SYMBOL–BY–SYMBOL AND TRELLIS–BASED EQUALIZATION WITH WIDELY LINEAR PROCESSING FOR SPACE–TIME BLOCK–CODED TRANSMISSION OVER FREQUENCY–SELECTIVE FADING CHANNELS

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***Abstract* — In this paper, we develop receiver concepts for transmission with space–time block codes (STBC's) over frequency–selective fading channels. The focus lies on Alamouti's STBC, but the results may be generalized to related STBC's. We show that a straightforward combination of conventional equalizers and a space–time block decoder is only possible if at least as many receive antennas as transmit antennas are employed, but not for the practically interesting case of pure transmit diversity. This restriction is circumvented by our approach. Equalizers with *widely linear (WL) processing* are designed, utilizing the structural properties of the transmit signal of space–time block coding, which is shown to be *improper (rotationally variant)*. These schemes are especially suited for equalization of high–level modulated signals, which are used in third–generation time–division multiple access mobile communications standards such as EDGE (Enhanced Data Rates for GSM Evolution).**

1. Introduction

Space–time coding (STC) has been introduced in order to be able to combat fading via diversity also in such situations where multiple transmit antennas but only a single receive antenna can be employed. This often is the case for downlink transmission (base station to mobile station). In literature, space–time trellis codes (STTC's) and space–time block codes (STBC's) have been proposed for systems with transmit diversity. Both classes of codes have been originally designed for flat fading scenarios.

Recently, several STBC coding schemes based on block processing have been proposed for frequency–selective fading channels producing intersymbol interference [1, 2, 3]. These schemes are well suited for block fading channels and yield a high gain over single–input single–output (SISO) transmission in this case. However, their extension to channels with time–variant behaviour inside a transmission block does not seem to be straightforward. Therefore, the application of a symbol–based STBC like Alamouti's code [4] along with equalization at the receiver side, as has been proposed e.g. in [5, 6], may be preferable. The approaches in [5, 6], however, are not applicable if more transmit than receive antennas are used. In this paper, we show that space–time block coding using Alamouti's or related schemes generates an *improper (rotationally variant)* [7] transmit signal, i.e., a signal with non-vanishing *pseudoautocorrelation*. Hence, *widely linear equalization* is recommendable, which has been recently introduced in a different context [8]. A widely linear equalizer (LE), a decision–feedback equalizer (DFE) with widely linear processing, and a delayed decision–feedback sequence estimator (DDFSE) with widely linear prefiltering are described. The latter is developed starting from a maximum–likelihood sequence estimation approach for the given problem. In contrast to the proposed concepts, conventional suboptimum equalizers without widely linear processing cannot yield a high performance for space–time block–encoded transmission without receive diversity.

2. System Model

In the following, the equivalent complex baseband representation of space–time block–encoded transmission over a frequency–selective fading channel is described. Here, we restrict ourselves to Alamouti's STBC [4] for most derivations for the sake of simplicity. Nevertheless, the presented receiver concepts can be generalized in a straightforward manner to related orthogonal STBC's. In Alamouti's scheme, $N_T = 2$ transmit antennas are employed. Two independent, identically distributed (i.i.d.) sequences $c_1[\mu]$, $c_2[\mu]$ of real– or complex–valued coefficients taken from an M–ary signal set $\mathcal{C}$ (e.g. $\mathcal{C} = \{\pm 1/\sqrt{2}\}$ for binary phase–shift keying (BPSK) or $\mathcal{C} = \{1/\sqrt{2}\,\mathrm{e}^{\frac{j\,2\pi\nu}{8}} \mid \nu \in \{0, 1, \ldots, 7\}\}$ for 8PSK[1]) with variance σ_c^2 are fed into the encoder, which generates transmit sequences $s_1[k]$, $s_2[k]$ according to $s_1[k] = c_1[\mu]$, $k = 2\,\mu$, $s_1[k] = -c_2^*[\mu]$, $k = 2\,\mu + 1$, $s_2[k] = c_2[\mu]$, $k = 2\,\mu$, $s_2[k] = c_1^*[\mu]$, $k = 2\,\mu + 1$ ($(\cdot)^*$: complex conjugation). Because of the repetition of signal points, the processes $s_1[\cdot]$, $s_2[\cdot]$ are jointly cyclostationary with period 2. Therefore, it is convenient to organize the *polyphase components* of $s_1[\cdot]$, $s_2[\cdot]$ in a vector

$$\boldsymbol{s}[\mu] \stackrel{\triangle}{=} [\, s_1[2\,\mu] \quad s_2[2\,\mu] \quad s_1[2\,\mu+1] \quad s_2[2\,\mu+1]\,]^T \quad (1)$$

[1]In general, a normalization factor of $1/\sqrt{N_T}$ is included in the signal set in order to guarantee that the total transmit power is equal for space–time encoded and uncoded transmission.

($(\cdot)^T$: transposition). $\boldsymbol{s}[\cdot]$ is a stationary vector process with autocorrelation

$$\boldsymbol{\Phi}_{ss}[\kappa] \stackrel{\triangle}{=} E\{\boldsymbol{s}[\mu]\,\boldsymbol{s}^H[\mu-\kappa]\} = \delta[\kappa]\cdot \left(\sigma_c^2\,\boldsymbol{I}_4 + \begin{bmatrix} 0 & 0 & 0 & \xi_c^2 \\ 0 & 0 & -(\xi_c^2) & 0 \\ 0 & -(\xi_c^2)^* & 0 & 0 \\ (\xi_c^2)^* & 0 & 0 & 0 \end{bmatrix}\right), \quad (2)$$

where $E\{\cdot\}$, $(\cdot)^H$, $\delta[\cdot]$, and $\boldsymbol{I}_K$ denote expectation, Hermitian transposition, the unit pulse sequence, and the $K \times K$ identity matrix, respectively. In addition, the definition $\xi_c^2 \stackrel{\triangle}{=} E\{(c_1[\mu])^2\} = E\{(c_2[\mu])^2\}$ has been used in (2). E.g., $\xi_c^2 = 1/2$ for BPSK and $\xi_c^2 = 0$ for M–ary PSK with $M > 2$. For complex random processes, the *pseudoautocorrelation* [7] $\boldsymbol{\Phi}_{ss^*}[\kappa] \stackrel{\triangle}{=} E\{\boldsymbol{s}[\mu]\,\boldsymbol{s}^T[\mu-\kappa]\}$ is a further important quantity containing information about the statistical properties of the process. For vanishing pseudoautocorrelation, the process is referred to as *proper (rotationally invariant)* [7]. This holds for most complex random processes in digital communications. Interestingly, $\boldsymbol{\Phi}_{ss^*}[0] \neq \boldsymbol{0}_4$ for the transmit vector of Alamouti's scheme (and most other relevant space–time block coding schemes):

$$\boldsymbol{\Phi}_{ss^*}[\kappa] = \delta[\kappa]\cdot \left(\sigma_c^2 \begin{bmatrix} 0 & 0 & 0 & 1 \\ 0 & 0 & -1 & 0 \\ 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{bmatrix} + \begin{bmatrix} \xi_c^2 & 0 & 0 & 0 \\ 0 & \xi_c^2 & 0 & 0 \\ 0 & 0 & (\xi_c^2)^* & 0 \\ 0 & 0 & 0 & (\xi_c^2)^* \end{bmatrix}\right). \quad (3)$$

Hence, $\boldsymbol{s}[\cdot]$ is an *improper (rotationally variant)* process [7] for *any* real or complex signal constellation $\mathcal{C}$. For improper transmit signals and/or channel noise, performance of a linear receiver can be improved by *widely linear* (WL) processing [9]. In a WL receiver, the received sequence and its complex conjugate are processed jointly. In [8], the concept of widely linear estimation has been applied to equalization for transmission of real–valued data over single–input single–output (SISO) ISI channels with complex–valued coefficients. In this paper, we utilize the concept of WL processing for low–complexity equalization of space–time block–encoded transmission over multiple–input multiple–output (MIMO) channels.

At the receiver, N_R antennas are employed. Here, we pay special attention to the case $N_R = 1$, which is most relevant for space–time coding. The discrete–time received signal at antenna m, $m \in \{1, \ldots, N_R\}$, may be written as

$$r_m[k] = \sum_{i=1}^{N_T} \sum_{\kappa=0}^{q_h} h_{im}[\kappa]\, s_i[k-\kappa] + n_m[k], \quad (4)$$

where $h_{im}[\cdot]$ denotes the causal discrete–time impulse response of (maximum) order q_h characterizing transmission from transmit antenna i to receive antenna m including continuous–time transmit and receiver input filtering. $n_m[\cdot]$, $m \in \{1, \ldots, N_R\}$, are mutually independent proper complex additive white Gaussian noise processes with equal powers, $E\{|n_m[k]|^2\} = \sigma_n^2,\ \forall m$.

3. Widely Linear Equalization

In a direct approach to receiver design, equalization and space-time block decoding may be separated, cf. e.g. [5]. First, the matrix channel is equalized into a temporal ISI–free channel impulse response. Assuming linear equalization, in this first stage an $N_T \times N_R$ linear MIMO filter with transfer matrix $\boldsymbol{F}(z)$ is employed, whose coefficients are selected according to the minimum mean–squared error (MMSE) or the zero–forcing (ZF) criterion. Spatial ISI is retained in the equalizer output signal in order to achieve a diversity gain with the subsequent space–time block decoding [5].

ZF equalization is possible if and only if

$$\exists z \in \mathbb{C}\backslash\{0,\infty\} \text{ with } \mathrm{rank}(\boldsymbol{H}(z)) = N_T, \quad (5)$$

where $\boldsymbol{H}(z)$ denotes the MIMO channel transfer matrix corresponding to the impulse responses $h_{im}[\cdot]$. Otherwise, the ZF condition

$$\boldsymbol{F}(z)\ \boldsymbol{H}(z) = \boldsymbol{I}_{N_T} \quad (6)$$

cannot be fulfilled. Eq. (5) is not valid for pure transmit diversity ($N_T > 1$, $N_R = 1$). Hence, ZF equalization is not possible in this case using the direct approach. On the other hand, also MMSE equalization usually yields a poor performance if a ZF equalizer does not exist.

In our approach, the space–time block encoder and the channel are viewed as an equivalent overall channel at the receiver side, for which a WL equalizer is designed for direct reconstruction of the sequences $c_1[\cdot]$, $c_2[\cdot]$. For this, the cyclostationarity of the received signal has to be taken into account. Time–invariant equalizer filters are obtained if a vector containing the polyphase components of each received sequence is taken as equalizer input. Because we apply WL processing in order to utilize the rotational variance of the transmit signal, this vector is augmented by the complex conjugates of the polyphase components. Hence, as input vectors for WL equalization, we define

$$\tilde{\boldsymbol{r}}_m[\mu] \stackrel{\triangle}{=} [\, r_m[2\,\mu] \quad r_m[2\,\mu+1] \quad r_m^*[2\,\mu] \quad r_m^*[2\,\mu+1]\,]^T. \quad (7)$$

Each sequence $\tilde{\boldsymbol{r}}_m[\cdot]$ represents a stationary vector process, and

$$\tilde{\boldsymbol{r}}_m[\mu] = \sum_{\nu=0}^{\tilde{q}_h} \tilde{\boldsymbol{H}}_m[\nu]\,\tilde{\boldsymbol{c}}[\mu-\nu] + \tilde{\boldsymbol{n}}_m[\mu], \quad (8)$$

with

$$\tilde{\boldsymbol{c}}[\mu] = [\, c_1[\mu] \quad c_2[\mu] \quad c_1^*[\mu] \quad c_2^*[\mu]\,]^T, \quad (9)$$

$$\tilde{\boldsymbol{n}}_m[\mu] = [\, n_m[2\,\mu]\ n_m[2\,\mu+1]\ n_m^*[2\,\mu]\ n_m^*[2\,\mu+1]\,]^T \quad (10)$$

The coefficient matrices $\tilde{\boldsymbol{H}}_m[\nu]$ depend in a straightforward way on the channel coefficients $h_{im}[\cdot]$. The channel orders of equivalent overall and original channel are related by $\tilde{q}_h = \lceil \frac{q_h}{2} \rceil$ ($\lceil x \rceil$: smallest integer $\geq x$). Finally,

$$\tilde{\boldsymbol{r}}[\mu] \stackrel{\triangle}{=} [\, \tilde{\boldsymbol{r}}_1^T[\mu] \ \ldots\ \tilde{\boldsymbol{r}}_{N_R}^T[\mu]\,]^T = \sum_{\nu=0}^{\tilde{q}_h} \tilde{\boldsymbol{H}}[\nu]\,\tilde{\boldsymbol{c}}[\mu-\nu] + \tilde{\boldsymbol{n}}[\mu] \quad (11)$$

is obtained, with obvious definitions of $\tilde{\boldsymbol{H}}[\nu]$ and $\tilde{\boldsymbol{n}}[\mu]$. Eq. (11) represents a MIMO system with 4 inputs and $4\,N_R$ outputs, with a $4\,N_R \times 4$ transfer matrix $\tilde{\boldsymbol{H}}\,(z)$. For all cases of practical interest, the subchannel polyphase components are sufficiently different in order to guarantee that $\tilde{\boldsymbol{H}}\,(z)$ has rank 4, i.e., a ZF equalizer $\tilde{\boldsymbol{F}}\,(z)$ with $\tilde{\boldsymbol{F}}\,(z)\;\tilde{\boldsymbol{H}}\,(z) = \boldsymbol{I}_4$ exists. Especially, for $N_R = 1$ an equalizable 4×4 MIMO system arises. Hence, for any $N_R \geq 1$ WL ZF equalization is possible. Therefore, it can be also expected that MMSE equalization performs well.

It can be shown that for a flat fading channel matrix with $N_R = 1$, widely linear ZF equalization becomes equivalent to the standard detection rule for orthogonal STBC's like Alamouti's code [4]. Hence, this rule arises naturally in a WL processing context.

A similar approach has been recently presented for space–time coded code–division multiple access (CDMA) transmission [10].

4. Widely Linear Decision–Feedback Equalization

Performance of a linear equalizer can be improved by employing noise prediction or, equivalently, decision feedback. A MIMO decision–feedback equalizer (DFE) consists of a feedforward filter $\tilde{\boldsymbol{F}}\,(z)$ and a feedback filter $\tilde{\boldsymbol{B}}\,(z) - \boldsymbol{I}_4$, where $\tilde{\boldsymbol{B}}\,(z)$ is a causal and monic matrix polynomial, $\tilde{\boldsymbol{B}}\,(z) = \sum_{\nu=0}^{\tilde{q}_b} \tilde{\boldsymbol{B}}[\nu]\,z^{-\nu}$, $\tilde{\boldsymbol{B}}[0] = \boldsymbol{I}_4$. For the special case $\tilde{\boldsymbol{B}}\,(z) = \boldsymbol{I}_4$ a widely linear equalizer according to Section 3 results.

For calculation of the optimum infinite–length MMSE–DFE filters, standard results from literature on the MIMO–DFE may be used. Alternatively, an FIR MMSE–DFE receiver with WL processing may be employed, whose filters can be derived in a straightforward way.

On the other hand, the direct approach for combination of DFE and block decoding according to [5] has to use a MIMO DFE of dimension $N_T \times N_R$, which only yields high performance for $N_R \geq N_T$. Otherwise, a ZF–DFE does not exist.

5. Maximum–Likelihood Sequence Estimation

For transmission with binary modulation over channels with reasonably low order, maximum–likelihood sequence estimation (MLSE) may be realizable at the receiver. We again consider the polyphase components of the received sequences of all antennas. Because optimum estimation is now applied instead of linear processing, a receiver extension in the sense of WL processing does not yield any performance gain here, and the complex conjugates of the received signals are not considered for MLSE. Therefore, for receive antenna m, a vector

$$\bar{\boldsymbol{r}}_m[\mu] \stackrel{\triangle}{=} [\, r_m[2\,\mu] \quad r_m[2\,\mu+1]\,]^T\,, \qquad (12)$$

$m \in \{1,\, \dots,\, N_R\}$ is now defined, which satisfies

$$\bar{\boldsymbol{r}}_m[\mu] = \sum_{\nu=0}^{\tilde{q}_h} \bar{\boldsymbol{H}}_m[\nu]\,\tilde{\boldsymbol{c}}[\mu-\nu] + \bar{\boldsymbol{n}}_m[\mu], \qquad (13)$$

where $\bar{\boldsymbol{n}}_m[\mu] \stackrel{\triangle}{=} [\, n_m[2\,\mu] \quad n_m[2\,\mu+1]\,]^T$, and $\tilde{\boldsymbol{c}}[\mu]$ is given by Eq. (9). The matrix $\bar{\boldsymbol{H}}_m[\nu]$ can be obtained from $\tilde{\boldsymbol{H}}_m[\nu]$ (Eq. (8)) by deleting the 3rd and 4th row. Finally, we obtain

$$\bar{\boldsymbol{r}}[\mu] \stackrel{\triangle}{=} [\, \bar{\boldsymbol{r}}_1^T[\mu] \quad \dots \quad \bar{\boldsymbol{r}}_{N_R}^T[\mu]\,]^T = \sum_{\nu=0}^{\tilde{q}_h} \bar{\boldsymbol{H}}[\nu]\,\tilde{\boldsymbol{c}}[\mu-\nu] + \bar{\boldsymbol{n}}[\nu], \qquad (14)$$

where the definition of $\bar{\boldsymbol{H}}[\nu]$ and $\bar{\boldsymbol{n}}[\mu]$ is obvious. With this signal model, the ML metric for combined equalization and space–time block decoding of a burst of length L reads

$$\bar{\Lambda}(c_1'[\cdot], c_2'[\cdot]) \stackrel{\triangle}{=} \sum_{\mu=0}^{L/2-1} \|\,\bar{\boldsymbol{r}}[\mu] - \sum_{\nu=0}^{\tilde{q}_h} \bar{\boldsymbol{H}}[\nu]\,\tilde{\boldsymbol{c}}'[\mu-\nu]\,\|^2, \qquad (15)$$

$$\tilde{\boldsymbol{c}}'[\mu] \stackrel{\triangle}{=} [\, c_1'[\mu] \quad c_2'[\mu] \quad {c_1'}^*[\mu] \quad {c_2'}^*[\mu]\,]^T\,, \qquad (16)$$

which has to be minimized with respect to the equalizer trial sequences $c_1'[\cdot]$, $c_2'[\cdot]$. This minimization can be done recursively using the vector Viterbi algorithm, where the underlying trellis diagram has $Z = M^{2\,\tilde{q}_h}$ states in each time interval, and M^2 branches emerge from and merge into each state. Because of $\tilde{q}_h = \lceil \frac{q_h}{2} \rceil$, $Z = M^{q_h}$ for even q_h, i.e., in this case MIMO and SISO Viterbi algorithm have the same number of states, and $Z = M^{q_h+1}$ for odd q_h. The total number of branch metric computations per burst for the MIMO MLSE is

$$\frac{L}{2}\,M^{2\,\lceil q_h/2 \rceil}\,M^2 = \begin{cases} \frac{L}{2}\,M^{q_h+2}, q_h \text{ even}, \\ \frac{L}{2}\,M^{q_h+3}, q_h \text{ odd}, \end{cases} \qquad (17)$$

whereas $L\,M^{q_h+1}$ computations are required for the SISO Viterbi algorithm. Thus, for binary transmission and even channel orders, the number of required branch metric computations is equal, and complexity of uncoded and space–time encoded transmission is roughly the same. On the other hand, for high–level modulation space–time coding increases complexity by a factor of $M/2$ (q_h even), if MLSE is employed in the receiver. Thus, for EDGE where 8PSK modulation is applied complexity of space–time encoded transmission is four times higher than for conventional transmission for a receiver with MLSE[2].

In [12], an Ungerböck–type MLSE has been derived for Alamouti's code, assuming matched filters for receiver input filtering.

6. Delayed Decision–Feedback Sequence Estimation with Widely Linear Prefiltering

Decision–feedback sequence estimation (DDFSE) closes the gap between DFE and MLSE. For DDFSE, a minimum–phase transfer function matrix is necessary in order to achieve high performance. Therefore, in general, a paraunitary prefilter matrix is required in front of DDFSE in order to transform the matrix channel into its minimum–phase equivalent. Again, we focus on the case $N_R = 1$. Here, with a 4×2 prefilter for the polyphase components $r_1[2\,\mu]$, $r_1[2\,\mu+1]$ ($r_1^*[2\,\mu+1]$),

[2] It should be noted that for EDGE it is even impossible to realize a conventional MLSE [11].

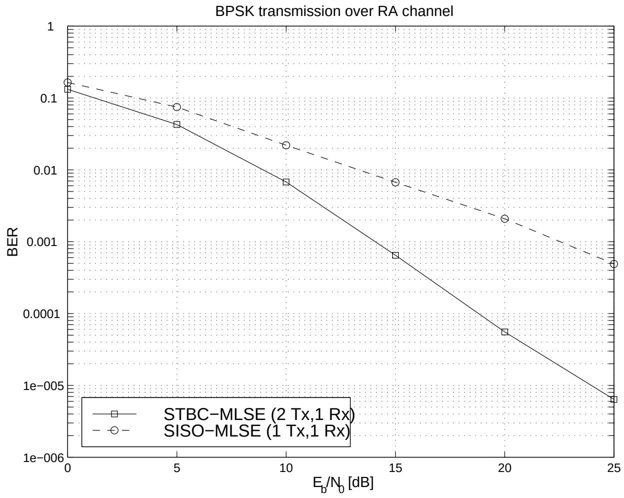

Figure 1: BER vs. $10\log_{10}(E_b/N_0)$ for BPSK transmission over RA channel and MLSE ($N_R = 1$). Tx: transmit antenna; Rx: receive antenna.

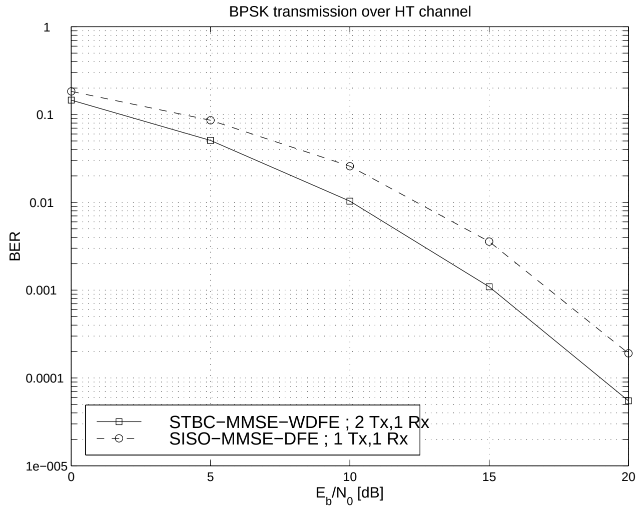

Figure 2: BER vs. $10\log_{10}(E_b/N_0)$ for BPSK transmission over HT channel and (W)DFE ($N_R = 1$).

an overall system suitable for DDFSE cannot be generated, because this prefilter has more outputs than inputs. Again, the solution to these difficulties is WL processing. We use the feedforward filter $\tilde{\boldsymbol{F}}(z)$ of a DFE with WL processing as a front–end filter for DDFSE. Considering also the complex conjugates of the received polyphase components, a 4×4 discrete–time channel arises, whose equivalent minimum–phase transfer function $\tilde{\boldsymbol{H}}_{\min}(z)$ is well defined. For an infinite–length ZF–DFE, $\tilde{\boldsymbol{F}}(z)\,\tilde{\boldsymbol{H}}(z) = \tilde{\boldsymbol{H}}_{\min}^{-1}[0]\,\tilde{\boldsymbol{H}}_{\min}(z)$ is valid. Hence, it can be also expected that the transfer function of the cascade of channel and a WL FIR MMSE–DFE feedforward filter is sufficiently close to the desired transfer function $\tilde{\boldsymbol{H}}_{\min}^{-1}[0]\,\tilde{\boldsymbol{H}}_{\min}(z)$. Assuming $\tilde{\boldsymbol{F}}(z)\,\tilde{\boldsymbol{H}}(z) = \tilde{\boldsymbol{B}}(z) = \sum_{\nu=0}^{\tilde{q}_b} \tilde{\boldsymbol{B}}[\nu]\,z^{-\nu}$, for trellis definition in DDFSE, the first $\tilde{q}_1 + 1$ taps of $\tilde{\boldsymbol{B}}(z)$ ($0 \leq \tilde{q}_1 \leq \tilde{q}_b$) are used, whereas the remaining taps are taken into account by state–dependent decision feedback. Hence, the number of states is reduced from $Z = M^{N_T\,\tilde{q}_h}$ (MLSE) to $Z = M^{N_T\,\tilde{q}_1}$.

The described DDFSE with prefiltering is also well suited for $N_R > 1$. Here, the feedforward filter again does not colour the noise, and $\tilde{\boldsymbol{B}}(z)$ is the minimum–phase solution of a generalized spectral factorization problem.

Finally, it should be noted that WL processing is also highly beneficial for channel memory shortening, which is another approach to reduced–complexity equalization.

7. Results for GSM/EDGE

In the following, the proposed receiver concepts are applied to an EGPRS (Enhanced General Packet Radio Service) transmission system, which is part of the EDGE standard. Nine modulation and coding schemes (MCS's) are specified for EGPRS, which use either 8PSK modulation or binary Gaussian minimum–shift keying (GMSK) modulation. The latter is also applied in GSM and can be well approximated by filtered BPSK modulation. Four different power delay profiles are specified: rural area (RA), hilly terrain (HT), typical urban area (TU), and equalizer test (EQ). For each transmit (Tx) and receive (Rx) branch, a GMSK pulse and a square–root raised cosine filter with roll–off factor $\alpha = 0.3$, respectively, has been selected, cf. also [11]. FIR equalizers of sufficient order are considered. Unless otherwise stated, only a single receive antenna has been used ($N_R = 1$).

First, we consider a binary transmission over the RA channel ($q_h = 2$). Fig. 1 shows the bit error rate (BER) versus $10\log_{10}(E_b/N_0)$ (E_b: average received bit energy per receive antenna, N_0: noise power spectral density) for a SISO transmission with MLSE in the receiver (SISO–MLSE) and a space–time encoded transmission with $N_T = 2$ (Alamouti's scheme) combined with MLSE according to Section 5 (STBC–MLSE). A gain of ≈ 8 dB can be observed at $\mathrm{BER} = 10^{-3}$, if space–time encoded transmission is applied instead of SISO transmission. This high gain is due to the small delay spread of the RA channel, which resembles a flat fading channel. On the other hand, equalization is necessary also for RA because transmit and receiver input filtering cause intersymbol interference, and the standard detector for space–time codes transmitted over a flat fading channel cannot be employed here. For TU, the gain of STBC–MLSE is reduced to ≈ 3dB at $\mathrm{BER} = 10^{-3}$. This is due to a higher inherent temporal diversity compared to RA because of a larger delay spread.

In Fig. 2, the performance of BPSK transmission over an HT channel is shown assuming a DFE in the receiver. STBC transmission with MMSE WL DFE (MMSE–WDFE) gains $\approx$2 dB at $\mathrm{BER} = 10^{-3}$ compared to SISO–MMSE–DFE.

In Fig. 3, we compare SISO–DDFSE ($Z = 4$) and STBC–DDFSE ($Z = 4$) for the TU profile. For SISO–DDFSE, the feedforward filter of a conventional FIR MMSE–DFE has been used in front of DDFSE in order to transform the chan-

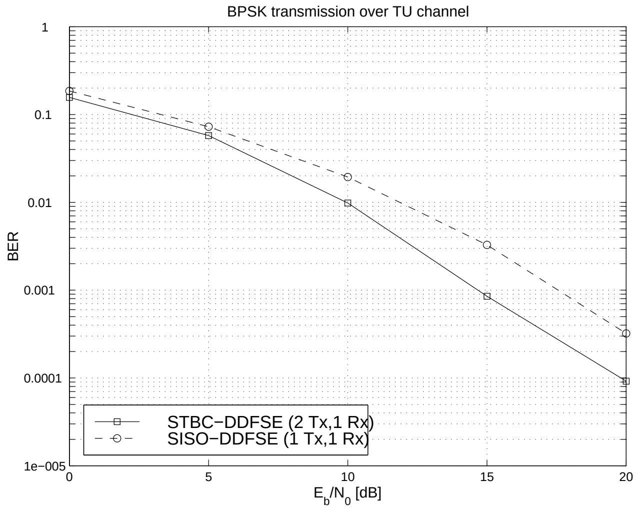

Figure 3: BER vs. $10\log_{10}(E_b/N_0)$ for BPSK transmission over TU channel and DDFSE ($Z = 4$) ($N_R = 1$).

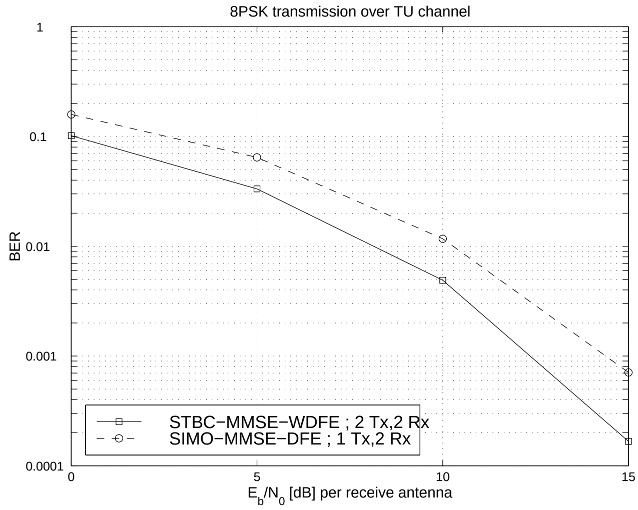

Figure 4: BER vs. $10\log_{10}(E_b/N_0)$ for 8PSK transmission over TU channel and (W)DFE (additional receive diversity, $N_R = 2$).

nel into its minimum–phase equivalent. For STBC–DDFSE, the feedforward filter of an FIR MMSE–WDFE has been selected. Fig. 3 shows that space–time encoded transmission is ≈ 2.5 dB better than SISO transmission at $\mathrm{BER} = 10^{-3}$. Finally, according to Fig. 4, which is valid for TU, 8PSK transmission, DFE, and $N_R = 2$, WL equalization using the structural properties of the STBC is also beneficial if additional receive diversity is available. A receiver using the direct combination of DFE and block decoding according to [5] requires $10\log_{10}(E_b/N_0) = 14.7$ dB for $\mathrm{BER} = 10^{-3}$ for this scenario, cf. Fig. 5 of [5]. Thus, WDFE is 2.3 dB better than the direct approach, and the latter is even slightly worse than SIMO–MMSE–DFE.

8. Conclusions

In this paper, we consider a space–time block–encoded transmission over frequency–selective fading channels. Utilizing the cyclostationarity and the rotational variance of the transmitted signal, we propose several widely linear detectors which also process the complex–conjugated versions of the received polyphase components. With this approach, high performance of detection can be also guaranteed for pure transmit diversity, in contrast to previously proposed schemes [5, 6]. Simulation results for EDGE indicate that space–time block–encoded transmission combined with a WL receiver outperforms a SISO transmission combined with the conventional counterpart of the corresponding WL receiver, if inherent temporal channel diversity is small to moderate.

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