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Compatible Improvement of the GSM/GPRS System by Means of Space-Time Block Codes

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***Abstract*—We investigate the application of the Alamouti Space-Time Block Code (STBC) on basis of the GSM/GPRS system, especially with regard to aspects of compatibility with current specifications. The performance improvements obtainable by means of this technique are demonstrated on basis of simulation results. A novel trellis-based soft-output equalization and detection algorithm for the Alamouti scheme is presented, which is of the same complexity as in the single transmit antenna case. Moreover, appropriate training sequence pairs are derived. In this context, optimized partner sequences with respect to the eight GSM sequences are introduced that significantly outperform any pair of GSM sequences. Analytical results are presented demonstrating the sensitivity of the Alamouti scheme to fast and/or frequency-selective fading as well as to non-perfect knowledge of the channel coefficients at the receiver.**

I. Introduction

RECENTLY, the application of multiple transmit antenna techniques in wireless communications systems has gained much interest. By introducing an additional *spatial* component to the signal processing carried out in the transmitter, significant performance improvements upon single transmit antenna systems can be obtained. In this context *Space-Time Trellis Codes (STTC)* [1] and *Space-Time Block Codes (STBC)* [2] are subject to current research activities. They exploit spatial diversity yielding an additional diversity and/or coding gain and thus an improved bit error performance compared to a system utilizing only a single transmit antenna ((1x1)-system). Spatial diversity results from the fact that the individual transmission paths from the transmit antennas to the receive antenna(s) are likely to fade *independently*. With STTC and STBC multiple antennas at the receiver are optional.

Deploying STTC and STBC in future wireless communications systems promises reliable transmission at high data rates, e.g., required for *3rd generation (3G)* bearer services. Such services are usually characterized by asymmetric data traffic, where the predominant part of data transfer occurs in the downlink (DL). Therefore, in order to enhance the crucial DL, the application of STTC and STBC is very attractive because only the base transceiver station (BTS) needs to be equipped with additional antennas.

This paper considers the application of the *Alamouti scheme* [3] in a GSM/GPRS[1] system [4], often referred to as a 2.5G system. The Alamouti scheme is the simplest special case of a STBC and employs two transmit antennas.

Original aspects of this paper include a transmitter structure presented in Section II, which is designed with special regard to compatibility aspects. In this context, appropriate pairs of training sequences are introduced in Section III. The proposed pairs consist of an original GSM sequence and an optimized partner sequence. They perform significantly better than any pair of GSM sequences. Moreover, a novel equalization and detection algorithm is presented in Section III, which is derived from the conventional Max-Log-MAP algorithm [5] for a single transmit antenna system and which is characterized by the same complexity. For related work see [6]-[10]. New analytical results in Section IV demonstrate the sensitivity of the Alamouti scheme to fast and to frequency-selective fading as well as to non-perfect knowledge of the channel coefficients at the receiver. Corresponding simulation results are given in Section V, illustrating the bit error performance improvements obtainable on a typical urban wireless channel by means of the Alamouti STBC. Finally, conclusions are drawn in Section VI.

[1] GPRS (General Packet Radio Service) is part of the GSM specifi cations (Phase 2+) and is used for the transfer of packet-switched data.

II. Structure of the Enhanced System

The compatible enhancement of the GSM/GPRS system by means of the Alamouti scheme shall be carried out in a way that preferably few changes have to be applied to current specifications [4]. In this context, the GSM/GPRS binary Gaussian minimum shift keying (GMSK) modulation scheme and the burst structure are retained in the extended system. Throughout this paper, the equivalent complex baseband notation is used.

In the Alamouti scheme, pairs $[x(k), x(k+1)]$ of M-ary data symbols are transmitted over two antennas using two consecutive time instants. In this context, a mapping of the symbols $x(k)$ and $x(k+1)$ is performed according to the *Alamouti matrix* $\mathbf{A}$ (throughout this paper the transposed of the original matrix [3] is used):

$$\mathbf{A} \doteq \begin{array}{cc} \begin{bmatrix} x(k) & -x^*(k+1) \\ x(k+1) & x^*(k) \end{bmatrix} & \begin{array}{l} \longleftarrow \text{Time index } k \\ \longleftarrow \text{Time index } k+1 \end{array} \\ \begin{array}{cc} \uparrow & \uparrow \\ \text{Ant. 1} & \text{Ant. 2} \end{array} & \end{array} \tag{1}$$

where $(.)^*$ denotes complex conjugation. The Alamouti matrix is *unitary* with $\mathbf{A}^H\mathbf{A} = (|x(k)|^2 + |x(k+1)|^2)\,\mathbf{I}_2$, where $\mathbf{A}^H$ is the Hermitian conjugate of $\mathbf{A}$ and $\mathbf{I}_2$ the (2×2)-identity matrix. Fig. 1 shows the proposed transmitter structure of the GSM/GPRS system enhanced by the Alamouti scheme. First, channel coding and interleaving is performed according to one of the GPRS coding schemes CS 1-4, yielding $2 \cdot 58$ data symbols $x(k) \in \{\pm1\}$ per burst. Then, the data symbols are space-time processed according to (1) and mapped on GSM/GPRS bursts. In contrast to the (1x1)-system, two *different* training sequences TS1 (first transmit antenna) and TS2 (second transmit antenna) are mandatory here. Finally, linearized GMSK pulse shaping is done (including a symbol-wise phase rotation

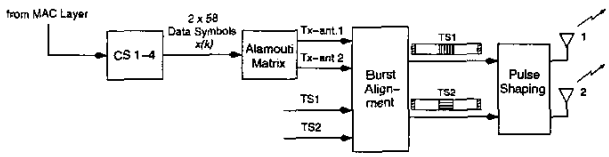

Fig. 1. Transmitter structure of the GSM/GPRS system enhanced by the Alamouti scheme ($n_T = 2$).

of $\pi/2$ rad/symbol), and the modulated signals are transmitted over the two antennas. Note that the single transmit antenna case is included in the enhanced structure as the special case, when the second transmit antenna is switched off. The data rate of the (1x1)-system is retained.

The same general receiver structure can be applied both for the (1x1)-system and for the enhanced system. The received signal is first filtered, then sampled at time instants kT and derotated yielding received symbols $y(k)$. A channel estimator provides estimates of the coefficients of the equivalent discrete-time channel model, here referred to as channel coefficients. Eventually, equalization/detection is performed utilizing the channel estimates. Throughout this paper we consider a Max-Log-MAP algorithm [5], which is a soft-output equalizer/detector approximating the log-likelihood ratios (LLRs) $L(x(k))$ of the transmitted data symbols $x(k)$.

The system described above applies, in principle, as well for EDGE/ EGPRS[2], when the GPRS coding schemes CS 1-4 are replaced by the EGPRS modulation and coding schemes MCS 1-9. The data symbols $x(k)$ are then either binary or 8-ary.

III. Data Detection for the Alamouti Scheme

In order to detect the transmitted data symbols, the equalizer/detector employed in the receiver requires knowledge about the channel coefficients. In a GSM receiver, channel estimation (CE) is typically performed by means of the *correlation method* [11]. In this method, the channel estimates are obtained by correlating the employed training sequence(s) with the corresponding received symbols. In the (1x1)-system solely the *autocorrelation* properties of the training sequence used are crucial for the quality of the resulting estimates (if co-channel interference is neglected). The 8 GSM sequences [4] are characterized by a *perfect* cyclic autocorrelation for any shift ξ with $|\xi| \leq 5$ and thus grant – with respect to the average estimation error – *optimal* channel estimates $\hat{\mathbf{h}} = [\hat{h}^{(0)}, \hat{h}^{(1)}, ..., \hat{h}^{(L)}]$ for channels with memory length $L \leq 5$.

However, in the case of the Alamouti scheme, where two different training sequences TS1 and TS2 are required (cf. Fig. 1), not only the cyclic autocorrelation of *both* sequences determines the quality of the channel estimates, but as well their *cross-correlation* properties. The correlation of TS1 and TS2 with the received midamble symbols yields channel estimates $\hat{\mathbf{h}}_1$ and $\hat{\mathbf{h}}_2$, respectively, associated with the transmission path corresponding to the first and the second transmit antenna.

The ETSI/3GPP specifications [4] do not comprise any partner sequences with respect to the eight GSM sequences at all. Pairs of GSM sequences are in fact characterized by comparably poor cross-correlation properties. Moreover, the existing eight training sequences are already invariably required in the single transmit antenna case, due to the cellular structure of a GSM radio network. Therefore, it is required to define adequate training sequence pairs for the Alamouti case.

[2] EDGE (Enhanced Data Rates for GSM Evolution) is a further development of GSM towards 3G data rates. The EDGE specifi cations comprise both binary GMSK and linearized GMSK with an 8-PSK (phase shift keying) mapping. The GSM burst structure as well as the eight training sequences have been retained for EDGE. EGPRS stands for Enhanced GPRS.

A. Definition of Appropriate Training Sequence Pairs

In Table I, optimized partner sequences with respect to the eight GSM training sequences are listed. These sequences have a perfect cyclic autocorrelation for shifts $|\xi| \leq 5$, the same length as the GSM sequences (26 symbols), and the same cyclic structure 'A B C A B'. Among all sequences with these properties, the ones given in Table I yield a minimum estimation error in the case of a typical urban wireless channel ($L = 3$). In the case of a channel memory length $L \leq 2$, even optimal CE is accomplished in the sense that the resulting sequence pairs are characterized by a *perfect* cross-correlation for shifts $|\xi| \leq 2$. Simulation results indicate that – with regard to the resulting bit error performance of the overall system – the proposed sequence pairs significantly outperform any pair of GSM sequences (refer to Section V-B). As a matter of fact, the performance accomplished on a typical urban channel is very close to the case of perfect cross-correlation.

B. Generalization of the Max-Log-MAP Algorithm

The conventional Max-Log-MAP algorithm [5] has to be modified in a way that – corresponding to the Alamouti transmitter structure – the received symbols are processed as *pairs*. For reasons of brevity, we will not recapitulate the Max-Log-MAP algorithm here, but rather outline the main differences between the conventional algorithm and the generalized Max-Log-MAP algorithm required in the Alamouti case.

In Fig. 2 a comparison is drawn between the conventional trellis and the trellis resulting for the Alamouti scheme with regard to trellis states and transitions (channel memory length $L = 4$ in both cases). In the first case, a trellis segment complies with a single data symbol $x(k)$. Presuming a single receive antenna ($n_R = 1$), the *Euclidean branch metric* $\mu(\mathcal{S}(k), \mathcal{S}(k+1))$ for a transition $\mathcal{S}(k) \rightarrow \mathcal{S}(k+1)$ is given by $|y(k) - \tilde{y}(k)|^2$, where $y(k)$ denotes the current received symbol and $\tilde{y}(k)$ a corresponding hypothesis. $\tilde{y}(k)$ is a replica of $y(k)$, determined by the state $\mathcal{S}(k)$, the symbol hypothesis $\tilde{x}(k)$ associated with the transition regarded and the (estimated) channel coefficients at time index k. The number of possible trellis states $\mathcal{S}(k)$ is M^L, where M is the cardinality of the symbol alphabet.

In the Alamouti scheme, a single trellis segment spans *two* consecutive time indices k and $k+1$. A transition from a state $\mathcal{S}(k)$ to a subsequent state $\mathcal{S}(k+2)$ therefore complies with a pair $[x(k), x(k+1)]$ of data symbols. As in the conventional trellis, the number of possible trellis states $\mathcal{S}(k)$ is given by M^L if L

TABLE I

Optimized partner sequences with respect to the 8 GSM training sequences.

Partner to GSM No.	A	B	C	A	B
1	00100	01001	011100	00100	01001
2	00011	10100	100010	00011	10100
3	01011	01111	000100	01011	01111
4	10111	01110	000100	10111	01110
5	01010	00110	111110	01010	00110
6	00000	10011	101011	00000	10011
7	01111	10110	001010	01111	10110
8	00001	11011	010001	00001	11011

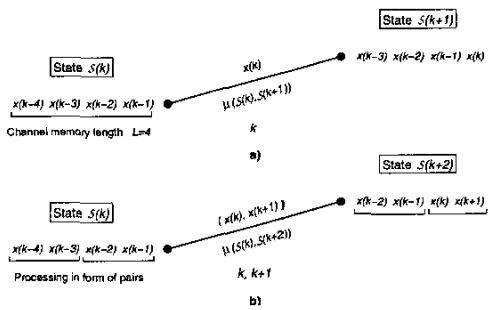

Fig. 2. State transition a) within the conventional trellis (channel memory length $L = 4$) and b) within the Alamouti-trellis ($L = 4$)

is even. This means that both equalizers are of the same complexity[3]. The Euclidean branch metric $\mu(\mathcal{S}(k), \mathcal{S}(k+2))$ for a transition $\mathcal{S}(k) \to \mathcal{S}(k+2)$ is calculated according to

$$\mu(\mathcal{S}(k), \mathcal{S}(k+2)) = \left|y(k) - \tilde{y}(k)\right|^2 + \left|y(k+1) - \tilde{y}(k+1)\right|^2 . \tag{2}$$

In the sequel, let the estimated channel coefficients be denoted as $\hat{\mathbf{h}}_1 = [\hat{h}_1^{(0)}, ..., \hat{h}_1^{(L)}]$ and $\hat{\mathbf{h}}_2 = [\hat{h}_2^{(0)}, ..., \hat{h}_2^{(L)}]$. The hypotheses $\tilde{y}(k)$ and $\tilde{y}(k+1)$ are then calculated as follows:

$$\tilde{y}(k) = \sum_{\kappa=0}^{L} \left[\hat{h}_1^{(\kappa)}\, \tilde{x}(k-\kappa) \;-\; \hat{h}_2^{(\kappa)}\, \tilde{x}^*(k-(\kappa-1)) \Big|_{\kappa \text{ even}} + \hat{h}_2^{(\kappa)}\, \tilde{x}^*(k-(\kappa+1)) \Big|_{\kappa \text{ odd}} \right] \quad \text{and}$$

$$\tilde{y}(k+1) = \sum_{\kappa=0}^{L} \left[\hat{h}_1^{(\kappa)}\, \tilde{x}((k+1)-\kappa) + \hat{h}_2^{(\kappa)}\, \tilde{x}^*((k+1)-(\kappa+1)) \Big|_{\kappa \text{ even}} - \hat{h}_2^{(\kappa)}\, \tilde{x}^*((k+1)-(\kappa-1)) \Big|_{\kappa \text{ odd}} \right]. \tag{3}$$

The symbol hypotheses $\tilde{x}(k-L), ..., \tilde{x}(k-1)$ are determined by the state $\mathcal{S}(k)$, and the hypotheses $\tilde{x}(k), \tilde{x}(k+1)$ are determined by the transition under consideration.

In the case of multiple receive antennas, a partial branch metric $\mu_j(\mathcal{S}(k), \mathcal{S}(k+2))$ is calculated for each individual receive antenna $1 \le j \le n_R$, which is done corresponding to (2) on basis of the received values $y_j(k)$, $y_j(k+1)$ and the corresponding hypotheses $\tilde{y}_j(k)$ and $\tilde{y}_j(k+1)$. The hypotheses $\tilde{y}_j(k)$ and $\tilde{y}_j(k+1)$ are obtained according to (3), where each receive antenna j is associated with dedicated vectors $\hat{\mathbf{h}}_{1j}, \hat{\mathbf{h}}_{2j}$ of channel coefficient estimates. The overall branch metric μ is obtained by a sum over the partial metrics μ_j:

$$\mu(\mathcal{S}(k), \mathcal{S}(k+2)) = \sum_{j=1}^{n_R} \mu_j(\mathcal{S}(k), \mathcal{S}(k+2)) . \tag{4}$$

[3] If L is an odd number, M^{L+1} states are required leading to a higher complexity compared to the conventional trellis.

IV. Orthogonal Properties of the Alamouti STBC

The Alamouti scheme has been designed for quasi-static flat fading channels. In the following it is shown that the orthogonal properties of the Alamouti scheme are lost in the case of non-perfect knowledge of the channel coefficients at the receiver as well as in the presence of fast fading or frequency-selective fading. This orthogonality loss will cause a performance degradation. The results obtained in the following hold, in principle, for an arbitrary STBC.

First of all, consider a quasi-static flat fading channel assuming $n_R = 1$ receive antenna. In this case, the two transmission paths from either transmit antenna to the receive antenna can be modeled by means of two complex-valued channel coefficients h_1 and h_2 (channel memory length $L = 0$), which are constant over the duration of an entire GSM burst. Taking into account the symbol mapping according to matrix $\mathbf{A}$, the received symbols $y(k)$ and $y(k+1)$ at the time instants k and $k+1$ are given by the following matrix equation:

$$\underbrace{\begin{bmatrix} y(k) \\ y^*(k+1) \end{bmatrix}}_{\mathbf{y}(k)} = \underbrace{\begin{bmatrix} h_1 & -h_2 \\ h_2^* & h_1^* \end{bmatrix}}_{\mathbf{H}} \underbrace{\begin{bmatrix} x(k) \\ x^*(k+1) \end{bmatrix}}_{\mathbf{x}(k)} + \underbrace{\begin{bmatrix} n(k) \\ n^*(k+1) \end{bmatrix}}_{\mathbf{n}(k)}, \tag{5}$$

where $n(k)$ and $n(k+1)$ denote samples of an additive white Gaussian noise (AWGN) process taken at time index k and $k+1$, respectively. $\mathbf{H}$ is unitary, which is due to the unitarity of the Alamouti matrix $\mathbf{A}$.

The detection of the symbols $x(k)$ and $x(k+1)$ can simply be performed by means of the matrix-vector multiplication $\mathbf{H}^H\mathbf{y}(k)$. In this context – assuming perfect knowledge of h_1 and h_2 at the receiver – the unitarity of $\mathbf{H}$ leads to a decoupling of $x(k)$ and $x(k+1)$ in terms of *independent* estimates $\hat{x}(k)$ and $\hat{x}(k+1)$ (see e.g. [12]):

$$\hat{\mathbf{x}}(k) = \mathbf{H}^H\mathbf{y}(k) = \mathbf{H}^H\mathbf{H}\,\mathbf{x}(k) + \mathbf{H}^H\mathbf{n}(k)\,, \quad \text{where}$$

$$\hat{\mathbf{x}}(k) = \begin{bmatrix} \hat{x}(k) \\ \hat{x}^*(k+1) \end{bmatrix} \quad \text{and}$$

$$\mathbf{H}^H\mathbf{H} = (|h_1|^2 + |h_2|^2)\,\mathbf{I}_2\,. \tag{6}$$

Computing $\hat{\mathbf{x}}(k)$ according to (6) is equivalent to minimizing the metric (2) for $L = 0$. Note that due to the diagonal structure of $\mathbf{H}^H\mathbf{H}$, the desired symbols are always combined in a constructive way, because they are multiplied by a sum of absolute terms. The noise, however, is combined incoherently (matrix $\mathbf{H}^H$), which leads to a diversity gain over a (1x1)-system.

A. Orthogonality Loss due to Non-Perfect Channel Estimation

Let $\hat{h}_1$ and $\hat{h}_2$ denote noisy estimates of the channel coefficients h_1 and h_2, where $\hat{h}_1 \doteq h_1 + \varepsilon_1$ and $\hat{h}_2 \doteq h_2 + \varepsilon_2$. The detection of the symbols $x(k)$ and $x(k+1)$ will then be performed using

$$\hat{\mathbf{H}} = \begin{bmatrix} \hat{h}_1 & -\hat{h}_2 \\ \hat{h}_2^* & \hat{h}_1^* \end{bmatrix}, \tag{7}$$

i.e., (6) becomes

$$\hat{\mathbf{x}}(k) = \hat{\mathbf{H}}^H\mathbf{H}\,\mathbf{x}(k) + \hat{\mathbf{H}}^H\mathbf{n}(k)\,. \tag{8}$$

In contrast to (6), $\hat{\mathbf{H}}^H\mathbf{H}$ does not yield a diagonal matrix:

$$\hat{\mathbf{H}}^H\mathbf{H} = \begin{bmatrix} \psi & \varsigma^* \\ -\varsigma & \psi^* \end{bmatrix},$$

$$\text{where} \quad \psi = |h_1|^2 + |h_2|^2 + \varepsilon_1^* h_1 + \varepsilon_2 h_2^* \qquad (9)$$

$$\text{and} \quad \varsigma = \varepsilon_2^* h_1 - \varepsilon_1 h_2^* .$$

Despite this loss of orthogonality, the desired symbols $x(k)$ and $x(k+1)$ are still combined in a constructive way. They are, however, weighted with different factors, as the diagonal elements of $\hat{\mathbf{H}}^H\mathbf{H}$ are not equal. The secondary diagonal elements $-\varsigma$ and ς^* introduce an additional error term to the estimates $\hat{x}(k)$ and $\hat{x}(k+1)$. Note that for $\varepsilon_1, \varepsilon_2 \to 0$, (8) reduces to (6).

B. Orthogonality Loss due to Fast Fading

In the case of fast fading, the channel coefficients vary over the duration of a symbol pair $[x(k), x(k+1)]$. In this context, let $h_1(k)$ and $h_2(k)$ denote the channel coefficients at time index k, and $h_1(k+1) \doteq h_1(k) + \Delta_1(k)$ and $h_2(k+1) \doteq h_2(k) + \Delta_2(k)$ the ones at time index $k+1$. Accordingly, the channel matrix is given by

$$\tilde{\mathbf{H}}(k) = \begin{bmatrix} h_1(k) & -h_2(k) \\ h_2^*(k+1) & h_1^*(k+1) \end{bmatrix}, \qquad (10)$$

i.e., (6) becomes

$$\hat{\mathbf{x}}(k) = \tilde{\mathbf{H}}(k)^H \tilde{\mathbf{H}}(k)\, \mathbf{x}(k) + \tilde{\mathbf{H}}(k)^H \mathbf{n}(k), \qquad (11)$$

provided that the channel coefficients are perfectly known at the receiver both for time index k and for time index $k+1$. The matrix $\tilde{\mathbf{H}}(k)^H\tilde{\mathbf{H}}(k)$ does not have a diagonal structure:

$$\tilde{\mathbf{H}}(k)^H\tilde{\mathbf{H}}(k) = \qquad (12)$$

$$\begin{bmatrix} |h_1(k)|^2 + |h_2(k) + \Delta_2(k)|^2 & \zeta^* \\ \zeta & |h_1(k) + \Delta_1(k)|^2 + |h_2(k)|^2 \end{bmatrix},$$

$$\text{where} \quad \zeta = h_1(k)\Delta_2^*(k) + h_2^*(k)\Delta_1(k) + \Delta_1(k)\Delta_2^*(k).$$

As in Section IV-A, the desired symbols $x(k)$ and $x(k+1)$ are still combined in a constructive way but weighted with different factors. The secondary diagonal elements ζ and ζ^* introduce an additional error term to the estimates $\hat{x}(k)$ and $\hat{x}(k+1)$. Note that for $\Delta_1(k)$, $\Delta_2(k) \to 0$, (11) reduces to (6).

C. Orthogonality Loss due to Frequency-Selective Fading

In the case of a quasi-static frequency-selective fading channel with memory length L, the two transmission paths from either transmit antenna to the receive antenna can be described by means of two complex-valued channel coefficient vectors $\mathbf{h}_1 = [h_1^{(0)}, h_1^{(1)}, ..., h_1^{(L)}]$ and $\mathbf{h}_2 = [h_2^{(0)}, h_2^{(1)}, ..., h_2^{(L)}]$. If a channel memory length of $L = 1$ is considered, (5) becomes

$$\mathbf{y}(k) = \qquad (13)$$

$$\mathbf{H}_0\, \mathbf{x}(k) + \mathbf{H}_{\mathrm{ISI}\,1}\, \mathbf{x}^*(k) + \mathbf{H}_{\mathrm{ISI}\,2}\, \mathbf{x}^*(k-2) + \mathbf{n}(k), \text{ where}$$

$$\mathbf{H}_0 = \begin{bmatrix} h_1^{(0)} & -h_2^{(0)} \\ h_2^{(0)*} & h_1^{(0)*} \end{bmatrix}, \mathbf{H}_{\mathrm{ISI}\,1} = \begin{bmatrix} 0 & 0 \\ h_1^{(1)*} & -h_2^{(1)*} \end{bmatrix}$$

$$\text{and } \mathbf{H}_{\mathrm{ISI}\,2} = \begin{bmatrix} h_2^{(1)} & h_1^{(1)} \\ 0 & 0 \end{bmatrix}.$$

The orthogonal properties of the Alamouti STBC are therefore lost in the presence of ISI (in this example represented by $\mathbf{H}_{\mathrm{ISI}\,1,2}$). Note that for $h_1^{(1)}, h_2^{(1)} = 0$, (13) reduces to (5).

V. Simulation Results

The simulation results presented next demonstrate the performance of the GSM/GPRS system enhanced by the Alamouti scheme for the example of a GSM 05.05 typical urban (TU) wireless channel [4], which is a frequency-selective channel model with a memory length of $L_{TU} \approx 3$. In particular, the application of different training sequence pairs is addressed. The simulations are restricted to the transmission of *uncoded* data, i.e., channel coding is not performed. Two transmit and up to two receive antennas are considered. In this context, the overall transmitter power is the same as for a single transmit antenna, i.e., the transmission power is normalized by a factor 1/2 at each antenna. Normalization with respect to the number n_R of receive antennas has not been performed. The receiver filter used is a non-adaptive root-raised-cosine filter with roll-off factor $r = 0.5$ and 3dB-bandwidth $f_{3\mathrm{dB}} = 180$kHz. The employed Max-Log-MAP equalizer/detector has $2^{L_{TU}} = 8$ states in the case of the (1x1)-system. In the Alamouti case $2^{L_{TU}+1} = 2^4$ states are required, since L_{TU} is an odd number.
All simulation results presented in the following were obtained by means of Monte Carlo simulations over 10.000 bursts.

A. Performance on the Typical Urban Channel

The bit error performance of the enhanced GSM/GPRS system on the time-invariant typical urban channel (TU0) is shown in Fig. 3. The channel coefficients are perfectly known at the receiver. The bit error rate (BER) curves are plotted as a function of the average signal-to-noise ratio (SNR). As a reference, the BER curve resulting for the (1x1)-system is included. Moreover, analytical curves for *diversity reception* of uncoded BPSK are included [13], where data symbols $x(k) \in \{\pm 1\}$ are transmitted over ν individual paths subject to *independent* Rayleigh fading and characterized by an identical average SNR of $(E_s/N_0)/\nu$ (E_s denotes the average energy per data symbol and N_0 denotes the single-sided noise power density).
As shown in Fig. 3, the Alamouti scheme yields significant performance improvements upon the (1x1)-system. The gain accomplished at a BER of 10^{-3} is about 1.5 dB, when a single receive antenna is employed ((2x1)-Alamouti) and about 9 dB in the case of a second receive antenna ((2x2)-Alamouti).
Further investigations showed that in the case of a time-varying typical urban channel, very little performance loss occurs as long as the mobile station moves with a speed $\leq$ 50 km/h, @ 900 MHz. At a BER of 10^{-3}, the degradation resulting for 50 km/h and $n_R = 1$ receive antenna is smaller than 0.4 dB. Generally, the enhanced system becomes even more robust when a second receive antenna is employed.

B. Choice of Different Training Sequence Pairs

Fig. 4 demonstrates the importance of an appropriate training sequence pair in the case of the Alamouti scheme ($n_R = 1$ receive antenna). Three different cases of training sequence pairs are considered, where each time GSM training sequence No.1 (GSM TS No.1) is employed at the first transmit antenna. In the first case, the optimized partner sequence with respect to GSM TS No.1 is employed at the second transmit antenna (cf. Table I). In the second and the third case, the pairs (GSM TS No.1 + No.3) and (GSM TS No.1 + No.2) are considered. The channel coefficient estimates are obtained by means of the correlation method. The estimates are not corrupted by additional noise. The performance degradations shown in Fig. 4 are therefore solely due to non-perfect cross-correlation properties of the

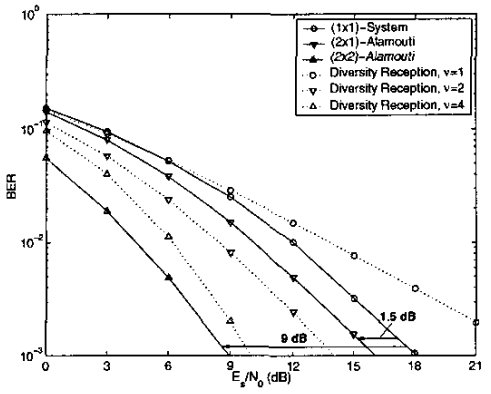

Fig. 3. Performance of the Alamouti scheme on the TU0 channel, perfect knowledge of channel coeffi cients at the receiver.

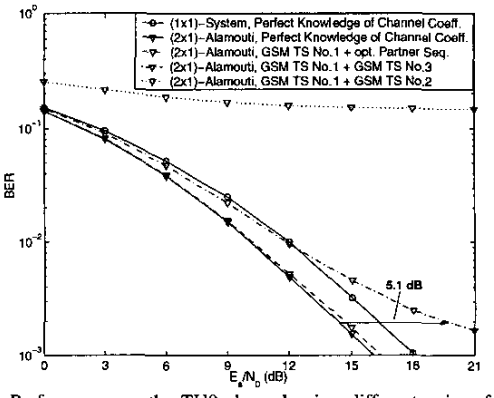

Fig. 4. Performance on the TU0 channel using different pairs of training sequences (channel estimation by means of correlation method (no AWGN)).

employed sequence pairs.

The application of the optimized partner sequence leads to a very small performance deterioration with respect to *perfect* knowledge of the channel coefficients at the receiver, which is about 0.3 dB at a BER of $2 \cdot 10^{-3}$. The pair (GSM TS No.1 + No.3) is the *best* choice among all possible pairs of GSM sequences in the sense of a minimum average channel estimation error for large SNR values (TU-profile). Still, its application leads to a performance degradation of about 5.1 dB at the same BER. The optimized partner sequence of Table I therefore significantly outperforms any pair of GSM sequences. Note that for a desired BER < 0.01, the (2x1)-Alamouti scheme employing the training sequence pair (GSM TS No.1 + No.3) is even less efficient than the (1x1)-system. The pair (GSM TS No.1 + No.2) turns out to be a particularly bad choice and leads to further significant performance deterioration. In this case the (2x1)-Alamouti scheme does not accomplish a BER ≤ 0.1.

When the optimized partner sequence is employed, noisy channel estimates cause a performance loss of 1.7 dB ($n_R = 1$) and 1.9 dB ($n_R = 2$) at a BER of 10^{-3}. For comparison, in the case of a (1x1)-system a degradation of about 1 dB results.

VI. Summary and Conclusions

The application of the Alamouti scheme in a GSM/GPRS system has been investigated. First of all, the structure of the enhanced GSM/GPRS system was presented, which is compatible with current specifications. A modified Max-Log-MAP equalizer/detector was introduced as well as appropriate pairs of training sequences, consisting of one of the eight GSM sequences and a corresponding optimized partner sequence. Simulation results for a typical urban environment showed that the proposed sequence pairs yield a system performance very close to the optimal case. In fact, they significantly outperform any pair of GSM sequences. In principle, the proposed system structure applies as well for the EDGE system. However, in the case of 8-PSK the detection complexity becomes impractical, and reduced-state techniques are required.

It was shown that the orthogonal properties of STBC are lost in the case of non-perfect knowledge of the channel coefficients at the receiver as well as in the presence of fast and/or frequency-selective fading, which will result in a performance loss.

On the basis of simulation results for a typical urban wireless channel, it was demonstrated that the Alamouti scheme yields significant performance improvements upon a single transmit antenna system. This is especially true in the case of two receive antennas, where a performance improvement of 9 dB was accomplished. The Alamouti STBC is therefore an attractive technique for an extension of the current GSM/GPRS specifications with regard to wireless services of high bit rates. The enhanced system is robust to vehicular speeds not significantly faster than 50 km/h (at a carrier frequency of 900 MHz). It is, however, more sensitive to non-perfect knowledge of the channel coefficients at the receiver than the (1x1)-system.

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References

[1] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: Performance criterion and code construction," *IEEE Trans. Inform. Theory*, vol. 44, no. 2, pp. 744-765, Mar. 1998.

[2] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Trans. Inform. Theory*, vol. 45, no. 5, pp. 1456-1467, June 1999.

[3] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1451-1458, Oct. 1998.

[4] 3GPP, "Digital cellular communications system (Phase 2+)," Technical Specifi cations 3GPP TS 05.01-05, 2001.

[5] P. Robertson, P. Hoeher, and E. Villebrun, "Optimal and sub-optimal maximum a posteriori algorithms suitable for turbo decoding," *Europ. Trans. Telecommun.*, vol. 8, no. 2, pp. 119-125, Mar./Apr. 1997.

[6] M. Coupechoux and V. Braun, "Space-time coding for the EDGE mobile radio system," in *Proc. IEEE Int. Conf. Pers. Wireless Commun.*, 2000, pp. 28-32.

[7] R. Srinivasan, M. J. Heikkila, and R. Pirhonen, "Performance evaluation of space-time coding for EDGE," in *Proc. IEEE Int. Conf. Commun.*, 2001, pp. 3056-3060.

[8] K. Zangi, D. Hui, and J.-F. Cheng, "Physical layer issues for deploying transmit diversity in GPRS/EGPRS networks," in *Proc. IEEE Veh. Technol. Conf. Fall'01*, 2001, pp. 538-542.

[9] C. Budianu and L. Tong, "Channel estimation for space-time orthogonal block codes," in *Proc. IEEE Int. Conf. Commun.*, 2001, pp. 1127-1131.

[10] W. J. Choi and J. M. Cioffi, "Multiple input/ multiple output (MIMO) equalization for space-time block coding," in *Proc. IEEE Pacific Rim Conf. Commun.*, 1999, pp. 341-344.

[11] A. Baier, "Correlative and iterative channel estimation in adaptive Viterbi equalizers for TDMA mobile radio systems," in *ITG-Fachbericht 107*, 1989, pp. 363-368.

[12] G. Bauch, *"Turbo-Entzerrung" und Sendeantennen-Diversity mit "Space-Time-Codes" im Mobilfunk.* PhD thesis, Department of Communications Engineering, Munich University of Technology, 2001 (in German).

[13] J. G. Proakis, *Digital Communications.* 4th ed., New York: McGraw-Hill, 2001.