aim for minimizing the peak power of the OFDM signal by employing different data encoding schemes before modulation. The philosophy of these data encoding schemes is that they employ specific block codes for encoding the modulating data symbols by concatenating appropriate redundant symbols, which mitigate the associated peak of the modulated signal. In other words, the legitimate crest-factor reduction code words exhibit low crest factors or peak-to-mean power envelope fluctuation. Müller [61], Pauli [62], May [63], and Wulich [64] suggested different algorithms for post-processing the time domain OFDM signal prior to amplification. By contrast, Schmidt and Kammeyer [65] employed adaptive subcarrier allocation in the spirit of Fig. 3, where the specific mapping of the modulating bits was simultaneously minimizing the resultant crest factor. Dinis and Gusmao [66]–[68] researched the employment of two-branch amplifiers, where each amplifier was fed with a near-constant-envelope signal. The clustered OFDM technique introduced by Daneshrad et al. [69] employs a set of parallel partial FFT processors in conjunction with separate radio-frequency transmission chains. OFDM systems exhibiting increased robustness to *nonlinear distortion have also been proposed by Okada et al. [70] as well as by Dinis and Gusmao [71]. These, as well as a range of other crest-factor related aspects of OFDM transmissions were treated in substantial depth both in Part 2 of [5] as well as by Han and Lee in [72]. Finally, crest-factor reduction may become even more critical in the context of MIMO-aided AOFDM transceivers, since the superposition of the multiple transmitted signals may potentially further aggravate the peak-to-mean ratio.*[3]

C. Synchronization

Time and frequency synchronization between the transmitter and receiver are of crucial importance as regards to the performance of an OFDM link [73], [74]. A wide variety of techniques have been proposed for estimating and correcting both timing and carrier frequency offsets at the OFDM receiver. These tend to rely on known pilot symbols or pilot tones embedded into the OFDM symbols, as suggested for example by Classen [75], Warner [76], Sari [77], Moose [78], as well as by Brüninghaus and Rohling [79]. The philosophy of these pilot-based methods is that they allow the receiver to estimate the associated change of phase imposed by the channel as well as by the timing- and frequency-differences of the transmitter's and receiver's oscillators. Once the composite misadjustment has been estimated, it may also be readily compensated. Frequency and timing fine-tracking algorithms exploiting the OFDM signal's cyclic extension were published by Moose [78], Daffara [80], and Sandell [81], where it was exploited that the phase-change across the OFDM symbol's duration imposed by the time-delay alone is known. If the actual phase difference deviates from the expected difference, this may be deemed to be the consequence of oscillator frequency- or phase errors and hence may be compensated. A high-performance timing synchronization technique was proposed by Williams *et al.* in [82], which employed a pre-FFT processing technique. A range of OFDM time- and frequency-domain synchronization techniques were proposed and investigated in [5, Ch. 5], but similarly to crest-factor reduction, further research advances are required in the field of synchronization schemes designed for MIMO-aided AOFDM.

D. Hybrid OFDM, CDMA, and MC-CDMA Systems

Combining multicarrier OFDM transmissions with code-division multiple access (CDMA) allows us to exploit the wideband channel's inherent frequency diversity by spreading each subcarrier's modulating symbol across multiple subcarriers. As mentioned before, the attraction of this operation is that even if a subcarrier and the corresponding chip of the spreading code mapped to it is obliterated, the spreading code and hence the corresponding original unspread symbol may still be recoverable. Furthermore, upon using unique user-specific spreading codes the system is also capable of supporting multiple users. This technique has been pioneered by Yee *et al.* [83], by Chouly *et al.* [84], as well as by Fettweis *et al.* [85]. Fazel and Papke [86] investigated convolutional coded OFDM/CDMA. Prasad and Hara [87] compared various methods of combining the two techniques, identifying three different structures, namely MC-CDMA [4], [5], multicarrier direct sequence CDMA (MC-DS-CDMA) [4], [5], and multitone CDMA (MT-CDMA). Like unspread AOFDM transmission, the various spreading assisted adaptive OFDM/CDMA methods also suffer from high peak-to-mean power ratios, which are dependent on the specific choice of the frequency-domain spreading scheme employed, as investigated by Choi *et al.* [88] and in Part 2 of [5]. The employment of MIMOs further increases the grade of design challenge.

E. Pilot-Aided Channel Estimation

The transmitted AOFDM signal is convolved with the CIR, or when interpreted in the frequency domain, it is multiplied by the FDCHTF. At the receiver these channel-induced linear distortions have to be removed, which is typically carried out by estimating the FDCHTF and then multiplying the received signal with the inverse of the FDCHTF.

In recent years, numerous research contributions have appeared on the topic of FDCHTF estimation techniques designed for employment in single-user, single transmit antenna-assisted AOFDM scenarios, since the availability of an accurate FDCHTF estimate is one of the prerequisites for coherent AOFDM detection. The techniques proposed in the literature can be classified as [5] *pilot-assisted*, *decision-directed* (DD), and *blind* channel estimation (CE) methods.

The simple philosophy of *pilot-assisted* frequency-domain channel estimation is that known pilot subcarrier symbols are allocated to the AOFDM subcarriers at a regular frequency spacing. The pilot spacing required for FDCHTF estimation is determined by the rate of FDCHTF fluctuation versus the frequency axis. More explicitly, these pilot subcarriers have to facilitate adequate sampling of the FDCHTF $H(n)$, requiring that the corresponding sampling frequency is set higher than the Nyquist frequency necessitated for the aliasing-free representation of the FDCHTF. This allows the FDCHTF's recovery from these pilot-aided "Nyquist-rate samples" [5]. The frequency separation between the FDCHTF fades observed in Fig. 2 depends on the maximum CIR duration observed for example in Fig. 3, since the CIR duration and the frequency-domain spacing of the deep fades constitute Fourier transform pairs. More

[3]Suffice to say that since both the in-phase and quadrature-phase components of the modulated signal are constituted by a multiplicity of superimposed modulated subcarrier signals, each of which may hence tend to obey a Gaussian distribution, the resultant complex-valued Gaussian, i.e., Rayleigh distributed signal exhibits a high peak-to-mean ratio, as illustrated for example in [5, Fig. 3.2]. Hence, this problem may also require further research.

explicitly, the longer the CIR, the more frequent the FDCHTF fades and vice versa.

The above-mentioned AOFDM symbol-by-symbol based FDCHTF estimation approach may be rendered more efficient, potentially requiring a lower pilot overhead, if the FDCHTF of consecutive OFDM symbols is predicted also as a function of time. This results in two-dimensional (2-D) pilot-aided FDCHTF estimation, as detailed in [5]. As in the context of the FDCHTF estimation versus frequency, the required time-direction pilot density is also determined by the Nyquist theorem, this time obeying twice the Doppler frequency encountered by the CIR taps, as it is illustrated both graphically as well as mathematically in [5]. The family of *pilot-assisted* channel estimation techniques was also investigated for example by Chang and Su [89], Höher [90]–[92], Itami *et al.* [93], Li [94], Tufvesson and Maseng [95], Wang and Liu [96], as well as Yang *et al.* [97]–[99]. Let us now briefly consider the family of decision-directed channel estimation (DDCE) schemes.

F. Decision-Directed Channel Estimation

In contrast to the above-mentioned pilot-aided schemes, in the context of DDCE all the sliced and remodulated subcarrier data symbols may be considered as pilots. In a slightly refined approach, an initial pilot-based FDCHTF estimate may be used for tentative subcarrier data detection, *as detailed in [5, Ch. 15 and 16] for both single- and SDMA-aided mulituser OFDM, invoking sophisticated iterative turbo-style joint parallel interference cancellation (PIC) and DDCE channel estimation*. To expound a little further, in a second stage all the detected subcarriers' data may be remodulated and used as a known pilot. The stored previous FDCHTFs may also be used for long-range FDCHTF prediction [34]. Provided that there are no subcarrier decision errors—which would lead to long-term error propagation—the performance of DDCE may approach that of the perfect channel estimation scenario. Phrased in more practical terms, in the absence of subcarrier symbol errors and also depending on the Doppler frequency, it was found that accurate FDCHTF estimates can be obtained, which may be of better quality in terms of the FDCHTF estimator's mean-square error (MSE), than the estimates offered by pure pilot-assisted schemes.

More explicitly, the family of *decision-directed* channel estimation techniques was investigated for example by van de Beek *et al.* [100], Edfors *et al.* [101], [102], Li *et al.* [103], Li [104], Mignone and Morello [105], Al-Susa and Ormondroyd [106], Frenger and Svensson [107], as well as Wilson *et al.* [108]. Furthermore, the family of *blind* channel estimation techniques using no channel sounding pilots was studied by Lu and Wang [109], Necker and Stüuber [110], as well as by Zhou and Giannakis [111].

In order to render the various DDCE techniques more amenable to employment in scenarios associated with a relatively high rate of FDCHTF variation expressed in terms of the OFDM symbol normalized Doppler frequency, linear prediction techniques well known from the speech coding literature [112], [113] can be invoked. To elaborate a little further, every FDCHTF sample may be predicted from a number of consecutive previous FDCHTF samples positioned at the specific subcarrier frequency considered. However, in case of for example 512 subcarriers, 512 predictors would be required, which results in a high complexity. A computationally more efficient solution is to transform the FDCHTF to the time domain, yielding the CIR. Since the CIR typically has only a low number of significant taps, their prediction is computationally less demanding, even when taking into account the complexity of the IFFT/FFT operation required for generating the CIR and transforming it back to the frequency domain [103]. The general concepts described by Duel-Hallen *et al.* [34] and the ideas presented by Frenger and Svensson [107], where a frequency domain prediction filter-assisted DDCE was proposed, also contributed substantially to the literature of DDCE. Furthermore, we should mention the contributions of Tufvesson *et al.* [114], [115], where a prediction filter-assisted frequency domain pre-equalization scheme was discussed in the context of OFDM. In a further contribution by Al-Susa and Ormondroyd [106], adaptive prediction filter-assisted DDCE designed for OFDM has been proposed upon invoking techniques known from speech coding, such as the Levinson–Durbin algorithm or the Burg algorithm [112], [116], [117] in order to determine the DDCE's predictor coefficients. Suffice to say in closing that in case of multiuser SDMA-aided OFDM modems the employment of adaptive modulation also becomes of high importance, since the number of interfering users is likely to fluctuate as a function of time, which results in time-variant MUI that cannot be readily combatted with the aid of fixed-mode OFDM. In the absence of interference-dependent adaptivity the decision errors may propagate catastrophically, since they would contaminate the channel estimates.

G. Multiuser MIMO OFDM Transmission

In recent years four basic MIMO designs have emerged, which have found application in diverse scenarios, as seen in Table 2. In this section, we briefly summarize the recent MIMO-OFDM research activities and suggest some open research problems.

A comprehensive review of MIMO OFDM air-interface proposals was provided by Yang in [119]. In contrast to STBC aided classic single-carrier modems, multicarrier modems offer the option of using the frequency domain subcarriers, rather than creating different time-slots for conveying the different antennas' signals, as it will be detailed in the context of Fig. 8. *In this OFDM context we may refer to these schemes as space-frequency coded arrangements*. As argued above, the full benefits of transmit

Table 2 Applications of MIMOs

Beamforming [1]	Typically $\lambda/2$-spaced antenna elements are used for the sake of creating a spatially selective transmitter/receiver beam. Smart antennas using beamforming have been employed for mitigating the effects of cochannel interfering signals and for providing beamforming gain.
Spatial Diversity [8] and Space-Time Spreading	In contrast to the $\lambda/2$-spaced phased array elements, in spatial diversity schemes, such as space-time block or trellis codes [8] the multiple antennas are positioned as far apart as possible, so that the transmitted signals of the different antennas experience independent fading, resulting in the maximum achievable diversity gain.
Space Division Multiple Access	SDMA exploits the unique, user-specific "spatial signature" of the individual users for differentiating amongst them. This allows the system to support multiple users within the same frequency band and/or time slot.
Spatial Division Multiplexing [119]	SDM systems also employ multiple antennas, but in contrast to SDMA arrangements, not for the sake of supporting multiple users. Instead, they aim for increasing the throughput of a wireless system in terms of the number of bits per symbol that can be transmitted by a given user in a given bandwidth at a given integrity.

diversity can only be exploited, if the individual space-time or space-frequency components experience independent fading. The performance versus complexity tradeoffs of MC-CDMA systems benefiting from both STBCs and cyclic delay diversity (CDD) was investigated by Lodhi *et al.* [120]. Different-throughput STBC and ST trellis coded AOFDM schemes combined with a variety of channel codes were studied by Liew and Hanzo [121] in terms of their performance versus complexity. For low-dispersion channels the STBC-aided AOFDM schemes performed better, while for high-dispersion channels the opposite was true.

During the same era Su, Safar, and Liu [122], [123] proposed an attractive space-frequency, rather than space-time block code, which is capable of supporting full-rate, full-diversity MIMO OFDM transmission. Zhang and Letaief [124] designed an adaptive power and bit allocation arrangement for multiuser MIMO OFDM systems, but further research is required for designing TVTBR-AOFDM schemes, which are capable of adapting the AOFDM QAM mode of the specific MIMO elements, each of which may be experiencing a potentially different channel quality. McLaughlin and Schafer [125] calculated a sphere packing lower bound and a pairwise error upper bound for space-time-frequency coded MIMO OFDM systems. A low-complexity space-time multiuser OFDM scheme was proposed by Wen *et al.* [126], while Su *et al.* [127] analyzed MIMO OFDM systems employing spatial, temporal, as well as frequency domain coding schemes. Although the design of MIMO-assisted TVTBR-AOFDM using high-throughput QAM modes was not considered by Shao and Roy [128], they suceeded in increasing the achievable diversity gain, when communicating over frequency-selective channels with the aid of a full-rate space-frequency block coding aided MIMO OFDM system.

Qiao et al. [129] proposed a novel iterative channel estimation algorithm for MIMO OFDM, while Kim et al. [133] combined the QR decomposition with the M-algorithm (QRD-M) for joint data detection and channel estimation in MIMO OFDM. Almost coincidentally, Ma [134] characterized a pilot-assisted modulation scheme designed for both carrier frequency offset (CFO) and channel estimation in wideband MIMO OFDM schemes. Sun et al. [137] contrived CFO estimation assisted expectation maximization (EM) based iterative receivers for MIMO OFDM systems, which may also be further developed for TVTBR-AOFDM systems. Wang et al. [138] also

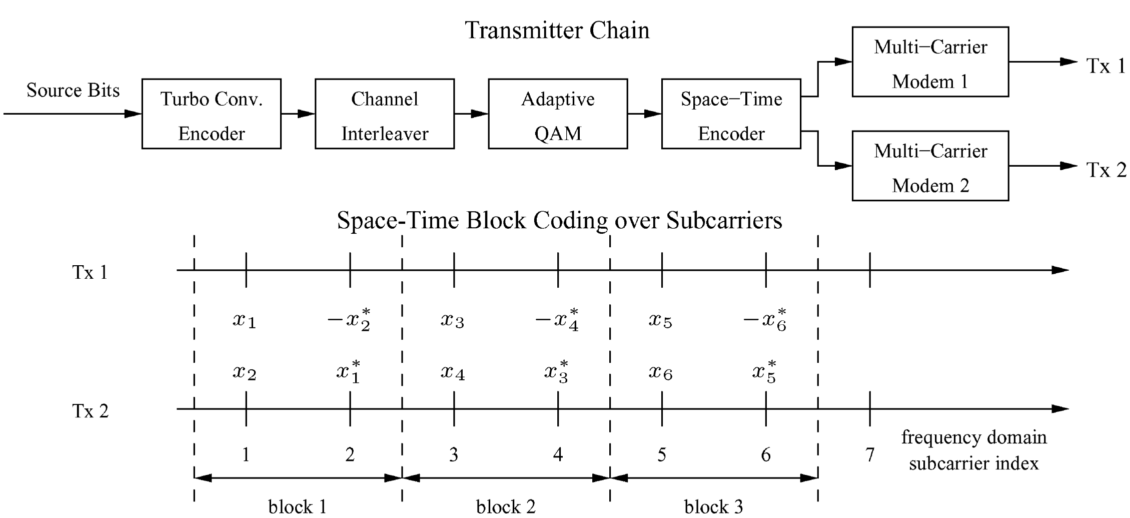


Fig. 8. ***Transmitter structure and space-frequency block encoding scheme.***

considered the MIMO OFDM channel estimation problem, which is quite challenging, since each MIMO link's FDCHTF has to be estimated. For example, for a 6×6-element adaptive MIMO OFDM system a total of 64 FDCHTFs has to be determined. Blind joint soft-detection aided slow frequency-hopped multicarrier DS-CDMA was the subject of [130], while space-time coded MIMO aided OFDM was the subject of [131] and [132].

Rey *et al.* [135] designed transmit prefiltering matrices for conveying as much of the desired signal power to the reference user as possible, and as little interference to the remaining users as possible. In case of the unrealistic simplifying assumption of having perfect channel information for all the users at the transmitter even before transmission it becomes plausible that the multiuser transmitter has the ability to separate the individual users' transmitted signals as well as to activate that particular TVTBR-AOFDM mode, which is capable of maintaining the target BER. However, in reality perfect noncausal multiuser channel information is unavailable and hence the authors aimed for minimizing the effects of channel estimation errors. A range of challenging open research problems can be found by considering the related design options. First, in time division duplex (TDD) systems the up-link (UL) and down-link (DL) signals are transmitted on the same frequency and hence they are likely to have a similar FDCHTF, which allows the transmitter to assume that the FDCHTF about to be experienced by the transmitter is similar to that estimated on the basis of the received signal. Another design option is to explicitly signal the FDCHTF from the remote receiver to the transmitter using for example high-compression vector-quantization. A third design option is to use long-range channel prediction for predicting the FDCHTF about to be experienced in the future based on previous FDCHTFs explicitly signaled by the receiver to the transmitter of the TVTBR-AOFDM system. The structure of the transmit correlation matrix was studied by Sampath *et al.* [136] with the aid of field trial results gleaned from a MIMO OFDM system and these results could be beneficially exploited for designing MIMO-aided TVTBR-AOFDM schemes.

Tan *et al.* [139] created a MIMO OFDM scheme by introducing a cross-antenna signal rotation arrangement for the sake of introducing additional degrees of freedom, which may be potentially exploited for example for generating independent fading and hence for increasing the achievable diversity order or for supporting an increased number of users. Park and Cho [140] aimed for reducing the inter-carrier interference (ICI) caused by high-Doppler time-varying channels, which may inflict a high residual BER in the context of MIMO-aided TVTBR-AOFDM using high-order QAM modes. Schenk *et al.* [141] characterized the transmitter/receiver phase noise effects on the achievable MIMO OFDM performance. These phase noise effects have to be carefully considered in the context of MIMO-assisted TVTBR-AOFDM using high-throughput QAM modes.

Various space-frequency code designs were also proposed by Borgmann and Bölcskei [142] for noncoherently detected frequency-selective MIMO OFDM systems. To elaborate a little further, noncoherently detected MIMO-OFDM is capable of dispensing with the above-mentioned high-complexity MIMO channel estimation problem. This becomes possible, since they consider the FDCHTF of the previously received and the current MIMO OFDM symbol to be identical, which allows the receiver to attribute the difference of the two consecutive subcarrier symbols entirely to the change imposed by the modulation process. Hence, noncoherently detected systems exhibit a low complexity, which is achieved at the cost of a typically 3 dB higher channel SNR requirement. When considering the high-throughput QAM modes of AOFDM, the channel estimation complexity may become extremely high and hence further research on noncoherently detected MIMO-aided TVTBR-AOFDM is required.

Tarighat and Sayed [143] characterized the effect of the in-phase/quadrature-phase (IQ) imbalances imposed by the imperfect orthogonality of the IQ components on MIMO OFDM systems and proposed various techniques of mitigating the associated BER degradations. These IQ imbalances may become particularly detrimental for the high-throughput QAM modes of AOFDM and hence have to be mitigated. In another practically motivated contribution Nanda *et al.* [144] reported their experience gleaned from a MIMO WLAN prototype that was capable of operating at rates in excess of 200 Mb/s.

A sophisticated transmit beamforming scheme was proposed by Choi and Heath [145], while Baek *et al.* [146] employed MIMOs in the context of a high-rate DAB system. Jiang *et al.* [147] contrived a joint transmitter and receiver design.

Zhang *et al.* [148] advocated the QR decomposition technique designed for a MIMO OFDM system, where successive interference cancellation can be used for mitigating the effects of inter-antenna interference or the MUI. The effects of carrier frequency offset (CFO) potentially result in a high residual BER, therefore Yao and Giannakis [149] proposed and characterized a low-complexity blind CFO estimator designed for OFDM.

H. Uplink Detection Techniques for Multiuser SDMA-OFDM

Combining adaptive antenna-aided techniques with OFDM transmissions was shown to be advantageous [5], for example in the context of suppressing the co-channel interference in cellular communications systems. Among others, Li, Cimini and Sollenberger [150]–[152], Kim *et al.* [153], Lin *et al.* [154], as well as Münster *et al.* [155] have investigated algorithms designed for multiuser channel estimation and interference suppression.

The related family of SDMA systems has recently drawn wide reseach interests [5]. Again, in these systems the L different users' transmitted signals are separated at the BS with the aid of their unique, user-specific spatial signature, which is constituted by their CIRs. A whole host of MUD techniques known from the literature of CDMA communications [4] lend themselves also to employment in the context of SDMA-OFDM on a per-subcarrier basis [5]. Some of these techniques are the least-squares (LS) [156]–[159], minimum mean-square error (MMSE) [150], [151], [156], [157], [160]–[166], successive interference cancellation (SIC) [118], [156]–[158], [162], [165], [167]–[170], parallel interference cancellation (PIC) [156], [171] and maximum likelihood (ML) MUD arrangements [156], [162], [165], [172]–[178]. A range of avantgarde sphere decoding aided SDM as well as GA-assisted SDMA near-ML designs and the radical so-called minimum bit error rate (MBER) MUD were detailed in [53, Ch. 10–12].

I. FDCHTF Estimation for Multiuser SDMA-OFDM

In contrast to the above-mentioned single-user OFDM scenarios, in a multiuser OFDM scenario the signal received by each antenna is constituted by the superposition of the different users' or antennas' signal contributions. Note that in terms of the MIMO structure of the channel the multiuser single-transmit antenna scenario is equivalent, for example, to a single-user space-time coded (STC) scenario using multiple transmit antennas. For the latter, a least-squares (LS) error channel estimator was proposed by Li *et al.* [179], which aims at recovering the different transmit antennas' FDCHTFs on the basis of the output signal of a specific reception antenna element and by also capitalizing on the remodulated received symbols associated with the different users. The performance of this estimator was found to be limited in terms of the mean-square estimation error in scenarios, where the product of the number of transmit antennas and the number of CIR taps to be estimated per transmit antenna approaches the total number of subcarriers hosted by an OFDM symbol. As a design alternative, in [180] a DDCE was proposed by Jeon *et al.* for a space-time coded OFDM scenario of two transmit antennas and two receive antennas.

Specifically, the FDCHTF associated with each transmit-receive antenna pair was estimated on the basis of the output signal of the specific receive antenna upon *subtracting* the interfering signal contributions of the remaining transmit antennas. These interference contributions were estimated by capitalizing on the knowledge of the FDCHTF of all interfering transmit antennas predicted during the $(n-1)$th OFDM symbol period for the nth OFDM symbol, also invoking the corresponding remodulated symbols associated with the nth OFDM symbol. To elaborate further, the difference between the subtraction-based FDCHTF estimator of [180] and the LS estimator proposed by Li *et al.* in [179] is that in the former the FDCHTFs predicted during the previous, i.e., the $(n-1)$th OFDM symbol period for the current, i.e., the nth OFDM symbol are employed for both symbol detection *as well as* for obtaining an updated channel estimate for employment during the $(n+1)$th OFDM symbol period. In the approach advocated in [180], the subtraction of the different transmit antennas' interfering signals is performed in the frequency domain.

By contrast, in [181] a similar technique was proposed by Li with the aim of simplifying the DDCE approach of [179], which operates in the time domain. A prerequisite for the operation of this parallel interference cancellation (PIC)-assisted DDCE is the availability of a reliable estimate of the various FDCHTFs for the current OFDM symbol, which are employed in the cancellation process in order to obtain updated FDCHTF estimates for the demodulation of the next OFDM symbol. In order to compensate for the channel's variation as a function of the OFDM symbol index, linear prediction techniques can be employed, as it was also proposed for example in [181]. However, due to the estimator's recursive structure, determining the optimum predictor coefficients is not as straightforward as for the transversal FIR filter-assisted predictor of single-user DDCE, both of which are detailed in [5].

J. Standardized OFDM Applications

Again, owing to their relatively high implementational complexity and crest-factor problems, OFDM applications have been scarce until the 1990s. In the ubiquitous subscriber telephone line environments OFDM is also referred to as a discrete multitone (DMT) scheme [182] and was ratified by the American National Standards Institute (ANSI) for asymmetric digital subscriber lines (ADSL) [183], for the high-bit-rate digital subscriber lines (HDSL) standard [184] as well as for very-high-speed digital subscriber lines (VDSL) [185] as a standard. OFDM has become the choice for numerous wireless standards, such as for digital audio broadcasting (DAB) [186], DVB terrestrial (DVB-T) [187] television as well as the standard of DVB for handheld (DVB-H) [188] terminals.

Wireless local area networks (WLANs) [189] and broadband radio access networks (BRANs) [190] also opted for employing OFDM. Furthermore, OFDM has been either approved or considered by the Institute of Electrical and Electronics Engineers (IEEE) for the following standards, which are amenable to AOFDM-style implementations:

- IEEE 802.11a [191]: An extension to IEEE 802.11 [192] that applies to WLANs and provides a bit rate of up to 54 Mb/s in the 5-GHz band. In comparison to IEEE 802.11, where frequency hopping spread spectrum (FHSS) or direct sequence spread spectrum (DSSS) are used, IEEE 802.11a employs an OFDM scheme which applies to wireless asynchronous transfer mode (WATM) networks and access hubs.

- IEEE 802.11g [193]: Offers wireless transmission over relatively short distances at 20–54 Mb/s in the 2.4-GHz band. It also uses an OFDM scheme.
- IEEE 802.11n [194]: Candidate standard for next generation WLANs, which was created from previous IEEE 802.11 standards by incorporating MIMO techniques. It offers high-throughput wireless transmission at 100–200 Mb/s.
- IEEE 802.15.3a [195]: Candidate standard using multiband OFDM (MB-OFDM), which grew out of the IEEE 802.15.3 [196] scheme proposed for wireless personal area networks (WPANs).
- IEEE 802.16 [197]: Defines wireless services operating in the 2–11 GHz band associated with wireless metropolitan area networks (WMANs), providing a communication link between a subscriber and a core network, e.g., the public telephone network and the Internet.

Following the above rudimentary overview of the past 40 years of OFDM research, further multicarrier research problems are addressed in [198]–[214]. Let us now focus our attention on a quantitative OFDM/MC-CDMA system design study, in an effort to contribute towards the next-generation adaptive multicarrier system research.

V. ADAPTIVE OFDM VERSUS MULTICARRIER CDMA

A. System Model and AOFDM Switching Levels

The family of OFDM-based multicarrier systems [5], [6] approach the theoretically *highest possible 2 Baud/Hz—i.e., 2 symbols/Hz—bandwidth efficiency quantified by Shannon, since in case of classic symbol-spaced Nyquist sampling at a rate of twice the bandwidth this is the highest achievable efficiency. This ultimate limit may be approached by OFDM modems, since they typically require only a small raised-cosine excess bandwidth for Nyquist-filtering, where typically unmodulated dummy subcarriers are used*. Hence, they are considered attractive for down-link wireless Internet services in future fourth generation (4G) systems as well as in high-speed WLANs. However, OFDM in its basic form cannot fully benefit from the multipath diversity potential of wideband channels.

It was reported in the literature that the synchronization requirements of MC-CDMA facilitate quasi-synchronous operation for low-speed vehicles and pedestrians [215]. Thus, an MC-CDMA system having the appropriate modem parameters constitutes an attractive down-link multiple access scheme for both fixed and slowly moving mobile terminals, where maintaining near-synchronous operation is feasible. Since MC-CDMA facilitates Rake reception, the performance of single-user MC-CDMA is characterised by that of an ideal Rake receiver. In a multiuser scenario joint-detection assisted MC-CDMA employing the minimum mean squared error block decision feedback equalizer (MMSE-BDFE) based MUD [216] is capable of approaching the single-user performance.

When channel coding is employed in conjunction with sufficiently long frequency domain interleaving, OFDM substantially benefits from the frequency domain diversity. However, OFDM may not be capable of exploiting the diversity potential of the channel to the same extent as MC-CDMA. This is because in OFDM the channel error statistics are more bursty than in MC-CDMA, since OFDM dispenses with spreading. By contrast, MC-CDMA may be able to still recover a spreading code and the corresponding bit, even if a few chips are partially corrupted. Hence, it is interesting to compare the coded BERs of OFDM and MMSE-BDFE aided MC-CDMA in conjunction with concatenated turbo codes and space-time block codes over wideband Rayleigh channels.

Various combinations of space-time (ST) codes and channel codes can be used for transmission over wideband fading channels [8], [217]. An attractive option is to use a half-rate turbo convolutional code concatenated to a space-time block code using two transmit antennas and to expand the signal constellation to a higher order modulation mode in order to match the throughput of the appropriate benchmark system having the same effective throughput, but using no channel coding. Another approach capable of maintaining a high effective throughput is to use a high-rate turbo BCH code in conjunction with a ST trellis code or ST block code and a lower order modulation mode, than in case of the half-rate FEC scheme. It was reported in [8], [217] that the former approach gives a lower BER, than the latter. Hence, we will employ a half-rate turbo convolutional code in our comparative study.

As argued in Fig. 2 of Section III, wideband fading channels exhibit two-dimensional channel quality variations, namely both time domain variation and frequency domain variation, and OFDM lends itself to exploiting these two-dimensional channel quality variations [5], [218], [219]. By contrast, although capable of providing frequency domain diversity with the aid of averaging the channel qualities of several subcarriers, MC-CDMA is less amenable to frequency domain adaptation, than to time domain adaptation. In order to contribute towards the next-generation system studies, the aim of the rest of this contribution is to compare the performances of space-time as well as turbo coded adaptive OFDM and MC-CDMA modems [218], [220], [221].

Fig. 8 portrays the stylized transmitter structure of the AOFDM system. The source bits are channel coded by a half-rate turbo convolutional encoder [8], [222] using a constraint length of $K = 3$ as well as an interleaver size of $L = 3072$ bits and interleaved by a random block interleaver. Then, the AOFDM block selects a modulation mode from the set of no transmission, BPSK, QPSK, 16-QAM, and 64-QAM modes, depending on the instantaneous channel quality perceived by the receiver, according to the predetermined SNR-dependent switching thresholds. It is assumed that the perfectly estimated channel quality experienced by receiver A is fed back to transmitter B superimposed on the next burst transmitted to receiver B. The modulation mode switching levels of the AOFDM scheme determine the average BER as well as the average throughput. A set of optimum switching thresholds was derived in [223] for transmission over flat Rayleigh fading channels. However, AOFDM modems employing these switching thresholds inevitably exhibit a variable average BER across the SNR range, despite aiming for a given fixed target BER, namely B_t. In order to achieve a constant target BER, while maintaining the maximum possible throughput, a new set of SNR-dependent switching thresholds was devised for transmission over wideband channels [224], [225].

Fig. 9 illustrates the switching levels optimized for both adaptive OFDM and adaptive MC-CDMA for maintaining the target BER of $B_t = 10^{-3}$.[4] Fig. 9 also shows the "avalanche" SNR, beyond which adaptive mode switching is abandoned in favor of the fixed highest order modulation mode, namely 64-QAM, since the BER of 64-QAM satisfies the target BER requirement.

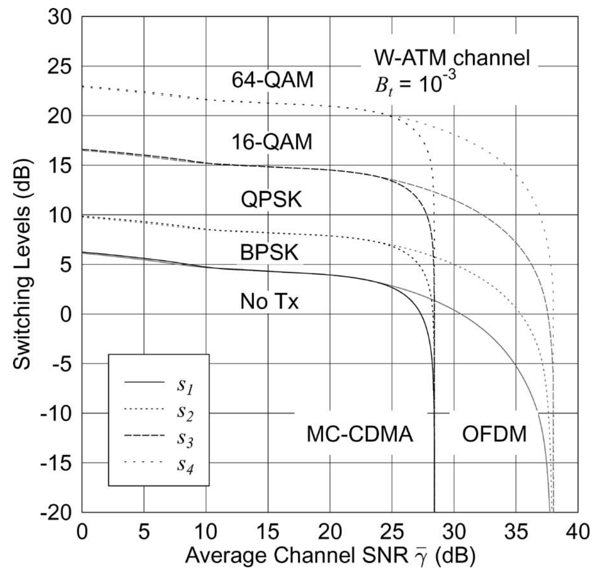


Fig. 9. ***Optimum switching levels devised for the target BER of $B_t = 10^{-3}$, when using 1 Tx antenna and 1 Rx antenna. The WATM channel model [6, p. 476] is assumed for MC-CDMA. The switching levels for OFDM were obtained for narrowband Rayleigh channels, which can be used for any multipath profile, since OFDM renders the dispersive channel nondispersive [5].***

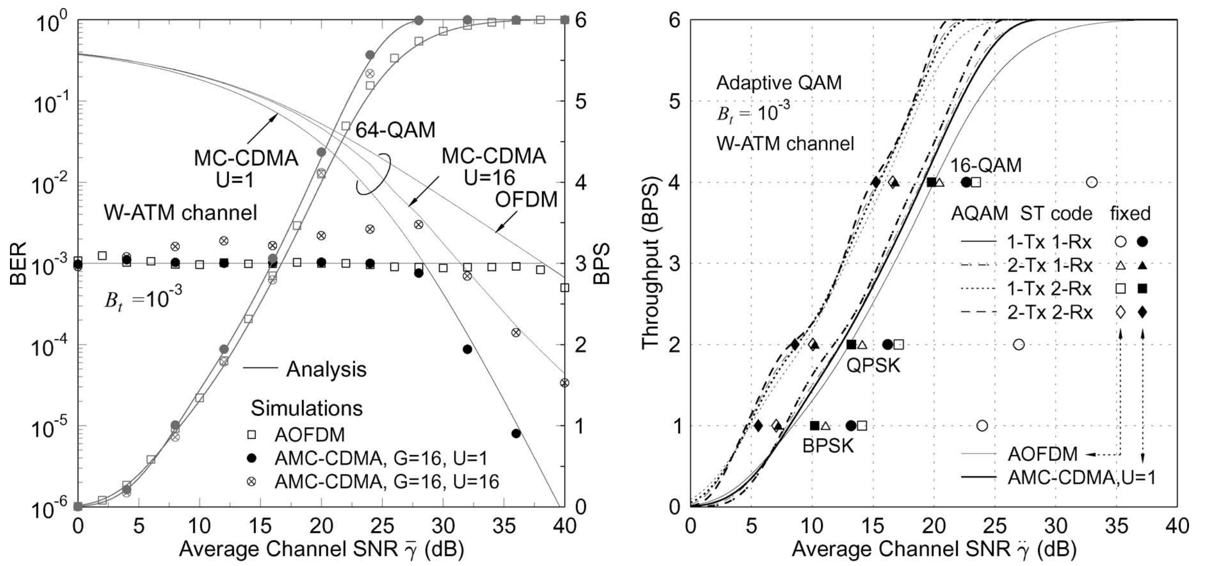
The modulated symbol is now space-time encoded. As seen at the bottom of Fig. 8, Alamouti's space-time block code [8], [51] is applied across the frequency domain. A pair of the adjacent subcarriers belonging to the same space-time encoding block is assumed to have the same channel quality. We employed the WATM channel model of Fig. 3 [6, p. 476] transmitting at a carrier frequency of 60 GHz, at a sampling rate of 225 MHz and employing 512 subcarriers. Specifically, we used a three-path fading channel model, where the average SNR of each path is given by $\bar{\gamma}_1 = 0.79192\bar{\gamma}$, $\bar{\gamma}_2 = 0.12424\bar{\gamma}$, and $\bar{\gamma}_3 = 0.08384\bar{\gamma}$. Each channel associated with a different pair of transmit and receive antennas was assumed to exhibit independent fading.

B. Uncoded Adaptive System

The simulation results characterizing the uncoded AOFDM modems are presented in Fig. 10. Since we employed the optimum switching levels of Fig. 9, both the AOFDM and the adaptive single-user MC-CDMA (AMC-CDMA) modems maintain the constant target BER of 10^{-3} up to the "avalanche" SNR value seen in Fig. 9 and then follow the BER curve of the 64-QAM mode. However, "full-user" AMC-CDMA defined as the system supporting $U = 16$ users with the aid of a spreading factor of $G = 16$ and employing the MMSE-BDFE joint detection (JD) receiver [216] exhibits a slightly higher average BER, than the target of $B_t = 10^{-3}$ due to the residual MUI of the imperfect joint detector. Since we derived the optimum switching levels based on a single-user system, the switching levels are no longer optimum, when residual MUI is present. The average throughputs expressed in terms of bits per symbol (BPS) steadily increase and reach the throughput of 64-QAM, namely 6 BPS. The throughput degradation of "full" user MC-CDMA in comparison to the single-user scenario was within a fraction of 1 dB. Observe in Fig. 10(a) that the analytical [5], [224] and simulation results are in good agreement, which we denoted by the lines and distinct symbols, respectively.

The effects of ST coding on the average BPS throughput are displayed in Fig. 10(b). Specifically, the thick lines represent the average BPS throughput of the AMC-CDMA scheme, while the thin lines represent those of the AOFDM modem. The four pairs of hollow and filled markers associated with four different ST-coded scenarios represent the BPS throughput versus SNR values associated with fixed-mode OFDM and fixed-mode MMSE-BDFE JD assisted MC-CDMA. Specifically, the right-most markers in Fig. 10(b) correspond to the 1-Tx/1-Rx, the second to the 2-Tx/1-Rx, the third to the 1-Tx/2-Rx, and the left-most to the 2-Tx/2-Rx scenarios. *First of all, we can observe that the BPS throughput curves of OFDM*

[4]These switching levels are defined as the average SNR levels, above which a specific modulation scheme may be safely activated, while maintaining the required target BER. Observe in Fig. 9 that the optimum switching levels decrease, as the average channel SNR increases, and hence higher throughput modulation modes can be invoked more frequently. This is plausible, since in case of a higher average SNR the variance of the SNR also tends to be lower and hence there is a reduced instantaneous SNR fluctuation around the average value, encouraging the employment of higher throughput modern modes.



(a) BER and Throughput, No ST coding (b) Throughput with ST coding

Fig. 10. ***Performance of uncoded five-mode AOFDM and AMC-CDMA. The target BER is $B_t = 10^{-3}$ transmitting over the WATM channel [6, p. 476]. (a) The constant average BER is maintained for AOFDM and single user AMC-CDMA, while "full-load" AMC-CDMA exhibits a slightly higher average BER due to the residual MUI. (b) The SNR gain of adaptive modems decreases as ST coding increases the diversity order. The BPS curves appear in pairs, corresponding to AOFDM and AMC-CDMA—indicated by the thin and thick lines, respectively—for each of the four different ST code configurations. The markers represent the SNRs required by the fixed-mode OFDM and MC-CDMA schemes for maintaining the target BER of 10^{-3} in conjunction with the four ST-coded schemes [5].***

and single-user MC-CDMA are close to each other, namely within 1 dB for most of the SNR range, regardless of the number of transmit and receive antennas. This is surprising, considering that the fixed-mode MMSE-BDFE JD assisted MC-CDMA scheme was reported to exhibit around 10-dB SNR gain at a BER of 10^{-3} and 30-dB gain at a BER of 10^{-6} over fixed-mode OFDM [5], [88]. This is confirmed in Fig. 10(b) by observing that the SNR difference between the ○ and ● markers is around 10 dB, regardless whether the 4, 2, or 1 BPS fixed-mode scenario is concerned, when using a single transmitter and receiver.

Still considering Fig. 10(b), let us now compare the SNR gains of the adaptive modems over the fixed modems. The SNR difference between the BPS curve of AOFDM and the fixed-mode OFDM represented by the symbol ○ at the same throughput is around 15 dB, when using a single transmitter and receiver. By contrast, the corresponding SNR difference between the adaptive and fixed-mode 4, 2, or 1 BPS MC-CDMA modem is around 5 dB in the single-transmitter, single-receiver scenario. More explicitly, since in the context of the WATM channel model [6, p. 476] fixed-mode MC-CDMA appears to exhibit a 10-dB SNR gain over fixed-mode OFDM, the additional 5-dB SNR gain of AMC-CDMA over its fixed-mode counterpart results in a total SNR gain of 15 dB over fixed-mode OFDM. Hence, ultimately the performance of AOFDM and AMC-CDMA becomes similar.

Let us now examine the effects of ST block coding more closely in Fig. 10(b). The SNR gain of the fixed-mode scheme due to the introduction of a 2-Tx/1-Rx ST block code is represented as the SNR difference between the two right-most markers. These gains are nearly 10 dB for fixed-mode OFDM, while they are only 3 dB for fixed-mode MC-CDMA modems. However, the corresponding gains are less than 1 dB for both adaptive modems. Since the transmitter power is halved due to using two Tx antennas in the ST codec, a 3-dB channel SNR penalty was already applied to the curves in Fig. 10(b). The introduction of the second receive antenna instead of the second transmit antenna eliminates this 3-dB penalty. Finally, the 2-Tx/2-Rx system gives around 3–4 dB SNR gain in the context of fixed-mode OFDM and a 2–3 dB SNR gain for MC-CDMA, in both cases over the 1-Tx/2-Rx system. By contrast, the gain of the 2-Tx/2-Rx scheme over the 1-Tx/2-Rx based adaptive modems was, again, less than 1 dB in Fig. 10(b). More importantly, for the 2-Tx/2-Rx scenario the advantage of employing adaptive modulation vanishes, since the fixed-mode MC-CDMA modem performs as well as the AMC-CDMA modem in this scenario. Moreover, the fixed-mode MC-CDMA modem still outperforms the fixed-mode OFDM modem by about 2 dB. We conclude that since the diversity order increases with the introduction of ST block codes, the channel quality variation becomes sufficiently low for the performance advantage of adaptive modems to vanish. This is achieved at the price of a higher complexity due to employing two transmitters and two receivers. These conclusions would be reversed, however, in the presence of the routinely encountered shadow fading or time-variant MUI fluctuations.

C. Turbo Coded Fixed Modem

When channel coding is employed in the fixed-mode multicarrier systems, it is expected that OFDM benefits more substantially than MC-CDMA, since the spreading operation of MC-CDMA may be viewed as a form of simple repetition coding, which was the main reason for the MC-CDMA system's better performance in the absence of FEC coding. The simulation results depicted in Fig. 11 show that the various half-rate turbo coded fixed-mode MC-CDMA systems consistently outperform OFDM. *However, the SNR differences between the turbo coded BER curves of multiuser detected OFDM and MC-CDMA are reduced considerably, namely to around 1–2 dB at a BER of* 10^{-4}.

D. Turbo Coded Adaptive System

The performance of the concatenated ST block coded and turbo convolutional coded adaptive modems is depicted in Fig. 12. We now applied a different set of switching levels, namely the optimum set of switching levels designed for the relatively high uncoded BER of 3×10^{-2}, rather than 10^{-3}, which was then further reduced by turbo coding. More explicitly, this uncoded target BER was obtained from the relationship of the uncoded and the turbo coded BPSK modems employing the same coding parameters over AWGN channels, with the ultimate objective of obtaining a coded BER of less than 10^{-7} for the adaptive modems. However, the simulations yielded zero bit errors when transmitting 10^9 bits, except for some SNRs, when employing only a single antenna.

Fig. 12(a) shows the BER of the turbo coded adaptive modems, when a single antenna is used. We observe in the figure that the BER reaches its highest value around the "avalanche" SNR point, where the adaptive modulation scheme consistently activates 64QAM. The system is most vulnerable around this point. In order to interpret this phenomenon, let us briefly consider the associated interleaving aspects. For practical reasons, we have used a fixed interleaver length of $L = 3072$ bits. When the instantaneous channel quality was high, the $L = 3072$ bits were spanning a shorter time-duration during their passage over the fading channel, because they were delivered by a six times lower number 64QAM symbols, than in case of BPSK symbols. Hence, the channel errors appeared more bursty, than in the lower throughput AOFDM modes, which transmitted the $L = 3072$ bits over a longer duration, hence dispersing the error bursts over a longer duration of time. The associated more random dispersion of erroneous bits indirectly enhances the error correction capability of the turbo code. On the other hand, in the SNR region beyond the "avalanche" SNR point the system exhibited a lower uncoded BER, reducing the coded BER even further. This observation suggests that *further research ought to determine the set of switching thresholds specifically and directly for a coded adaptive system. Alternatively, adaptive threshold-learning techniques, such as that proposed by Tang [226], have to be designed*.

We can also observe that the turbo coded BER of AOFDM is higher than that of AMC-CDMA in the SNR rage of 10 dB–20 dB, even though the uncoded BER is the same. This appears to be the effect of the limited exploitation of frequency domain diversity in coded OFDM, compared to MC-CDMA, which leads to a more bursty uncoded error distribution, hence degrading the turbo coded performance. The fact that ST block coding aided multiple antenna systems show virtually error free performance corroborates our argument.

Fig. 12(b) compares the throughputs of the turbo coded and uncoded adaptive modems exhibiting a comparable average BER. The SNR gains due to channel coding were in the range of 0 to 8 dB depending on the SNR region and the employed scenarios. Each of the two distinctive bundles of throughput curves corresponds to the scenarios of 1-Tx/1-Rx OFDM, 1-Tx/1-Rx MC-CDMA, 2-Tx/1-Rx OFDM, 2-Tx/1-Rx MC-CDMA, 1-Tx/2-Rx OFDM, 1-Tx/2-Rx MC-CDMA, 2-Tx/2-Rx OFDM, and 2-Tx/2-Rx MC-CDMA starting from the far right curve for the throughput values higher than 0.5 BPS. The SNR difference between the throughput curves of the ST- plus turbo-coded AOFDM and those of the corresponding AMC-CDMA schemes was reduced compared to the uncoded performance curves of Fig. 10(b). The SNR gain owing to ST block coding in the context of AOFDM and AMC-CDMA was limited to about 1 dB due to the halved transmitter power, when using two transmit antennas. Therefore, again, ST block coding provides limited additional gains in the context of adaptive modems, since either of these fading countermeasures is capable of mitigating the effects of fading in the absence of

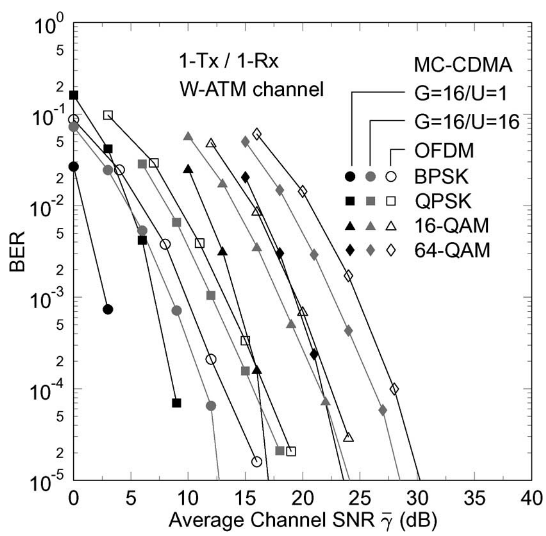
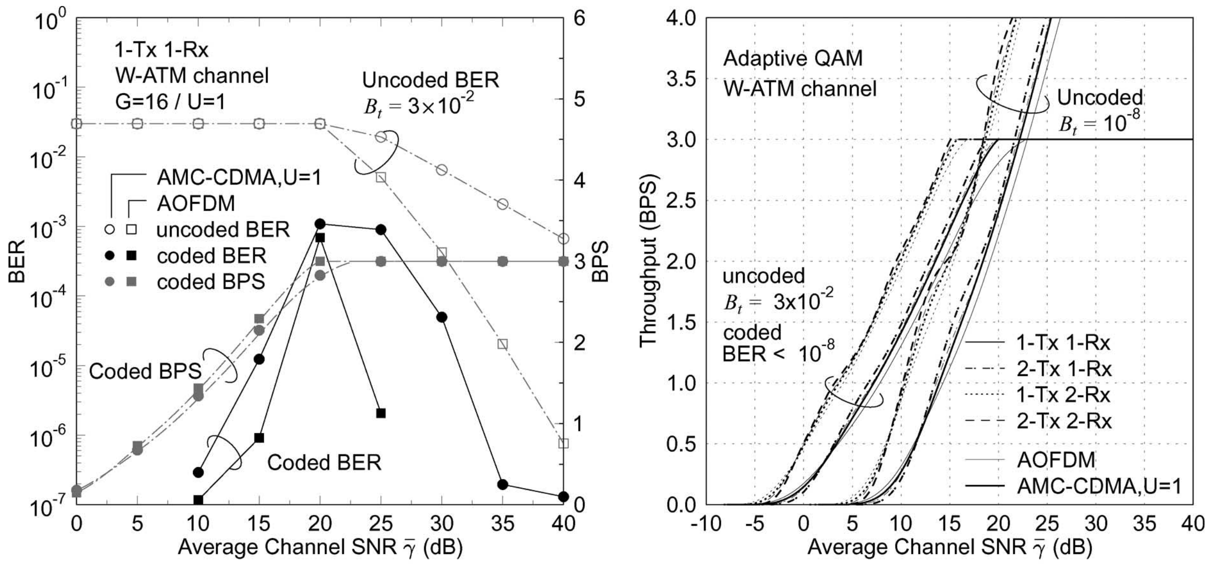


Fig. 11. ***Performance of turbo convolutional coded fixed-mode OFDM and MC-CDMA over the WATM channel of [6, p. 476]. JD MC-CDMA still outperforms OFDM. However, the SNR gain of JD MC-CDMA over OFDM is reduced to 1–2 dB at a BER of 10^{-4} [6].***



(a) BER (b) Throughput

Fig. 12. ***Performance of the concatenated ST block coded and turbo convolutional coded adaptive OFDM and MC-CDMA systems over WATM channel of [6, p. 476]. The uncoded target BER is 3 $\times$ 10^{-2}. (a) The coded BER becomes about 10^{-3} near the "avalanche" SNR point around 20 dB, when a single transmit and receive antenna was used. (b) By contrast, the coded BER was less than 10^{-8} in subfigure (b) for most of the SNR range, although this is not explicitly shown in the figure [6].***

shadow fading. By contrast, in the presence of shadow fading only a combination of ST and adaptive modulation are capable of combatting the hostile channel effects. Please note that the bunch of uncoded BPS curves seen in Fig. 10(b) was recorded for the scenario, when we have $B_t = 10^{-8}$ rather than for $B_t = 3 \cdot 10^{-2}$, as indicated by the text at the upper right corner of Fig. 10(b). These curves are there in order to explicitly contrast the coding gain of the curves in the left bunch corresponding to the half-rate turbo coded scenario.

VI. CONCLUSION

An overview of HSDPA-style BbB-adaptive OFDM/MC-CDMA multicarrier systems was provided, identifying their most critical design aspects. These systems exhibit numerous attractive features, rendering them eminently eligible for employment in next-generation wireless systems. The second half of the paper provided a system design example and characterized the achievable performance.

Explicitly, the performance of ST block coded constant-power adaptive multicarrier modems employing optimum SNR-dependent modem mode switching levels was investigated. The adaptive modems maintained the constant target BER specified, while maximizing the average throughput. As expected, it was found that ST block coding reduces the relative performance advantage of adaptive modulation, since it increases the diversity order and eventually reduces the channel quality variations. When turbo convolutional coding was concatenated to the ST block codes, near-error-free transmission was achieved at the expense of the halving the average throughput. Compared to the uncoded system, the turbo coded system was capable of achieving an increased throughput in the low SNR region at the cost of an increased complexity. The study of the relationship between the uncoded BER and the corresponding coded BER showed that adaptive modems obtain higher coding gains, than that of fixed modems. This was due to the fact that the adaptive modem avoids burst errors even in deep channel fades by reducing the number of bits per modulated symbol eventually to zero. However, these conclusions were based on assuming independent fading for all the individual ST antennas and if this assumption is no longer valid due to shadow fading or MUI fluctuations, only adaptive modulation techniques have the potential of maintaining a certain target error rate. ■

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ABOUT THE AUTHORS

Lajos Hanzo (Fellow, IEEE) received the first-class degree in electronics in 1976 and the doctorate degree in 1983, both from the Technical University of Budapest, Hungary. In 2004, he was awarded the D.Sc. degree from the University of Southampton, Southampton, U.K.

During his career in telecommunications he has held various research and academic posts in Hungary, Germany, and the U.K. Since 1986 he has been with the Department of Electronics and Computer Science, University of Southampton, where he holds the chair in telecommunications. He co-authored 15 books in excess of 10 000 pages on mobile radio communications, published about 700 research papers, organized and chaired IEEE conferences, presented overview lectures, and has been awarded a number of distinctions. Currently he heads an academic research team, working on a range of research projects in the field of wireless multimedia communications sponsored by industry, the Engineering and Physical Sciences Research Council (EPSRC) U.K., the European IST Programme, and the Mobile Virtual Centre of Excellence (VCE), U.K. He is an enthusiastic supporter of industrial and academic liaison and he offers a range of industrial courses.

Dr. Hanzo is a Fellow of the Royal Academy of Engineering. He is also an IEEE Distinguished Lecturer of both the Communications as well as the Vehicular Technology (VT) Society and a Fellow of the IEE/IET. He is a Governor of the IEEE VT Society. For further information on research in progress and associated publications, please refer to http://www-mobile.ecs.soton.ac.uk.

Byoung-Jo Choi received the B.Sc. and M.Sc. degrees in electrical engineering from KAIST, Korea, in 1990 and 1992, respectively, and the Ph.D. degree in mobile communications from the University of Southampton, Southampton, U.K.

He was a Postdoctoral Research Assistant at the University of Southampton from 2001 to 2002. He has been working for LG Electronics, Korea, since January 1992, where he was involved in developing the KoreaSat monitoring system, digital DBS transmission system, and W-CDMA based wireless local loop (WLL) system. He is a recipient of the British Chevening Scholarship awarded by the British Council, U.K. His current research interests are related to mobile communication systems design with emphasis on adaptive modulation aided OFDM, MC-CDMA, and W-CDMA.