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Boosting the Performance of Wireless Communication Systems: Theory and Practice of Multiple-Antenna Techniques

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ABSTRACT

Multiple-antenna systems, denoted multiple-input multiple-output systems, promise huge performance gains over conventional single-antenna systems. Many MIMO transmission schemes proposed in the literature are, however, based on idealized assumptions. Several effects that may occur in a practical system, such as intersymbol interference, nonperfect channel knowledge, fast fading, and correlation between the individual transmission paths, significantly influence system performance. On the basis of two particularly simple but nonetheless efficient MIMO transmission schemes, the above effects are illustrated, and possible countermeasures are discussed.

INTRODUCTION

What can be done in order to enhance the data rate of a wireless communication system? One can choose a shorter symbol duration T. This, however, implies that a larger fraction of the frequency spectrum will be occupied, since the bandwidth required by a system is determined by the baud rate $1/T$. Moreover, wireless channels are normally characterized by multipath signal propagation caused by reflections, scattering, and diffraction. A shorter symbol duration might therefore cause an increased degree of intersymbol interference (ISI) and thus performance loss. As an alternative to a shorter symbol duration, one can go for a multicarrier approach and multiplex data onto multiple narrow subbands. Thus, the problem of ISI may be circumvented. But still, the requirement for increased bandwidth remains, which is crucial in regard to the fact that frequency spectrum has become a valuable resource.

The seminal work by Foschini and Gans [1] and, independently, by Telatar [2] at the end of the 1990s suggested that there is another alternative to accomplish higher data rates over wireless links: the use of multiple transmit (Tx) and receive (Rx) antennas, an alternative that does not require any extra bandwidth at all (and no extra transmission power). Therefore, multiple antennas provide a very promising means to increase the spectral efficiency of a system. In his paper on the capacity of multi-antenna Gaussian channels [2], Telatar showed that given a wireless system employing N_T Tx antennas and N_R Rx antennas, the maximum data rate at which error-free transmission over a fading channel is theoretically possible is proportional to the minimum of N_T and N_R (provided that the N_TN_R transmission paths between the Tx and Rx antennas are statistically independent). Thus, huge throughput gains may be achieved by $N_T \times N_R$ systems compared to conventional 1×1 systems that use only a single antenna at either end of the wireless link. With multiple antennas, the *spatial* domain is exploited, as opposed to the first two approaches mentioned above, in which the time and frequency domain are utilized, respectively.

To date, many transmission schemes for multiple-antenna systems — often referred to as multiple-input multiple-output (MIMO) systems — have been proposed in the literature. These schemes may coarsely be classified as *spatial multiplexing* schemes and *space-time coding* techniques. Spatial multiplexing (SM) schemes employ multiple antennas at both the transmitter and receiver, and send independent data streams over the individual Tx antennas. At the receiver, the data streams are separated by employing an interference cancellation type of algorithm. In this context, the number of Rx antennas should not be less than the number of Tx antennas. With SM schemes the overall data throughput is enhanced and significant fractions of the data rates promised in theory are accomplished. A well-known example of an SM scheme is the Bell Labs Layered Space-Time Architecture (BLAST) [3].

As opposed to SM schemes, space-time coding (STC) techniques exploit spatial diversity, which yields an additional diversity and/or coding gain compared to a 1×1 system. Spatial diversity is available if the individual transmission paths from the Tx antennas to the Rx antennas fade more or less independently. In this case, the probability that all paths are degraded at the same time is significantly smaller than the probability that a single transmission path is in a deep fade. By means of STC techniques, improved bit error rate performance is accomplished. This may in turn be translated to higher throughput if an adaptive modulation/channel coding scheme is applied (which is done in most up-to-date wireless communication systems). Examples of STC techniques are space-time trellis codes (STTCs, e.g., [4]) and space-time block codes (STBCs, e.g., [5, 6]). STTCs are a generalization of trellis-coded modulation (TCM) to multiple Tx antennas. They may be represented by means of a trellis structure. STBCs, on the other hand, essentially perform a mapping of data symbols onto multiple Tx antennas, which may be described by means of a space-time (ST) mapping matrix.

With STC techniques, multiple antennas at the receiver are optional, which is a major advantage, particularly in cellular systems. Multimedia services are usually characterized by asymmetric data traffic, where the predominant part of data transfer occurs in the downlink (DL), in the direction from the base station (BS) to the mobile station (MS). This is because a single mobile user may download comparably large amounts of data from the BS, whereas rather little data traffic is required in the uplink (UL) in order to initiate the download. Therefore, STC techniques are very attractive in order to enhance the crucial DL, because only the BS needs to be equipped with additional antennas. However, in order to provide a capacity gain, multiple antennas are also required at the MS [2].

Often, MIMO transmission schemes proposed in the literature are based on somewhat idealized assumptions. For example, many schemes are designed for frequency flat fading channels (i.e., channels without ISI). However, depending on the delay spread of the physical channel (due to multipath signal propagation), the employed transmitter and receiver filter, and the symbol duration T, this assumption might not be valid in a practical system. Another common premise is the assumption of block fading, where the channel is presumed to be invariant over the duration of a complete data block. This assumption is questionable, for instance, if the MS moves at high speed. The same is true if there is a significant frequency offset between the local oscillators at the transmitter and receiver. Concerning recovery of the transmitted data symbols, most MIMO transmission schemes rely on accurate knowledge of the channel at the receiver in order to enable either good separation of the individual transmission signals in an SM scheme or appropriate combining/decoding in an STC technique. However, this assumption is questionable if a low signal-to-noise ratio (SNR) at the receiver is given or the channel is fast fading. Another important issue is that the performance gain over a 1×1 system actually achievable in a practical MIMO system might be smaller than promised in theory, because implementing an optimal transmitter/receiver strategy might be too complex, so one has to resort to suboptimal solutions. Finally, in some cases the individual transmission paths between the Tx and Rx antennas may not be statistically independent. The assumption of independent paths often made in theory is only valid if the antenna spacing at the transmitter and receiver is sufficiently large and there is a rich scattering environment. Insufficient antenna spacing, lack of scattering, or a significant line-of-sight (LOS) signal component between transmitter and receiver introduces *correlation* between the individual transmission paths.

In the following, the individual effects that may occur in a practical MIMO system are investigated in more detail. Their influence on system performance is illustrated and possible countermeasures are discussed. This is done on the basis of *delay diversity* and the *Alamouti scheme*. These two schemes are particularly simple but nonetheless efficient STC techniques, and are therefore of great interest for practical implementations. Delay diversity [7] may be interpreted as the simplest special case of an STTC. The Alamouti scheme [5] is the simplest example of an STBC.

For mathematical details concerning some of the topics addressed in the following, please refer to [8]. Related work may, for example, be found in [9, ref. 5 therein]. We start with an introduction to delay diversity and the Alamouti scheme.

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Delay Diversity and the Alamouti Scheme

Throughout this article the equivalent complex baseband notation is used. It is assumed that an M-ary modulation scheme is used, such as M-ary amplitude shift keying (M-ASK) or M-ary phase shift keying (M-PSK). The coefficients of the equivalent discrete time channel model shall be denoted channel coefficients in the following. They comprise the physical channel, pulse shaping filter, and receiver filter, as well as the sampling rate and phase. In this article baud-rate sampling is assumed.

Transmitter Structures

The transmitter structures of delay diversity and the Alamouti scheme are depicted in Fig. 1a and b. In the delay diversity scheme, the number of Tx antennas, N_T, is arbitrary. The same M-ary data symbol $x[k]$ is transmitted over all N_T antennas. At each Tx antenna Tx_i, a delay of $\delta_{i-1} \cdot T$ is applied, where δ_{i-1} is a non-negative integer for all $i = 1, \ldots, N_T$ and $\delta_0 = 0$. This means that at a given time index k the symbol $x[k - \delta_{i-1}]$ is transmitted over antenna Tx_i. Note that any 1×1 system may be enhanced by delay diversity.

The Alamouti scheme employs two Tx antennas. The M-ary data symbols are processed as pairs where two subsequent time indices are used in order to transmit a pair of data symbols over the two Tx antennas. The symbol rate of the corresponding 1×1 system is therefore retained. The mapping of a pair $[x[k], x[k + 1]]$ of data symbols onto the two Tx antennas and the two subsequent time indices k and $k + 1$ is done according to the Alamouti matrix $\mathbf{A}$, which is depicted in Fig. 1b. At time index k, the first antenna transmits $x[k]$ and the second antenna transmits $-x^*[k + 1]$, where $(.)^*$ denotes complex

In order to accomplish a larger diversity gain, the Alamouti scheme may be extended to multiple Rx antennas by stacking the received vectors of the individual Rx antennas in a joint vector and using the corresponding joint equivalent channel matrix for detection.

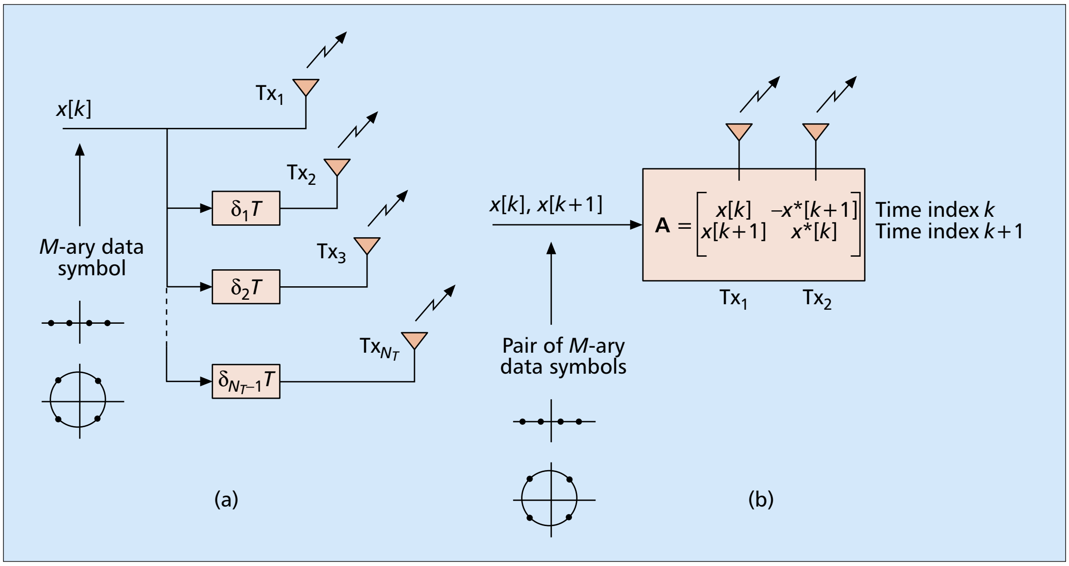

Figure 1. *Transmitter structure of: a) delay diversity; b) the Alamouti scheme.*

conjugation. At time index $k + 1$, the first antenna transmits $x[k + 1]$ and the second $x^*[k]$. The Alamouti scheme is backward compatible with the corresponding 1×1 system.

Receiver Structures

As seen in the next section, a frequency-flat channel model is transformed into a frequency-selective channel model if delay diversity is applied at the transmitter. Therefore, for delay diversity one will always require an equalizer at the receiver to mitigate ISI. For example, a trellis-based equalizer/detector may be used based on the well-known Viterbi algorithm performing maximum likelihood sequence estimation (MLSE) of the transmitted data sequence. With delay diversity, spatial diversity is transformed to diversity in the frequency domain. One or more antennas may be employed at the receiver; multiple Rx antennas will lead to an additional diversity gain.

In the Alamouti scheme, data detection may be performed by means of a simple matrix vector multiplication, provided that the channel model is frequency-flat. For the time being, let us assume a single Rx antenna and frequency-flat fading. The transmission paths from the two Tx antennas to the Rx antenna may then be described by complex-valued channel coefficients h_1 and h_2, which are (for the purpose of simple receiver design) assumed to be constant over two consecutive symbol durations. When grouping the received samples $y[k]$ and $y[k + 1]$ in a vector $\mathbf{y}[k]$, the overall transmission model including the Alamouti mapping may be written

$$\mathbf{y}[k] = \mathbf{H}_{eq} \cdot \mathbf{x}[k] + \mathbf{n}[k], \qquad (1)$$

where $\mathbf{H}_{eq}$ denotes an equivalent 2×2 channel matrix. The vector $\mathbf{x}[k]$ comprises the transmitted data symbols $x[k]$ and $x[k + 1]$ and the vector $\mathbf{n}[k]$ the noise samples $n[k]$ and $n[k + 1]$ at the receiver. Due to the special structure of the Alamouti matrix $\mathbf{A}$, the product matrix $\mathbf{H}_{eq}^H \mathbf{H}_{eq}$ is diagonal. Therefore, the transmitted data symbols may be estimated by the matrix vector multiplication $\mathbf{H}_{eq}^H \mathbf{y}[k]$, provided that the channel coefficients are perfectly known at the receiver:

$$\mathbf{H}_{eq}^H \mathbf{y}[k] = (|h_1|^2 + |h_2|^2) \cdot \mathbf{x}[k] + \mathbf{H}_{eq}^H \mathbf{n}[k], \qquad (2)$$

where $(.)^H$ denotes the hermitian conjugate of a matrix. In block fading, the above estimation of the data symbols $x[k]$ and $x[k + 1]$ is optimal in the sense of maximum likelihood (ML) estimation. Note that the desired symbols are combined in a constructive way because they are multiplied by a sum of absolute terms. The noise, however, is combined incoherently. Due to spatial diversity, the probability that the factor $(|h_1|^2 + |h_2|^2)$ is close to zero is comparably small. The detection complexity for the conventional Alamouti receiver according to Eq. 2 is virtually the same as for the corresponding 1×1 system. In order to accomplish a larger diversity gain, the Alamouti scheme may be extended to multiple Rx antennas by stacking the received vectors of the individual Rx antennas in a joint vector and using the corresponding joint equivalent channel matrix for detection. If the channel model is frequency-selective an appropriate equalizer is required, which will be discussed below.

Performance on a Frequency-Flat Fading Channel

The bit error rate (BER) performance of the Alamouti scheme as a function of the SNR in dB is depicted in Fig. 2 for the case of statistically independent transmission paths characterized by frequency-flat fading. Rayleigh fading has been assumed, which implies that there is a rich scattering environment but no LOS signal component between transmitter and receiver. The overall transmission power is normalized by the number of Tx and Rx antennas in order to provide a fair comparison with the corresponding 1×1 system. As can be seen from Fig. 2, the Alamouti scheme with one Rx antenna yields a gain of about 8.5 dB over the corresponding 1×1 system at a BER of $2 \cdot 10^{-3}$. By employing a second Rx antenna, an additional 3.5 dB gain is accomplished at the same BER.

The BER performance of delay diversity

given a frequency-flat fading channel is virtually the same as that of the Alamouti scheme, apart from a small loss due to the ISI introduced by delay diversity.

Effects in a Practical MIMO Transmission Scheme

In this section the individual effects that may occur in a practical MIMO system — ISI, nonperfect channel knowledge, fast fading, frequency offsets between transmitter and receiver as well as spatial correlation — shall be investigated in more detail. Their influence on system performance as well as possible countermeasures shall be discussed. We restrict ourselves to the single-user case; interference between multiple users shall not be considered here.

Frequency-Selective Fading

Most MIMO transmission schemes proposed in the literature (e.g., STTCs or orthogonal STBCs such as the Alamouti scheme) are designed for frequency-flat fading channels. However, if there are multipath signals with rather large propagation delays, the assumption of a frequency-flat fading channel might not be valid, depending on the symbol duration used. Then, an ST equalizer is required at the receiver in order to counteract ISI. Alternatively, a multicarrier approach may be pursued using multiple frequency-flat subbands to circumvent the problem of ISI.

In general, the ST equalizer has to account for the specific structure of the employed STC scheme. This means that standard equalizer algorithms already available for a 1×1 system are not suitable, and generalized algorithms are required. For example, a trellis-based equalizer/detector may be used that is based on the Viterbi algorithm or the Bahl-Cocke-Jelinek-Raviv (BCJR) algorithm. As opposed to the Viterbi algorithm, which provides "hard" outputs (e.g., ± 1 in the binary case), the BCJR algorithm calculates "soft" outputs that are optimal in the sense of the maximum a posteriori (MAP) criterion. Since the invention of turbo codes in 1993, the BCJR algorithm — already introduced back in 1972–1974 — has experienced a remarkable renaissance. This is because soft outputs are of particular interest in systems employing concatenated codes in conjunction with an iterative ("turbo") detection scheme at the receiver. Such turbo schemes are known to yield excellent system performances. The turbo principle may, for example, be applied in a MIMO transmission scheme consisting of an inner STC scheme and an outer channel code, used in order to further improve performance.

In the following we assume that the individual transmission paths from Tx antenna Tx_i to Rx antenna Rx_j are described by $L + 1$ complex-valued channel coefficients $h_{ij}^{(0)}, \ldots, h_{ij}^{(L)}$, where L denotes the effective channel memory length. The ISI is represented by the channel coefficients $h_{ij}^{(1)}, \ldots, h_{ij}^{(L)}$. Now, what would a generalization of a standard trellis-based equalizer algorithm look like in the case of the Alamouti scheme? It turns out that the received samples should be processed as pairs, corresponding to the transmitter structure, because then the equalizer complexity in terms of number of trellis states is minimized [8]. To be specific, if the effective channel memory length L is an even number, the number of states for MLSE resulting for the Alamouti scheme is the *same* as in the corresponding 1×1 system, namely M^L. If L is an odd number, M^{L+1} trellis states are required. Due to the pairwise processing of the received samples, each trellis segment in the Alamouti trellis spans two consecutive time indices, k and $k + 1$, as opposed to a single time index k in the conventional algorithm. As in the frequency-flat case, one or more Rx antennas may be employed. Additional Rx antennas, however, increase the computational complexity for the individual branch metrics within the trellis. The above concept may be extended to more general STBCs, in which a single ST matrix spans more than two consecutive symbol durations.

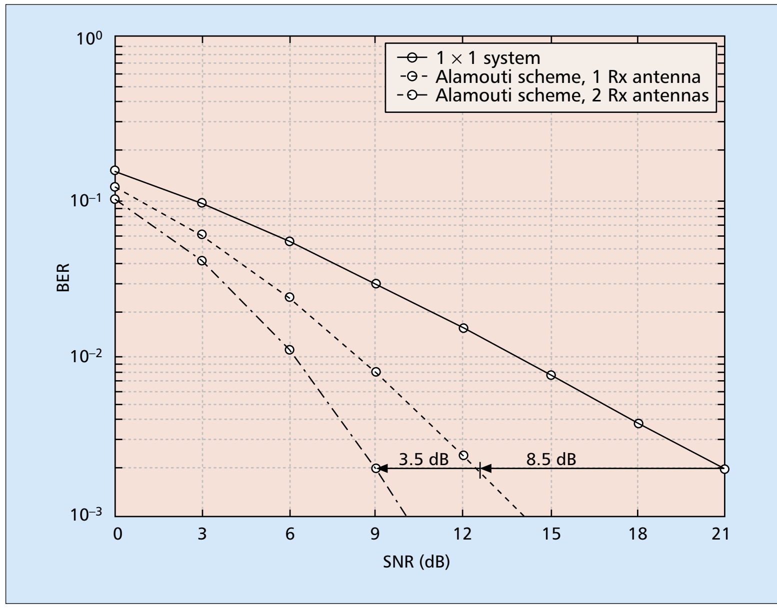

Figure 2. *BER performance of the Alamouti scheme for the case of frequency-flat fading.*

In contrast to the above considerations, a significant advantage of delay diversity is that virtually the same equalizer algorithm may be used as in the corresponding 1×1 system. However, as seen below, the equalizer has to handle an increased channel memory length and is therefore more complex. Due to the fact that the same data symbols are transmitted over each Tx antenna, the delay elements applied at the individual Tx antennas may be interpreted as part of the discrete time channel model. By this means, one obtains an equivalent single-Tx-antenna channel model. This principle is illustrated in Fig. 3 for the case of two Tx antennas and a delay of $\delta = 1$ at the second antenna. Based on the channel coefficients $h_{1j}^{(l)}$ (transmission path $Tx_1 \rightarrow Rx_j$) and $h_{2j}^{(l)}$ (transmission path $Tx_2 \rightarrow Rx_j$), where $l = 0, \ldots, L$, the equivalent single-Tx-antenna channel model for Rx antenna Rx_j may be described by the channel coefficients $[h_{1j}^{(0)}, h_{1j}^{(1)} + h_{2j}^{(0)}, \ldots, h_{1j}^{(L)} + h_{2j}^{(L-1)}, h_{2j}^{(L)}]$. Note that the resulting channel memory length is actually a function of the delay δ. It is increased from L to

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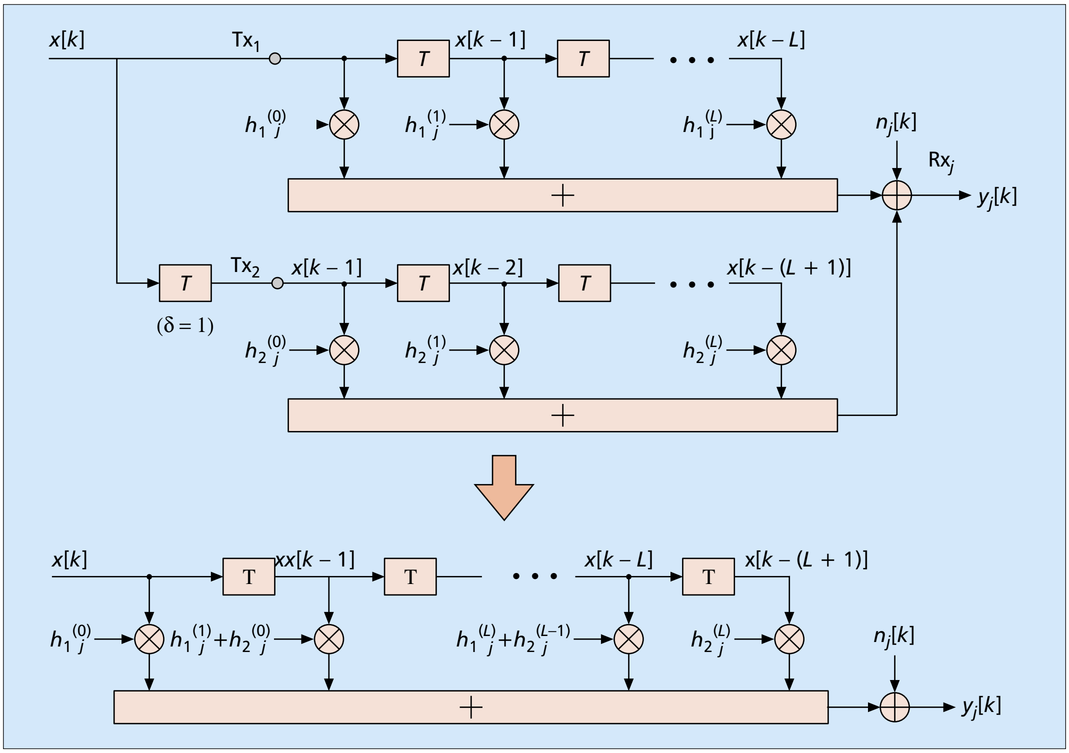

Figure 3. *Equivalent single-Tx-antenna channel model for delay diversity (two Tx antennas, δ = 1).*

$L + 1$, since δ equals one. In general, if δ_{max} denotes the maximum delay applied at the transmitter, the memory length resulting for the equivalent single-Tx-antenna channel model is given by $L + \delta_{max}$. The required number of equalizer states is $M^{L+\delta_{max}}$ for MLSE here, instead of M^L in the corresponding 1×1 system.

Let us restrict ourselves to the case of two Tx antennas. An interesting question is, what is the optimal delay δ at the second antenna in the sense of a maximum diversity gain? In the original delay diversity scheme [7], a delay of $\delta = 1$ is applied at the second Tx antenna. This choice is in fact optimal in the case of frequency-flat fading. In general, if the channel model is frequency-selective, δ should be chosen such that the channel coefficients $h_{1j}^{(l)}$ and $h_{2j}^{(l)}$ do not overlap in the time domain but are completely separated [8]. Therefore, a delay of $\delta > L$ yields the optimal diversity gain, provided that $M^{L+\delta}$ equalizer states are indeed employed. However, this equalizer complexity might be too large for a practical implementation. Suppose that a maximum number of M^K equalizer states shall not be exceeded, where $K < L + \delta$. Then only the first $K + 1$ channel coefficients of the equivalent single-Tx-antenna channel model are considered in the equalizer branch metrics, whereas the remaining $L + \delta - K$ channel coefficients are discarded. This leads to residual ISI and thus to a performance loss. Therefore, given a fixed equalizer complexity, there is a trade-off concerning the optimal choice of δ. If δ is chosen too small, only a small diversity gain is accomplished. However, if δ is chosen too large, the performance loss due to residual ISI dominates. Figure 4 illustrates this trade-off for binary antipodal signaling ($M = 2$) and the example of the typical urban (TU) channel model defined by the European Telecommunications Standards Institute (ETSI) as part of the specifications for the Global System for Mobile Communications (GSM). The RAKE receiver bound (RRB), which is a lower bound on the bit error probability (BEP), is plotted as a function of the delay δ for an SNR of 10 dB. The average powers of the channel coefficients resulting for the equivalent single-Tx-antenna channel model are qualitatively depicted as well for delays $\delta = 0$ and $\delta = 4$. The TU channel model is characterized by an effective channel memory length of about $L = 3$. If a delay of $\delta > 3$ is chosen, the channel coefficients $h_{1j}^{(l)}$ and $h_{2j}^{(l)}$ are completely separated in the time domain. In this case, provided an equalizer with $2^{L+\delta}$ states is used (blue curve), the RRB is minimized (i.e., the diversity gain is maximized). However, if an equalizer with a fixed number of 2^4 equalizer states is employed (red curve), the RRB is minimized for a delay of $\delta = 2$.

Channel Estimation, Nonperfect Channel Knowledge, and Fast Fading

If *coherent* reception is applied, for each of the $N_T N_R$ transmission paths the coefficients of the corresponding equivalent discrete time channel model have to be estimated. The quality of the resulting channel estimates is crucial for overall system performance. In a *training-based* system, channel estimation (CE) is performed on the basis of a known pilot sequence inserted at the transmitter. Alternatively, there are *blind* CE schemes that require no training symbols at all and thus save valuable bandwidth.

Let us focus on training-based CE in the following. A popular training-based CE method is to correlate the known pilot sequence with the corresponding received samples. In a 1×1 system, the quality of the resulting channel estimates is determined by the *auto-correlation* properties of the employed pilot sequence, the number of pilot symbols used, and the noise variance (the same is true for delay diversity, since an equivalent single-Tx-antenna channel model can be found, as seen in the previous subsection). If the channel model is frequency-selective with an effective channel memory length L, good auto-correlation properties are required for *any* cyclic shift of the pilot sequence by up to L symbols. In a general MIMO system, a pilot sequence is required for each individual Tx antenna. The quality of the channel estimates thus is not only determined by the (cyclic) auto-correlation properties of each pilot sequence, but is significantly influenced as well by the mutual (cyclic) *cross-correlation* properties of the employed pilot sequences. In fact, to define good pilot sequences for multiple Tx antennas can be a demanding task. An alternative to the above correlation method is the least squares (LS) CE method, which tends to be less sensitive to imperfect cross-correlation properties of the individual pilot sequences. Recently, *differential* STC techniques have gained considerable interest, in which the receiver does not need any channel knowledge at all. However, differential STC techniques were designed for frequency-flat fading channels.

Let us consider an example concerning orthogonal STBCs such as the Alamouti scheme. Due to the inherent orthogonality, the pilot sequences for the individual Tx antennas do not have to be designed explicitly, provided that the channel model is frequency-flat. One may just take *any* pilot sequence for the first Tx antenna and then use its space-time encoded versions (obtained on the basis of the STBC matrix) for the remaining Tx antennas. Performing CE given known pilot symbols is dual to estimating the transmitted data symbols given known channel coefficients (illustrated earlier for the Alamouti scheme, cf. Eqs. 1 and 2). In the Alamouti scheme, for example, matrix **A** depicted in Fig. 1b may be used in order to obtain ML estimates for channel coefficients h_1 and h_2. Assuming block fading, the channel estimates may in turn be used in order to estimate the unknown data symbols according to Eq. 2. Due to noise the channel estimates will not be accurate, however. Therefore, the receiver will use a noisy channel matrix $\hat{\mathbf{H}}_{\mathrm{eq}}$ instead of the correct one in order to perform the matrix vector multiplication in Eq. 2. Unfortunately, the product matrix $\hat{\mathbf{H}}_{\mathrm{eq}}^{\mathrm{H}}\mathbf{H}_{\mathrm{eq}}$ is in general not a diagonal matrix (i.e., the orthogonality of the scheme is lost) [8]. The secondary diagonal elements of $\hat{\mathbf{H}}_{\mathrm{eq}}^{\mathrm{H}}\mathbf{H}_{\mathrm{eq}}$ cause interference between the data symbols $x[k]$ and $x[k + 1]$. Therefore, independent estimates for the two symbols are no longer feasible. In a practical system, a performance loss of about 1–2 dB can be expected, concerning the BER as a function of the SNR.

Similarly, an orthogonality loss occurs if the block fading assumption does not hold, even if the channel coefficients are perfectly tracked. A time-varying channel model may be due to motion of the MS, causing a Doppler spread that is a function of the speed of the MS and the carrier frequency used. However, the influence

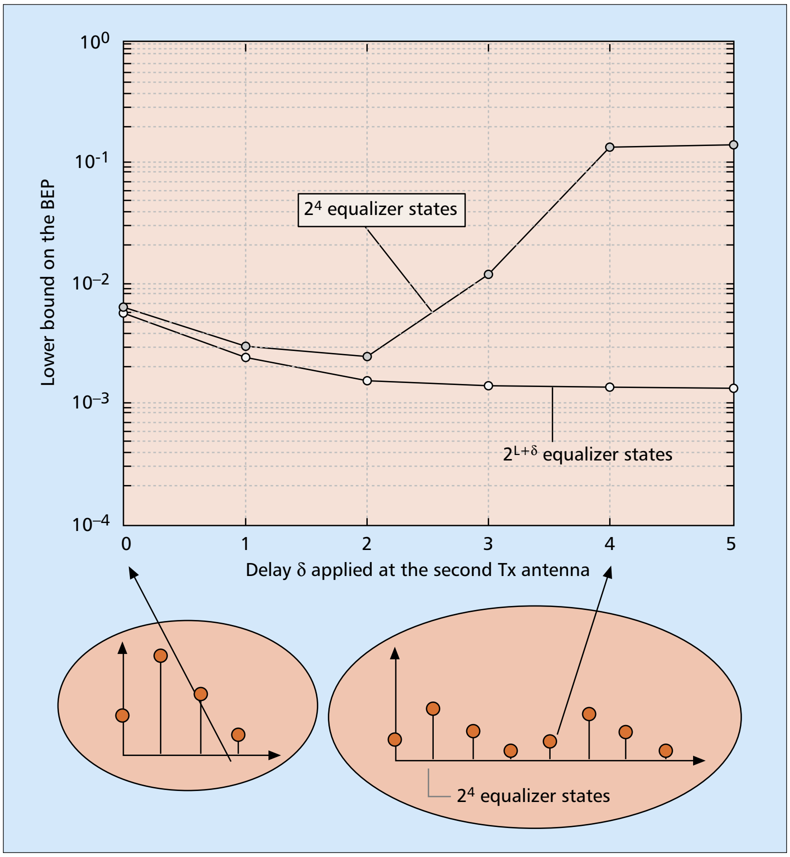

Figure 4. *Trade-off concerning the optimal choice of the delay δ applied at the second Tx antenna.*

on system performance is normally quite small in a practical system, even if tracking is not applied [8]. Another reason for a time-varying channel model may be a significant frequency offset Δf between the local oscillators at the transmitter and receiver. In this case, the channel coefficients will rotate within the complex plane with an angular speed determined by Δf. However, if the channel is perfectly tracked, the influence on system performance is normally negligible in a practical system.

SPATIAL CORRELATION

If the antenna spacing at the transmitter or receiver is small, or there is a lack of scattering from the physical environment, the $N_T N_R$ transmission paths might be correlated. The MS is often assumed to be surrounded by a ring of local scatterers, which leads to comparably small correlations between the individual antennas. For example, in the case of a mobile phone operating at a carrier frequency of 900 MHz, the maximum possible antenna spacing is about one quarter of the wavelength, which results in a correlation of about ρ = 0.4 to ρ = 0.7. At the BS, however, one usually observes significantly larger correlation values given the same antenna spacing. This is due to the fact that in a typical urban scenario, the BS is normally located at rooftop level and is therefore not surrounded by local scatterers. This means that the signals transmitted/received by the BS are usually characterized by a comparably small angular spread and thus strong correlation. On the other hand, it is rather

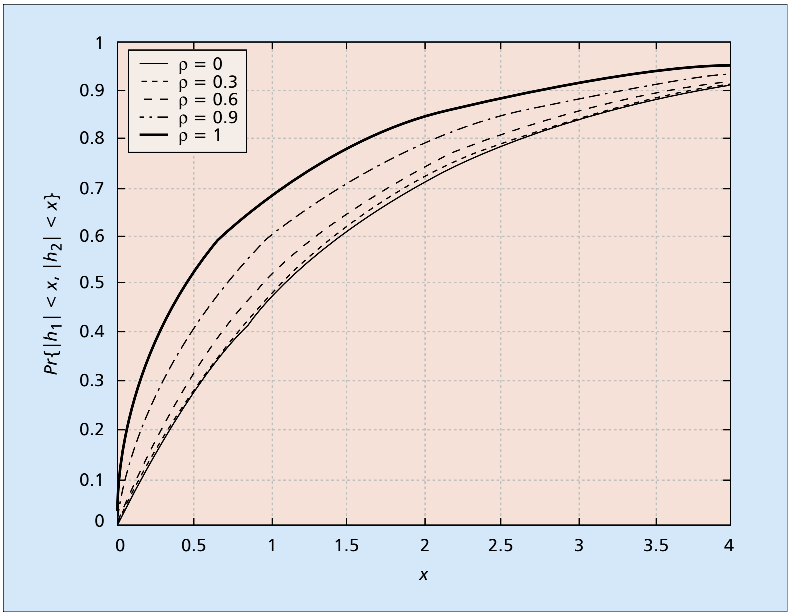

Figure 5. *Probability that both* $|h_1|$ *and* $|h_2|$ *are smaller than a certain value* x, *given different correlation values* ρ.

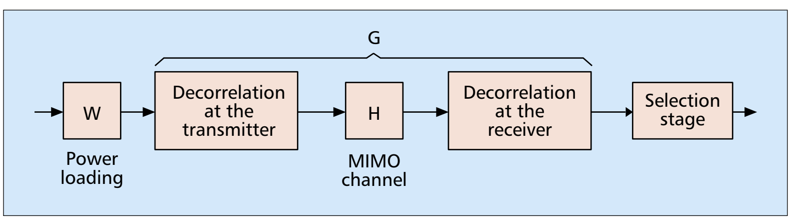

Figure 6. *The overall transmission model in the case of a spatially correlated MIMO channel.*

uncritical to provide generous antenna spacings at the BS; therefore, correlation due to the BS antenna array may often be neglected.

Spatial correlation reduces the degree of spatial diversity offered by a system. Therefore, the probability that multiple correlated transmission paths are simultaneously in a deep fade increases with growing correlation. This is illustrated in Fig. 5 for the case of two transmission paths described by channel coefficients h_1 and h_2, which have been modeled as real-valued Gaussian random variables with zero mean and unit variance. The probability that both $|h_1|$ and $|h_2|$ are smaller than a certain value x is plotted as a function of x, for different correlations ρ between zero and one. For example, if h_1 and h_2 are uncorrelated (ρ = 0), the probability that both $|h_1|$ and $|h_2|$ are smaller than $x = 0.5$ is 0.27. This value increases only slightly as long as the correlation does not exceed 0.6. Only for strong correlations is the above probability significantly larger (e.g., 0.53 for ρ = 1).

A spatially correlated MIMO channel may be described by means of a channel matrix

$$\mathbf{H} = \mathbf{R}^{1/2}\, \mathbf{G}\, \mathbf{T}^{1/2}, \tag{3}$$

where **R** denotes the Rx correlation matrix, **G** a spatially uncorrelated channel matrix, and **T** the Tx correlation matrix. Although diversity loss due to correlation cannot be undone, the transmitter and receiver may still take measures in order to improve the effective SNR, provided that the correlation matrices **T** and **R** are known. For example, the correlation matrices **T** and **R** may be estimated on the basis of long-term observation of the channel matrix **H**. Based on a singular-value decomposition (SVD) of **T**, the channel may be decorrelated at the transmitter [10]. The transmitter may then apply a power loading strategy, say, according to the "water-filling" rule ("statistical water-filling"), by using a diagonal weight matrix **W**. Based on an SVD of **R**, the channel may be decorrelated at the receiver. The receiver may then employ a selection stage that selects the dominating subchannels represented by the strongest eigenvalues of **R** [10]. By this means, the receiver complexity may be reduced considerably without causing a significant performance loss. The overall transmission model is depicted in Fig. 6.

Conclusions

As we have seen, the theoretical assumptions underlying many MIMO transmission schemes may sometimes be questionable in practice. Several effects which may occur in a practical system have been addressed in this article and possible counter measures have been discussed. Some of these effects may, in fact, cause significant performance degradations and need to be tackled in order to exploit the huge gains promised by theory. This offers many opportunities for exciting future research.

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