PERFORMANCE OF SPACE-TIME-CODED OFDM WITH SECTORIZED ANTENNA RECEPTION

Peter Klenner, Karl-Dirk Kammeyer

University of Bremen, Department of Communications Engineering
28359 Bremen, Germany
{klenner,kammeyer}@ant.uni-bremen.de

ABSTRACT

Diversity is a well-known means to counter the detrimental effect of fading. Against this background, space-time codes have been designed to facilitate transmit diversity with a suitable receiver. Furthermore, space-time codes combined with OFDM allow for the utilization of frequency selective channels. However, a general assumption underlying the design of space-time codes is the time-invariance of the channel, which is violated in a high mobility environment. Large Doppler spread causing rapidly time-varying channels can be compensated at the receiver by sectorized antennas, which restrict the angle of incidence of imping waves to a finite range. Compared to an omnidirectional antenna, the resulting Doppler spread of a sector antenna and the channels time-variance are reduced. In this paper, we demonstrate the benefits of Doppler compensating sectorized antennas for space-time coded OFDM.

1. INTRODUCTION

Space-time (ST) codes are a tool to capitalize on transmit diversity by coding across space and time and can be categorized into several categories. Orthogonal ST block codes (OSTBC) with the Alamouti scheme [1] as its most prominent representative are specifically designed to allow for low-complex linear processing at the receiver which is identical to Maximum-Likelihood detection [2]. The fact, that for more than two transmit antennas OSTBCs with full rate and full diversity do not exist, lead to the design of quasi-orthogonal ST codes [3] with full rate at the expense of decreased diversity. Unlike OSTBCs and QOSTBs, the class of unitary or differential space-time-codes do not require any channel knowledge neither at the transmitter nor at the receiver. In [4] differential unitary space-time modulation (DUSTM) is proposed based on a group of unitary diagonal matrices, in [5] differential space-time modulation (DSTM) based on groups of unitary matrices is proposed. Both schemes can be interpreted as a matrix-valued generalization of the well-known differential modulation. Furthermore, for the special case of two transmit antennas, differential transmission can also be based on orthogonal designs [6].

ST codes are seamlessly combined with OFDM to overcome the influence of multipath propagation [7, 8]. However, due to the prolonged symbol duration, a ST coded OFDM system is likely to be more sensitive to channel fluctuations than a comparable single carrier system. Basically, a time-variant channel destroys the subcarriers orthogonality and introduces intercarrier interference [9]. Hence, interference between ST codewords arises requiring more complex equalization schemes other than linear processing [10].

Using sectorized receive antennas allows for compensation of large Doppler spread at the receiver [11]. The effect of directional antennas is such that impinging waves are separated according to their angle of incidence [12]. Thereby, the effective Doppler spread in each sector is reduced, and the resulting impulse response exhibits less time-variance compared to an omni-directional antenna which does not restrict the angle of incidence. Thus, under high mobility conditions sectorized antenna reception can render a time-variant channel quasi-static, such that the design criterion of ST codes is fulfilled.

In this paper, we consider a ST-coded OFDM system aided by sectorized antennas with either coherent or noncoherent reception. We restrict to the case of two transmit antennas. Coherent reception is based on the Alamouti scheme [1] and is facilitated by an equidistant pilot pattern. In principle, the benefit of rendering the channel time-invariant by sectorized reception lies in the increased bandwidth-efficiency due to a large pilot spacing in time direction. We investigate different types of equalizers to illustrate the influence of their respective trade-off between complexity and performance, i.e., Maximum Likelihood (ML), Zero-Forcing (ZF) and Matched Filtering (MF). On the other hand, differential ST schemes (DUSTM, DSTM) circumvent the problem of channel estimation and facilitate data detection without sacrificing bandwidth efficiency to pilot symbols. However, unlike differential modulation in a single carrier context, where the differentially modulated information is transmitted during two symbol durations, differential modulation for multiple transmit antennas is based on matrix-valued differential symbols. These require a much larger signaling interval, during which the transmitted signal is exposed to channel fluctuations. We demonstrate the benefits of sectorized antennas for this situation.

The paper is organized as follows. In Section 2 we detail our system model, i.e., transmitter, channel and receiver. In Section 3 we demonstrate the benefits of sectorized reception by simulations. Section 4 concludes this work.

2. SYSTEM MODEL

2.1. Transmitter

We consider a MIMO-OFDM system with N_T transmit antennas, N_R receive antennas, N subcarriers, and a cyclic prefix with N_g symbols. At time instant k the complex envelope of the t-th transmit antenna reads

$$s(k,t) = \sqrt{\frac{\mathcal{P}}{N_\mathrm{T} N}} \sum_{i=-\infty}^{\infty} \sum_{\nu=0}^{N-1} d(\nu,i,t) \mathrm{e}^{\mathrm{j}\frac{2\pi\nu(k-iZ)}{N}} g(k-iZ)\,, \tag{1}$$

where the transmit antennas are indexed by $0 \le t < N_\mathrm{T}$, the transmit power by $\mathcal{P}$, the transmitted symbol on subcarrier ν in the i-th OFDM symbol from the t-th antenna by $d(\nu,i,t)$, and $Z = N + N_\mathrm{g}$. The filter function $g(k)$ delimits the OFDM symbols in time, i.e., $g(k) = 1$ for $-N_\mathrm{g} \le k < N$ and $g(k) = 0$ for other time instants k.

Let us denote the information carrying symbols by $d_\mathrm{I}(k)$. They are drawn from an M-PSK or M-QAM signal constellation, and in case of one transmit antenna ($N_\mathrm{T} = 1$) they are directly mapped to the subcarriers $d(\nu,i,t)$ after multiplexing them with pilot symbols to enable channel estimation. Pilots are distributed in an equidistant pattern with pilot spacing in frequency direction denoted by Δ_f and in time direction denoted by Δ_t (cf. Fig 1a). In case of two transmit antennas ($N_\mathrm{T} = 2$) we resort to the Alamouti scheme

$$\begin{aligned} d(\nu,2i,0) &= d_\mathrm{I}(2\nu + iN)\,, \\ d(\nu,2i,1) &= d_\mathrm{I}(2\nu + iN + 1)\,, \\ d(\nu,2i+1,0) &= d_\mathrm{I}^*(2\nu + iN + 1)\,, \\ d(\nu,2i+1,1) &= -d_\mathrm{I}^*(2\nu + iN)\,. \end{aligned} \tag{2}$$

i.e., a ST codeword is transmitted on one subcarrier within two OFDM symbols. Thus, for the number of OFDM symbols to be even it becomes necessary to insert additional pilot symbols. An example is given in Figure 1b where the pilot spacing Δ_t = 2 requires the first two OFDM symbols to carry pilot symbols, whereas subsequently only every second OFDM symbol carries pilot symbols. Except for reference symbols, differential modulation can bypass channel estimation and the need for pilot symbols. It can be applied either in time or in frequency direction, depending on the channel constraints. It is advisable to choose the respective direction such that the correlation between consecutive subcarriers permits data recovery by differential demodulation. We choose differential modulation in time direction to render the system robust against severe frequency-selectivity for which

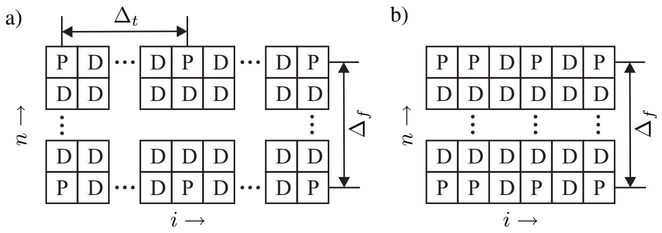

Fig. 1. Pilot schemes, OFDM symbol index i, subcarrier index n, pilot spacing in frequency Δ_f and time Δ_t; a) for $N_\mathrm{T} = 1$ transmit antennas the last OFDM symbol is required to contain pilot symbols; b) for $N_\mathrm{T} = 2$ transmit antennas an even number of OFDM symbols requires insertion of additional pilot symbols

the subcarrier correlation in frequency direction is low. Unfortunately, the system will be less tolerant to time-selectivity. However, Doppler compensation by sectorized antennas maintains the necessary correlation between subcarriers in neighboring OFDM, and aids the differential demodulator in recovering the transmitted data.

Differential modulation for $N_\mathrm{T} = 2$ is based on a group of M-ary orthogonal matrices [2, 13]

$$\mathcal{D} = \left\{ \frac{1}{\sqrt{2}} \begin{bmatrix} x & -y^* \\ y & x^* \end{bmatrix} \right\} \tag{3}$$

with $M = 2^{2p}$, $p \in \mathbb{N}$, and $x, y \in \exp(\mathrm{j}2\pi m/\sqrt{M})$, $0 \le m < \sqrt{M}$. Differential encoding is accomplished by computing

$$\boldsymbol{D}(k) = \boldsymbol{D}_\Delta(k)\boldsymbol{D}(k-1)\,, \tag{4}$$

where both the information carrying matrix symbol $\boldsymbol{D}_\Delta(k)$ and the differentially encoded matrix symbol $\boldsymbol{D}(k)$ are drawn from the group $\mathcal{D}$. Eventually, $\boldsymbol{D}$(k) is transmitted on one subcarrier within two OFDM symbols. In this paper, we fix $p = 2$, such that $M = 16$, i.e., the underlying signal constellation is QPSK, and one can transmit 4 bits per differential matrix symbol.

2.2. Channel

We are concerned with the performance of ST-coded OFDM under high mobility conditions received by either omnidirectional antennas or sectorized antennas. The former is characterized by the full Doppler spread and rapidly fading channels, whereas the latter reduces the Doppler spread and leads to less time-variant channel. Furthermore, omnidirectional reception is likely to benefit from time-diversity, however, it is susceptible to ICI, whereas this situation is reversed for sectorized antennas.

To assess the performance of such antenna structures we resort to a WSSUS (wide sense stationary uncorrelated scattering) channel, which assumes a richly scattering receive environment with impinging plane waves from uniformly distributed angles of incidence. The respective channel impulse response (CIR) at time instant k and delay ℓ reads [14]

$$h_{\mathrm{O}}(k,\ell)=\frac{1}{\sqrt{N_{\mathrm{e}}}}\sum_{\mu=0}^{N_{\mathrm{e}}-1}a(\mu)\mathrm{e}^{\mathrm{j}2\pi f(\mu)Tk}\delta(\ell-\ell(\mu))\,,\quad(5)$$

where $0\leq\ell<L$ with channel length L, sampling period T. We implicitly assumed that an realization of (5) models a CIR between the t-th transmit and r-th receive antenna. The N_{e} paths are characterized by their amplitude $a(\mu)$, Doppler frequency $f(\mu)$ and delay $\ell(\mu)$. The familiar Jakes' or bathtube-shaped Doppler spectrum follows, if the Doppler frequencies are modelled according to $f(\mu)=f_{\mathrm{D}}\cos((\mu))$, where the angles of incidence $\theta(\mu)$ are uniformly distributed between 0 and 2π. The maximum Doppler frequency is denoted as f_{D}. Eq. (5) models omnidirectional reception by superimposing paths from all directions leading to a maximally large Doppler spread.

On the other hand, a sectorizing antenna restricts the angles of incidence to a finite range. Thereby, the maximal Doppler spread can be reduced. We illustrated this approach in Fig. 2. As an example the horizontal plane is divided into 4 sectors, each corresponding to a sectorized antenna, e.g. sector antenna $r=0$ allows angles of incidence smaller than 70.5 and larger than 289.5. In Fig. 2b the corresponding subspectrum comprises Doppler frequencies from $1/3f_{\mathrm{D}}$ to f_{D}, i.e, the effective Doppler spread amounts now to $2/3f_{\mathrm{D}}$ instead of $2f_{\mathrm{D}}$. In general, the Doppler spread can be reduced by a factor $N_{\mathrm{R}}/2+1$, if the aperture angles are chosen such that the Doppler spectrum is divided into equisized subspectra. Eventually, we model the impulse response at the r-th sector antenna by

$$h_{\mathrm{S}}(k,\ell)=\frac{1}{\sqrt{N_{\mathrm{e}}}}\sum_{\mu=0}^{N_{\mathrm{e}}-1}a(\mu)\mathrm{e}^{\mathrm{j}2\pi f(r,\mu)Tk}\delta(\ell-\ell(\mu))\,,\quad(6)$$

where the Doppler frequencies follow from

$$f(r,\mu)=f_{\mathrm{D}}\cos(\theta(r,\mu))\,.\quad(7)$$

Due to the sectorization, the angles of incidence $\theta(r,\mu)$ are now restricted to the aperture angle of the r-th sector. We implicity assumed that (6) models the CIR between the t-th transmit and r-th receive antenna. A careful look at Fig. 2b reveals that the resulting subspectra have a reduced Doppler spread, but additionally a Doppler shift is associated with them, which needs to be corrected prior to the FFT. Omitting this frequency compensation would result into leakage between the subcarriers.

2.3. Receiver

The receive signal via omnidirectional reception reads

$$y_{\mathrm{O}}(k,r)=\sum_{t=0}^{N_{\mathrm{T}}-1}\sum_{\ell=0}^{L-1}h_{\mathrm{O}}(k,\ell,t,r)s(k-\ell,t)+\eta_{\mathrm{O}}(k,r)\,,\quad(8)$$

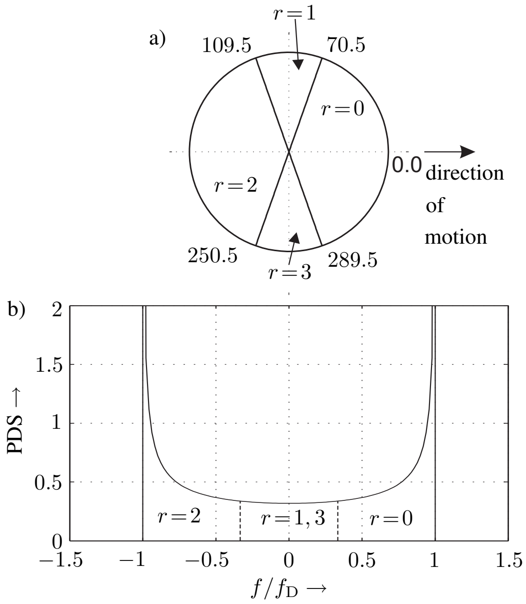

Fig. 2. a) Horizontal plane of reception divided into $N_{\mathrm{R}}=4$ sectors, b) Doppler spectrum divided into $N_{\mathrm{R}}/2+1$ subspectra with reduced Doppler spread

where the AWGN term $\eta_{\mathrm{O}}(k,r)$ of unit variance has been added. Upon guard removal the FFT is performed yielding

$$z_{\mathrm{O}}(n,i,r)=\sqrt{\frac{N_{\mathrm{T}}}{N\mathcal{P}}}\sum_{k=0}^{N-1}y_{\mathrm{O}}(k+iZ)\mathrm{e}^{-\mathrm{j}2\pi kn/N}\,.\quad(9)$$

The received signal via sectorizing antennas reads

$$y_{\mathrm{S}}(k,r)=\sum_{t=0}^{N_{\mathrm{T}}-1}\sum_{\ell=0}^{L-1}h_{\mathrm{S}}(k,\ell,t,r)s(k-\ell,t)+\eta_{\mathrm{S}}(k,r)\,.\quad(10)$$

Unlike omnidirectional reception we are required to perform frequency compensation prior to the FFT in order to avoid leakage. Let us denote the necessary frequencies by $f_{\mathrm{C}}(r)$, which in [15] are shown to correspond approximately to the average frequencies of the Doppler subspectra. Frequency compensation results from

$$\tilde{y}_{\mathrm{S}}(k,r)=\mathrm{e}^{-\mathrm{j}2\pi f_{\mathrm{C}}(r)Tk}y_{\mathrm{S}}(k,r)\,,\quad(11)$$

followed by the FFT

$$z_{\mathrm{S}}(n,i,r)=\sqrt{\frac{N_{\mathrm{T}}}{N\mathcal{P}}}\sum_{k=0}^{N-1}\tilde{y}_{\mathrm{S}}(k+iZ)\mathrm{e}^{-\mathrm{j}2\pi kn/N}\,.\quad(12)$$

In the following, we substitute $z(n,i,r)$ as a placeholder for either $z_{\mathrm{O}}(n,i,r)$ or $z_{\mathrm{S}}(n,i,r)$, which is valid since the subsequently described receiver approachs apply to both antenna structures. In general, we find that the n- th subcarrier in the i-th OFDM symbol at the r-th receive antenna can be separated into the data carrying subcarrier superimposed by AWGN and ICI [9]. The channel coefficient is denoted as $H_\mathrm{e}(\cdot)$ and the equivalent noise term by $\eta_\mathrm{e}(\cdot)$. The latter captures both ICI and AWGN

$$z(n,i,r) = \sum_{t=0}^{N_\mathrm{T}-1} H_\mathrm{e}(n,i,t,r)d(n,i,t,r) + \eta_\mathrm{e}(n,i,r)\,. \tag{13}$$

2.4. Channel estimation for coherent reception

We resort to Least Squares channel estimation for $N_\mathrm{T} = 1, 2$ antennas. For a pilot spacing in time direction $\Delta_t > 1$ some OFDM symbols are not carrying any pilot symbols. To estimate the corresponding channel transfer functions, we apply linear interpolation. In a first step, the CIRs are estimated, secondly the FFT of these CIRs deliver an estimate of the transfer functions, in a third step the remaining transfer functions are found by linear interpolation. For $N_\mathrm{T} = 1$, omitting the OFDM symbol index i, we collect the received pilot symbol in the vector

$$\boldsymbol{z}_P(r) = [z(0,r),\ z(\Delta_f,r),\ z(2\Delta_f,r),\cdots]^\mathrm{T}\,. \tag{14}$$

Defining $\mathbf{F}$ as the $(N\times N)$-DFT matrix and $\mathbf{F}_P = \mathbf{F}(0 : \Delta_f : N-1, 0 : L-1)$ we find an estimate of the CIR by computing

$$\hat{\boldsymbol{h}}(r) = \mathbf{F}_P^+ \boldsymbol{z}_P(r)\,. \tag{15}$$

Similarly, for $N_\mathrm{T} = 2$ we define the matrix

$$\boldsymbol{\Phi}(i) = [\mathrm{Dg}\{\mathbf{d}_P(i,0)\}\mathbf{F}_P]^\mathrm{T},\ \mathrm{Dg}\{\mathbf{d}_P(i,1)\}\mathbf{F}_P]^\mathrm{T} \tag{16}$$

where we have made use of the operator $\mathrm{Dg}\{\cdot\}$ which places a vector as its argument on a diagonal matrix. Aditionally introduced were the vectors $\mathbf{d}_P(i,0)$ and $\mathbf{d}_P(i,1)$ which denote the transmitted pilot symbols on Δ_f -spaced subcarriers. We estimate the CIRs by computing

$$[\hat{\boldsymbol{h}}^\mathrm{T}(0,r), \hat{\boldsymbol{h}}^\mathrm{T}(1,r)]^\mathrm{T} = \boldsymbol{\Phi}^+(i)\boldsymbol{z}_P(r)\,. \tag{17}$$

The channel transfer functions follow from (15) and (17) by applying the FFT to the respective CIRs.

2.5. Coherent detectors

In case of $N_\mathrm{T} = 1$, we perform Maximum Ratio Combining (MRC) before demodulation, in case of $N_\mathrm{T} = 2$, we consider three different detectors to illustrate the trade-off between complexity and performance of the respective devices. We are constructing equations to describe the received Alamouti ST Codewords in terms of the transmitted data and to arrive at suitable detection rules to retrieve this data. To this end, omitting the subcarrier index n we define the vectors

$$\begin{aligned}
\boldsymbol{z}_D(i,r) &= [z(2i,r),\ z^*(2i+1,r)]^\mathrm{T} && (18)\\
\boldsymbol{H}(i,r) &= \begin{bmatrix} H_\mathrm{e}(2i,0,r) & H_\mathrm{e}(2i,1,r) \\ -H_\mathrm{e}^*(2i+1,1,r) & H_\mathrm{e}^*(2i+1,0,r)\end{bmatrix} && (19)\\
\boldsymbol{d}_D(i) &= [d_\mathrm{I}(2i), d_\mathrm{I}(2i+1)]^\mathrm{T} && (20)\\
\boldsymbol{\eta}(i,r) &= [\eta_\mathrm{e}(2i,r),\ \eta_\mathrm{e}^*(2i+1,r)]^\mathrm{T} && (21)
\end{aligned}$$

to arrive at the system equation for an Alamouti codeword at the r-th receive antenna

$$\boldsymbol{z}_D(i,r) = \boldsymbol{H}(i,r)\boldsymbol{d}_D(i) + \boldsymbol{\eta}(i,r)\,. \tag{22}$$

Stacking all antenna signals yields

$$\begin{aligned}
\boldsymbol{z}_D(i) &= [\boldsymbol{z}_D(i,0),\cdots,\boldsymbol{z}_D(i,N_\mathrm{R}-1)]^\mathrm{T}\,, && (23)\\
\boldsymbol{H}(i) &= [\boldsymbol{H}^\mathrm{T}(i,0),\cdots,\boldsymbol{H}^\mathrm{T}(i,N_\mathrm{R}-1)]^\mathrm{T}\,, && (24)\\
\boldsymbol{\eta}(i) &= [\boldsymbol{\eta}^\mathrm{T}(i,0),\cdots,\boldsymbol{\eta}^\mathrm{T}(i,N_\mathrm{R}-1)]^\mathrm{T}\,. && (25)
\end{aligned}$$

We are eventually arriving at an equation which allows the application of the different detectors

$$\boldsymbol{z}_D(i) = \boldsymbol{H}(i)\boldsymbol{d}_D(i) + \boldsymbol{\eta}(i)\,. \tag{26}$$

Maximum-Likelihood, ML: This detector performs an exhaustive search over all hypotheses, selecting the candidate with minimum Euclidean distance to the receive signal

$$\hat{\boldsymbol{d}}_{D,ML}(i) = \underset{\tilde{\boldsymbol{d}}_D(i)}{\mathrm{argmin}}\, \|\boldsymbol{z}_D(i) - \boldsymbol{H}(i)\tilde{\boldsymbol{d}}_D(i)\|^2\,. \tag{27}$$

Zero-Forcing, ZF: This detector multiplies the received signal with the pseudo-inverse of the estimated channel matrix.

$$\hat{\boldsymbol{d}}_{D,ZF}(i) = \mathcal{Q}\{\boldsymbol{H}^+(i)\boldsymbol{z}_D(i)\,.\} \tag{28}$$

Matched-Filtering, MF: This detector multiplies the received signal with the Hermitian of the estimated channel matrix. If the channel is time-invariant over a ST-codeword, both symbols are perfectly decoupled, and MF- is identical to ML-detection. In case of time-variant conditions, MF suffers from interfering codesymbols.

$$[\hat{\boldsymbol{d}}_{D,ZF}(i)]_m = \mathcal{Q}\left\{\frac{[\boldsymbol{H}(i)]_m^\mathrm{H}\boldsymbol{z}_D(i)}{\|\boldsymbol{H}(i)\|^2}\right\},\ m=0,1 \tag{29}$$

The operator $\mathcal{Q}\{\cdot\}$ decides for the symbol in the applied signal constellation with smallest Euclidean distance to the passed argument.

2.6. Differential demodulation

In case of $N_\mathrm{T} = 1$, differential demodulation is simply achieved by

$$\hat{d}(n,i) = \mathcal{Q}\left\{\sum_{r=0}^{N_\mathrm{R}-1} z(n,i,r)z^*(n,i-1,r)\right\}. \tag{30}$$

Note, that the sum can be interpreted as Maximum Ratio Combining. For $N_\mathrm{T} = 2$, we follow the approach of [4]. Omitting subcarrier index n, the detector reads

$$\hat{\boldsymbol{D}}(i) = \underset{\tilde{\boldsymbol{D}}(i)\in\mathcal{D}}{\mathrm{argmax}} \left\|\boldsymbol{Z}(i) + \tilde{\boldsymbol{D}}(i)\boldsymbol{Z}(i-1)\right\|^2, \tag{31}$$

where we defined

$$\boldsymbol{Z}(i) = \begin{bmatrix} z(2i,0) & z(2i,1) & \cdots & z(2i,N_\mathrm{R}-1) \\ z(2i{+}1,0) & z(2i{+}1,1) & \cdots & z(2i{+}1,N_\mathrm{R}{-}1)\end{bmatrix}. \tag{32}$$

3. SIMULATION RESULTS

To illustrate the performances of transmitting with $N_{\mathrm{T}} = 1, 2$ transmit antennas and either omnidirectional or sectorized reception we performed several simulations based on the channel model described in Section 2.2. The power delay profile of the channel was chosen to be exponential such that each channel taps power decays by 3 dB compared to its predecessor. The number of subcarriers is fixed to $N = 64$, the cyclic prefix to $N_{\mathrm{g}} = 16$. Let γ^2 be the ratio between the number of totally transmitted symbols and the number of information carrying symbols, and define $E_{\mathrm{b}}/N_0 = \mathcal{P}\gamma^2(N + N_{\mathrm{g}})/(NN_{\mathrm{R}} log_2 M R_{\mathrm{c}})$ as the ratio of total bit energy to noise power density at the receiver.

In Figure 3 the bit error rate of single antenna transmission is depicted over the normalized Doppler frequency $f_{\mathrm{D}}/T_{\mathrm{s}}$, thereby T_{s} corresponds to the OFDM symbol duration and, likewise, to the inverse of the subcarrier spacing. Figures 3a-c illustrate the performance of omnidirectional reception for different pilot spacings in time $\Delta_t = 1, 2, 3$. One recognizes the general trend that larger Doppler frequencies lead to an increased number of bit errors, since the influence of intercarrier-interference increases as well. Furthermore, for the given system the sampling theorem demands that the normalized Doppler frequency adheres to $f_{\mathrm{D}}T_{\mathrm{s}} < 0.4/\Delta_t$ for omnidirectional reception and to $f_{\mathrm{D}}T_{\mathrm{s}} < 0.4(N_{\mathrm{R}}/2 + 1)/\Delta_t$ for sectorized reception. Figures 3d-f reveal that sectorized reception enlarges the range of Doppler frequencies, for which reliable detection is possible.

In Figure 4 performance results are given for $N_{\mathrm{T}} = 2$ transmit antennas with either omnidirectional or sectorized reception. The normalized Doppler frequency is fixed to the value $f_{\mathrm{D}}T_{\mathrm{s}} = 0.2$ corresponding to a rapidly time-variant channel. The three different detectors from Section 2.5 are applied, and the pilot spacings in time $\Delta_t = 1, 3$ are used.

Omnidirectional reception leads in all considered cases into an error-floor, due to the uncompensated Doppler influence and the therefore considerably large intercarrier-interference. Sectorized reception, on the other hand, successfully compensates the large Doppler influence and performs significantly better at high SNR compared to omnidirectional reception. Furthermore, the performance of the ML and ZF detectors are very similar, with a slightly better performance of the former. As it turns out, the low-complex Matched Filtering (MF) approach, which completely disregards the time-variance of the channel performs closely to ML and ZF in the case of $N_{\mathrm{R}} = 8$ receive antennas. For $N_{\mathrm{R}} = 4$, one can recognize an SNR-loss of approx. $3 - 4$ dB to ML and ZF.

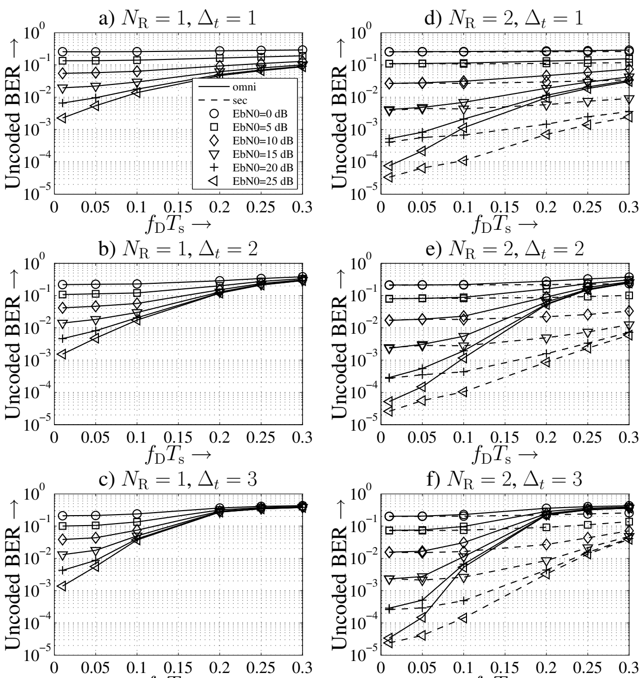

Fig. 3. Uncoded BER vs. normalized Doppler frequency $f_{\mathrm{D}}T_{\mathrm{s}}$ for single antenna transmission with either omnidirectional or sectorized reception, $N_{\mathrm{R}} = 1, 2$ receive antennas, different pilot spacings $\Delta_f = 3$, $\Delta_t = 1, 2, 3$, exponential power delay profile, QPSK

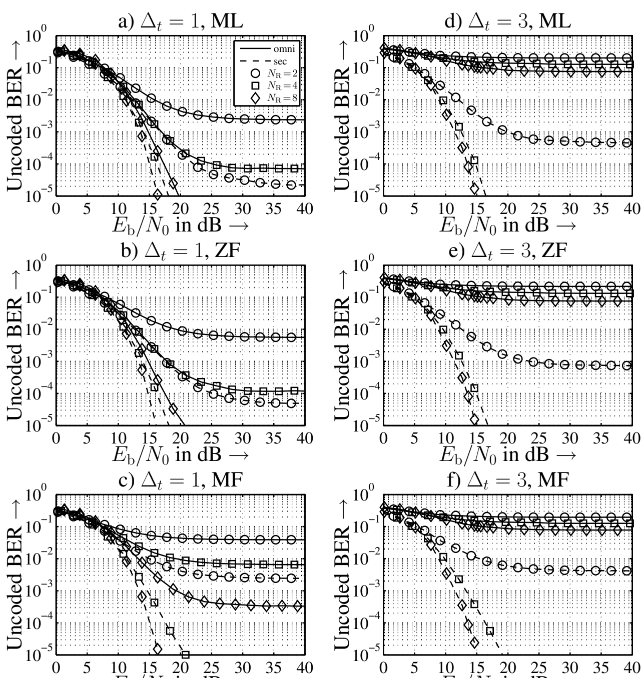

Fig. 4. Uncoded BER vs. E_{b}/N_0, normalized Doppler frequency $f_{\mathrm{D}}/T_{\mathrm{s}}$ for two transmit antennas with either omnidirectional or sectorized reception, $N_{\mathrm{R}} = 2, 4, 8$ receive antennas, different pilot spacings $\Delta_f = 3$, $\Delta_t = 1, 3$, exponential power delay profile, QPSK

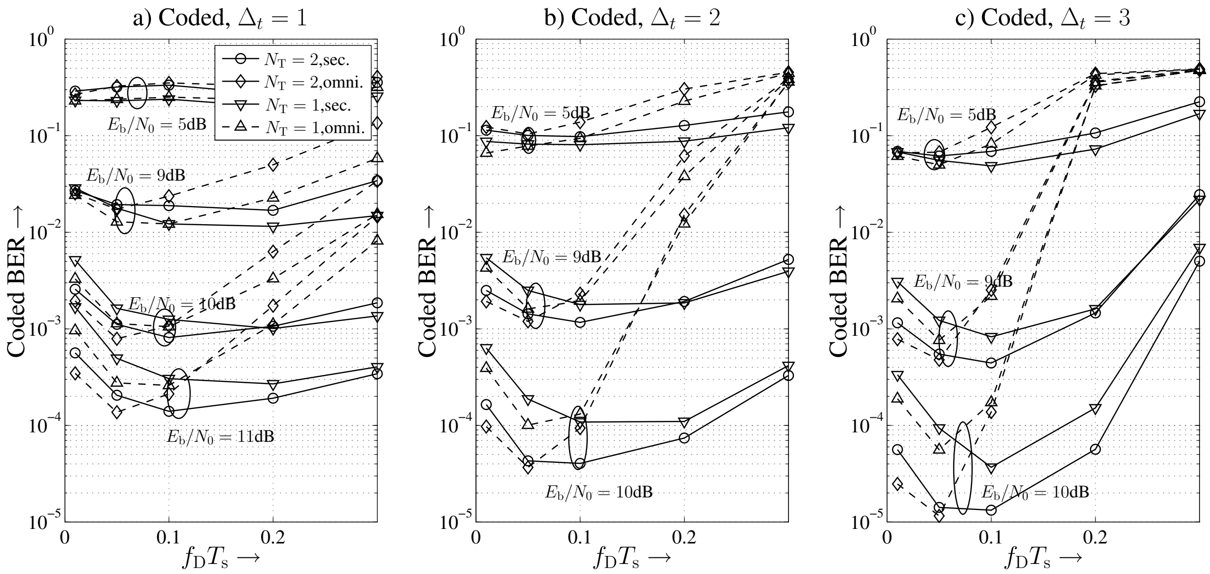

Fig. 5. Coherent modulation with $N_{\mathrm{T}} = 1, 2$ transmit antennas and either sectorized or omnidirectional reception with $N_{\mathrm{R}} = 2$, MRC for $N_{\mathrm{T}} = 1$, ML detection for $N_{\mathrm{T}} = 2$, QPSK, convolutional code $[7, 5]_8$, Least Squares channel estimation

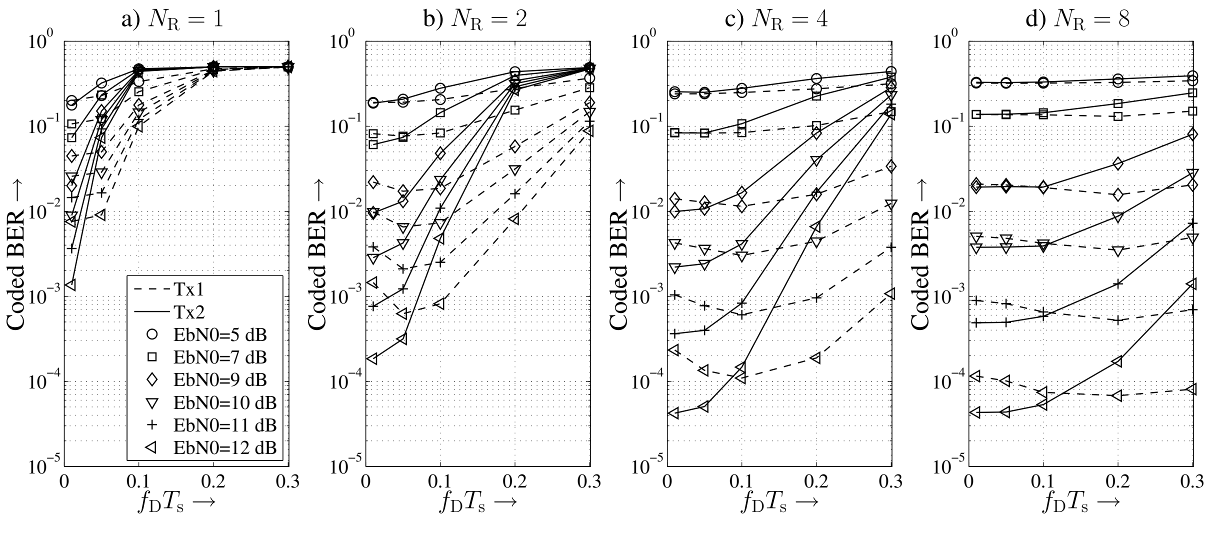

Fig. 6. Differential modulation with $N_{\mathrm{T}} = 1, 2$ transmit antennas and sectorized reception with $N_{\mathrm{R}} = 1, 2, 4, 8$

In Fig. 5 we present results for a convolutionally coded and randomly bit-interleaved ST-coded OFDM system with single and two-fold transmit diversity based on the Alamouti scheme. The pilot spacing in time is varied within $\Delta_t = 1, 2, 3$. For $N_\mathrm{T} = 1$ we perform Maximum Ratio Combining, for $N_\mathrm{T} = 2$ we perform ML detection.
We observe that for small Doppler frequencies sectorized reception with a single transmit antenna performs best. This is due to the negligible intercarrier-interference. Furthermore, omnidirectional reception leads here to an increased time diversity, which can be exploited by the underlying channel code. For larger Doppler frequencies ($f_\mathrm{D}T_\mathrm{s} > 0.05$), sectorized reception in conjunction with two-fold transmit diversity performs best. However, in this Doppler range a single transmit antenna combined with sectorized reception performs quite closely to the two-fold transmit diversity case. Additionally, omnidirection reception for large Doppler frequencies leads to a rather harsh impairment of the bit error rate since intercarrier interference is not compensated.
Figure 6 depicts simulation results for differential modulation with $N_\mathrm{T} = 1, 2$ transmit antennas and sectorized reception with $N_\mathrm{R} = 1, 2, 4, 8$. For both single and double transmit diversity the coded BER benefits from a finer sectorization at the receiver, since the channel is rendered the less time-variant, the more sectors are provided. It is also apparent from Figure 6 that two-fold transmit diversity performs superior over single transmit diversity only for small Doppler frequencies. This is due to the prolonged signalling interval of the matrix-valued differential symbols. In the given case it takes the duration of four OFDM symbols to transmit one matrix-valued differential symbol.

4. CONCLUSION

We have described a space-time coded OFDM system in a high mobility environment. To compensate for the rapidly time-variant channelwe employed sectorizing antennas, which lead to a separation of the Doppler spectrum into a set of subspectra. Those are characterized by a smaller Doppler spread, thus they are associated with a slowly time-varying channel. We investigated both coherent and noncoherent (differential) ST-codes, both based on the Alamouti scheme. For coherent reception we examined three different detectors, Maximum-Likelihood dection, Zero-Forcing and Matched-Filtering.
Whereas the former two perform quite similary due to their high complexity, the latter Matched-Filtering as a low-complexity approach was shown to benefit greatly from sectorized reception. This is due to rendering the channel quasi-static if an appropriate number of sector antennas is provided. Differential ST-coded were shown to benefit from sectorized reception only for the low Doppler case. For the high Doppler case, the signaling interval becomes too large, since matrix valued differential symbols need to be transmitted. Hence, single antenna transmission leads to symbols of shorter duration, which improves the performance under high Doppler influence.

5. REFERENCES

[1] S. M. Alamouti, "A Simple Transmit Diversity Technique for Wireless Communications," *IEEE Journal Selected Areas in Comm.*, vol. 16, pp. 1451–1458, Oct. 1998.

[2] H. Jafarkhani V. Tarokh and A. R. Calderbank, "Space-Time Block Codes from Orthogonal Designs," *IEEE Trans. Info. Theory*, vol. 45, pp. 1456–1467, July 1999.

[3] H. Jafarkhani, "A Quasi-Orthogonal Space-Time Block Code," *IEEE Trans. Comm.*, vol. 49, pp. 1–4, Jan. 2001.

[4] B. M. Hochwald and W. Sweldens, "Differential Unitary Space-Time Modulation," *IEEE Trans. Comm.*, vol. 49, pp. 2041–2052, March 1999.

[5] B. L. Hughes, "Differential Space-Time Modulation," *Trans. Info. Theory*, vol. 46, pp. 2567–2578, Nov. 2000.

[6] V. Tarokh and H. Jafarkhani, "A Differential Detection Scheme for Transmit Diversity," *IEEE Journal Selected Areas in Comm.*, vol. 18, pp. 1169–1174, July 2000.

[7] A. J. Paulraj H. Bölcskei, "Space-Frequency Coded Broadband OFDM Systems," in *IEEE Proc. WCNC*, Sept. 2000, pp. 1–6.

[8] G. J. Stüber, J. R. Barry, S. W. Mclaughlin, Y. Li, M. A. Ingram, and T. G. Pratt, "Broadband MIMO-OFDM Wireless Communications," *Proc. IEEE*, vol. 92, pp. 271–294, Feb. 2004.

[9] A. Stamoulis, S. N. Diggavi, and N. Al-Dhahir, "Intercarrier Interference in MIMO OFDM," *IEEE Trans. Signal Processing*, vol. 50, pp. 2451–2464, Oct. 2002.

[10] T. A. Tran and A. B. Sesay, "A Generalized Simplified ML Decoder of Orthogonal Space-Time Block Code for Wireless Communications over Time Selective Fading Channels," *IEEE Proc. VTC*, pp. 1911–1915, 2002.

[11] P. Klenner, S. Vogeler, and K.-D. Kammeyer, "Doppler-Compensation for OFDM-Transmission by Sectorized Antenna Reception," in *6th Intern. Workshop on Multi-Carrier Spread Spectrum, MCSS 2007*, 2007, pp. 237–246.

[12] W. T. Ng and V. K. Dubey, "Effect of Employing Directional Antennas on Mobile OFDM System with Time-Varying Channel," *IEEE Communications Letters*, pp. 165–167, April 2003.

[13] A. Shokrollahi, B. Hassibi, B.M. Hochwald, and W. Sweldens, "Representation Theory for High-Rate Multiple-Antenna Code Design," *IEEE Trans. Info. Theory*, pp. 2335–2367, Sept. 2001.

[14] P. Höher, "A Statistical Discrete-Time Model for the WSSUS Multipath Channel," *IEEE Trans. Veh. Tech*, pp. 461–468, Nov. 1992.

[15] O. Norklit and R. G. Vaughan, "Angular Partitioning to Yield Equal Doppler Contributions," *IEEE Trans. Veh. Tech*, pp. 1437–1442, March 2005.