Modified Space-Time Transmission in DS-CDMA Downlink Facilitating MISO Channel Equalization

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***Abstract*— A downlink space-time block coding transmission scheme is assessed via Monte Carlo simulations in a direct sequence code division multiple access context. Raw symbol error rate performance of the space-time transmit diversity scheme defined in the UTRA/FDD standard serves as a benchmark. Exploiting space-time code properties, the new scheme facilitates radio channel equalization and space-time decoding prior to despreading. Assuming ideal equalization, the proposed diversity scheme is therefore own cell interference free and outperforms conventional, schemes in frequency selective fading environments. With increasing channel estimation error however the proposed scheme tends to suffer from multiple access interference.**

***Index Terms*— DS-CDMA, Transmit Diversity, STBC, STTD, MISO, MIMO, Equalization .**

I. INTRODUCTION

IN the downlink direction of a Direct Sequence Code Division Multiple Access System (DS-CDMA) like UTRA (Universal Terrestrial Radio Access) a Base Station (BS) transmits simultaneously signals for many different users in the same frequency band. Orthogonal spreading codes are applied to make signals separable in the code domain. Due to multipath propagation different time delayed versions of the transmitted signals arrive at a receiving Mobile Station (MS). Conventionally, a Rake receiver coherently combines the multipath signals. Nevertheless, since time misaligned spreading codes are usually nonorthogonal, Rake receiver performance is often interference limited. To circumvent this problem [1] and [2], among others, propose to equalize a Single Input Single Output (SISO) radio channel. Assuming perfect equalisation, orthogonality between the users' signals is restored, and the receiver can detect the wanted signal by a simple despreading operation.

In uncorrelated frequency flat fading channels, Space-Time Block Coding (STBC) as proposed in [3] can deliver a second order diversity gain. Therefore, a two transmit antenna STBC scheme called Space-Time Transmit Diversity (STTD) was included in the UTRA FDD (Frequency Division Duplex) standard [4]. In a frequency selective propagation environment however an STTD decoding Rake receiver is then confronted with Inter Path- and Multiple Access Interference (IPI and MAI) from two transmit antennas. The total interference power however remains constant, as the transmit power is usually split evenly between the two transmit antennas. Using at least two receive antennas, own cell interference can be avoided by equalizing the Multiple Input Multiple Output (MIMO) radio channel with a similar equalization approach as in [1] and [2]. Keeping the advantage of requiring only a single antenna receiver RF-chain, a joint STBC multi user detector is proposed in [5]. Multi user detection requires however that the receiver knows the other users' spreading codes, which is usually not the case in a downlink situation. Here we propose a modified STTD transmitter architecture (M-STTD), where space-time block encoding and spreading/scrambling operations are swapped. This enables a single antenna receiver to perform joint STBC decoding and Multiple Input Single Output (MISO) channel equalization using standard estimation techniques like Zero Forcing (ZF) or Minimum Mean Square Error (MMSE) estimation. Under ideal conditions IPI and MAI can then be avoided.

After a short introduction to the 2 to 1 antenna MISO propagation channel representation, section II introduces a matrix vector notation for conventional and modified STTD. The performance of both schemes in different propagation environments (section III) will then be compared via single cell, multi user, Monte Carlo simulations (section IV). A brief discussion and concluding remarks are given in section V and VI respectively.

II. MATRIX VECTOR REPRESENTATION OF STTD AND M-STTD

STTD and M-STTD are both based on STBC [3], where the space-time coding block length is chosen, so that it does not exceed the channel's coherence time. Furthermore, for simplicity it is assumed, that the number of symbols (N_{sym}) within every block is even, and all users' spreading codes are of the same length N_{sf}.

A. The MISO Radio Channel

The chip sampled channel impulse response from transmit antenna m to receive antenne n is donated by

$$\mathbf{h}_{n,m} = \begin{bmatrix} h_{n,m}^{(1)} \\ h_{n,m}^{(2)} \\ \vdots \\ h_{n,m}^{(N_h)} \end{bmatrix}, \tag{1}$$

where N_h is the channel delay spread (DS) measured in chips. Stacking time shifted replicas of the channel impulse response vector into a matrix

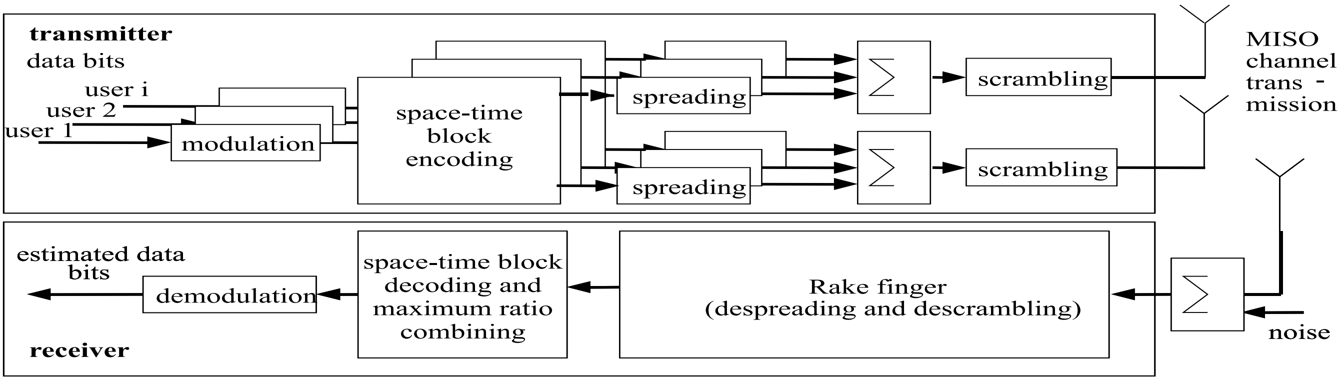

Fig. 1. Conventional STTD Transmit/Receive Chain

$$\mathbf{H}_{n,m} = \begin{bmatrix} \mathbf{h}_{n,m} & 0 & 0 & & 0 \\ \vdots & \mathbf{h}_{n,m} & 0 & & 0 \\ 0 & \vdots & \mathbf{h}_{n,m} & & 0 \\ 0 & 0 & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & & \mathbf{h}_{n,m} \end{bmatrix} \quad (2)$$

creates a convolution SISO channel matrix with the dimensions $(N_{sf} \cdot N_{sym} + N_h - 1) \times (N_{sf} \cdot N_{sym})$. The time shift between neighboring columns amounts to one chip period.

The 2 to 1 MISO channel can be described by concatenation of two SISO convolution channel matrices:

$$\mathbf{H}_{MISO} = \begin{bmatrix} \mathbf{H}_{1,1} & \mathbf{H}_{1,2} \end{bmatrix} \quad (3)$$

The underlying spatial correlated impulse responses $\mathbf{h}_{1,1}$ and $\mathbf{h}_{1,2}$ are obtained from the stochastic METRA MIMO radio channel model [6]. Detailed channel parameter settings are presented in section III.

B. Conventional STTD

Fig.1 displays a block diagram of a conventional STTD transmit/receive chain. The modulated and space-time encoded data symbols of all users are spread separately. Afterwards they are added and scrambled. The receiver's Rake fingers descramble and despread the time dispersed received signal prior to joint STTD-decoding and maximum ratio combining (MRC). The receive vector $\mathbf{r}$ can be written as

$$\begin{aligned} \mathbf{r} &= [\mathbf{H}_{1,1}, \mathbf{H}_{1,2}] \sum_{i=1}^{N_{user}} \begin{bmatrix} \mathbf{C}_i & \mathbf{0} \\ \mathbf{0} & \mathbf{C}_i \end{bmatrix} \cdot \begin{bmatrix} \mathbf{I} & \mathbf{0} \\ \mathbf{0} & \mathbf{T} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix} + \mathbf{n} \\ &= \sum_{i=1}^{Nuser} \begin{bmatrix} \mathbf{H}_{1,1}\mathbf{C}_i\mathbf{I} & \mathbf{H}_{1,2}\mathbf{C}_i\mathbf{T} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix} + \mathbf{n}\,, \end{aligned} \quad (4)$$

where $\mathbf{d}_i$ is a column vector of the i^{th} user's data symbols and $\mathbf{d}_i^*$ is its complex conjugate. The number of active users is N_{user}. $\mathbf{C}_i$ is a block diagonal matrix carrying complex, user- and symbol specific codes, constructed through an element by element multiplication of Welsh-Hadamard spreading codes and fractions of a Gold code scrambling sequence. The $\mathbf{0}$s represent zero padding matrices. The vector $\mathbf{n}$ contains complex additive white gaussian noise samples. An STTD encoding matrix is generated by combining identity matrix $\mathbf{I}$, zero padding matrices $\mathbf{0}$, and a permutation matrix $\mathbf{T}$. If, for example, STTD encoding is performed on a two-symbol block, $\mathbf{d}_i$, $\mathbf{I}$ and $\mathbf{T}$ are given by

$$\mathbf{d}_i = \begin{bmatrix} d_{i1} \\ d_{i2} \end{bmatrix}, \mathbf{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \mathbf{T} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}, \quad (5)$$

leading to the space-time encoded data sequence

$$\begin{aligned} \begin{bmatrix} d_{i1} \\ d_{i2} \\ -d_{i2}^* \\ d_{i1}^* \end{bmatrix} &= \begin{bmatrix} \mathbf{I} & \mathbf{0} \\ \mathbf{0} & \mathbf{T} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix} \\ &= \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \end{bmatrix} \cdot \begin{bmatrix} d_{i1} \\ d_{i2} \\ d_{i1}^* \\ d_{i2}^* \end{bmatrix}. \end{aligned} \quad (6)$$

Similar as in [5], the received signal vector $\mathbf{r}$ can be stacked with its complex conjugate to deliver

$$\begin{bmatrix} \mathbf{r} \\ \mathbf{r}^* \end{bmatrix} = \sum_{i=1}^{N_{user}} \underbrace{\begin{bmatrix} \mathbf{H}_{1,1}\mathbf{C}_i\mathbf{I} & \mathbf{H}_{1,2}\mathbf{C}_i\mathbf{T} \\ \mathbf{H}_{1,2}^*\mathbf{C}_i^*\mathbf{T}^* & \mathbf{H}_{1,1}^*\mathbf{C}_i^*\mathbf{I}^* \end{bmatrix}}_{\mathbf{A}_i} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix} + \begin{bmatrix} \mathbf{n} \\ \mathbf{n}^* \end{bmatrix}. \quad (7)$$

It can be seen that the channel components cannot be left extracted from the user specific transmission matrix $\mathbf{A}_i$. Channel equalisation can therefore not be separated from a despreading operation.

Ideal Rake reception and STTD decoding of user k's data symbols can be performed by

$$\begin{bmatrix} \widehat{\mathbf{d}}_{k,\mathrm{RAKE}} \\ \widehat{\mathbf{d}}_{k,\mathrm{RAKE}}^* \end{bmatrix} = \mathbf{A}_k^H \cdot \begin{bmatrix} \mathbf{r} \\ \mathbf{r}^* \end{bmatrix}, \quad (8)$$

where $(\cdot)^H$ and $\widehat{(\cdot)}$ indicates hermitian transpose and estimation respectively.

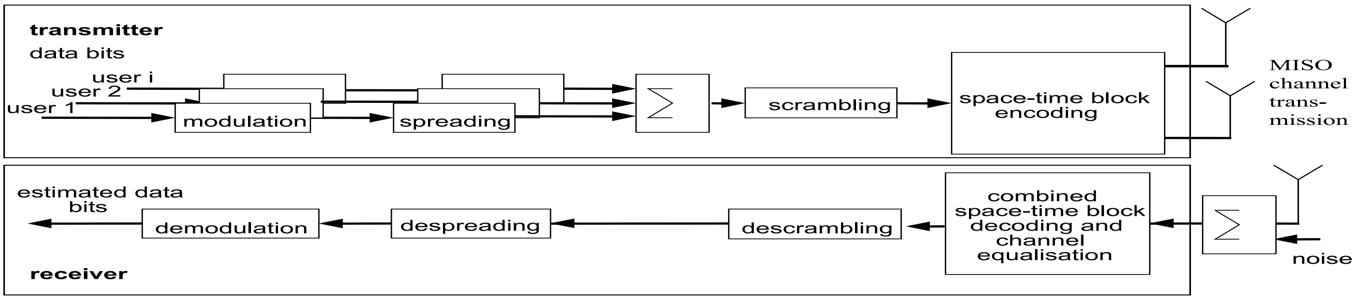

Fig. 2. Modified STTD Transmit Receive Chain

C. Modified STTD

Fig.2 shows the modified transmit/receive chain. Comparing Fig.1 and 2, it can be seen that on the transmit side STBC and spreading/scrambling operations are swapped. Using a similar mathematical notation as in (4) the receive vector can be written as

$$\widetilde{\mathbf{r}} = [\mathbf{H}_{1,1}, \mathbf{H}_{1,2}] \cdot \begin{bmatrix} \widetilde{\mathbf{I}} & \mathbf{0} \\ \mathbf{0} & \widetilde{\mathbf{T}} \end{bmatrix} \cdot \sum_{i=1}^{Nuser} \begin{bmatrix} \mathbf{C}_i & \mathbf{0} \\ \mathbf{0} & \mathbf{C}_i^* \end{bmatrix} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix} + \mathbf{n}. \tag{9}$$

$\tilde{\mathbf{I}}$ and $\tilde{\mathbf{T}}$ are now given by

$$\widetilde{\mathbf{I}} = \mathbf{I} \otimes \mathbf{I}' , \tag{10}$$

$$\widetilde{\mathbf{T}} = \mathbf{T} \otimes \mathbf{I}' , \tag{11}$$

where $\otimes$ represents the Kronecker Product [7] and $\mathbf{I}'$ is an identity matrix of dimensions $N_{sf} \times N_{sf}$. Substituting

$$\begin{bmatrix} \mathbf{S} \\ \mathbf{S}^* \end{bmatrix} = \sum_{i=1}^{N_{user}} \begin{bmatrix} \mathbf{C}_i & \mathbf{0} \\ \mathbf{0} & \mathbf{C}_i^* \end{bmatrix} \cdot \begin{bmatrix} \mathbf{d}_i \\ \mathbf{d}_i^* \end{bmatrix}, \tag{12}$$

it becomes clear that STTD encoding is now commonly performed on the spread and scrambled symbols of all users:

$$\begin{aligned} \widetilde{\mathbf{r}} &= [\mathbf{H}_{1,1}, \mathbf{H}_{1,2}] \cdot \begin{bmatrix} \widetilde{\mathbf{I}} & \mathbf{0} \\ \mathbf{0} & \widetilde{\mathbf{T}} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{S} \\ \mathbf{S}^* \end{bmatrix} + \mathbf{n} \\ &= \begin{bmatrix} \mathbf{H}_{1,1}\widetilde{\mathbf{I}} & \mathbf{H}_{1,2}\widetilde{\mathbf{T}} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{S} \\ \mathbf{S}^* \end{bmatrix} + \mathbf{n} \end{aligned} \tag{13}$$

A similar stacking operation as in (7) delivers:

$$\begin{bmatrix} \widetilde{\mathbf{r}} \\ \widetilde{\mathbf{r}}^* \end{bmatrix} = \underbrace{\begin{bmatrix} \mathbf{H}_{1,1}\widetilde{\mathbf{I}} & \mathbf{H}_{1,2}\widetilde{\mathbf{T}} \\ \mathbf{H}_{1,2}^*\widetilde{\mathbf{T}}^* & \mathbf{H}_{1,1}^*\widetilde{\mathbf{I}}^* \end{bmatrix}}_{\tilde{\mathbf{A}}} \cdot \begin{bmatrix} \mathbf{S} \\ \mathbf{S}^* \end{bmatrix} + \begin{bmatrix} \mathbf{n} \\ \mathbf{n}^* \end{bmatrix} \tag{14}$$

The transmission matrix $\tilde{\mathbf{A}}$ is now common to all users. Using MMSE decoding the spread and scrambled signal estimate of all users' symbols becomes

$$\begin{bmatrix} \hat{\mathbf{S}}_{MMSE} \\ \hat{\mathbf{S}}_{MMSE}^* \end{bmatrix} = (\tilde{\mathbf{A}}^H \cdot \tilde{\mathbf{A}} + \sigma^2 \cdot \mathbf{I}'')^{-\mathbf{1}} \cdot \tilde{\mathbf{A}}^{\mathbf{H}} \cdot \begin{bmatrix} \widetilde{\mathbf{r}} \\ \widetilde{\mathbf{r}}^* \end{bmatrix} , \tag{15}$$

where σ^2 is the noise variance and $\mathbf{I}''$ is an identity matrix with dimensions $(2 \cdot N_{sf} \cdot N_{sym}) \times (2 \cdot N_{sf} \cdot N_{sym})$. Similarly, the zero forcing estimate is given by setting σ^2 to zero. Descrambling and despreading can then be performed by

$$\hat{\mathbf{d}}_{k,MMSE} = \mathbf{C}_k^H \cdot \hat{\mathbf{S}}_{MMSE}, \tag{16}$$

to deliver the data estimate of user k.

III. Simulation Procedure

Single cell performance of both space-time coding schemes is assessed via Monte Carlo simulations, where the BS transmits equally powered signals to 10 MS each using a spreading factor of 16. The soft data estimates, obtained by solving (8) and (16), are hard delimited by mapping them to the closest modulation scheme constellation point, whereby Gray coding is used in connection with higher order modulation.

The MISO channel matrices are generated using the METRA MIMO channel simulator [6]. The total average channel power is normalized to one, i.e. large scale effects like path loss and shadow fading are not included in the model. Spatial correlation between the radio propagation links is introduced in form of correlation matrices, where the entries in the correlation matrices are derived from parameters like antenna element spacing, Power Azimuth Spectrum (PAS), mean Angle of Arrival (AoA) and Azimuth Spread (AS). Simulation results are obtained for different propagation scenarios specified in [8]. Table I gives an overview of the two propagation scenarios "ITU Generalised Pedestrian A" (Ped.A) and "ITU Generalised Vehicular A" (Vehic. A) used to generate the symbol error rate (SER) results in Section IV.

Initially, simulations are performed assuming ideal channel knowledge at the receiver, which means that an unmodified MISO channel matrix is used for coherent reception. In a second stage, the effect of channel estimation error on receiver performance is modelled by perturbing the taps of the channel impulse responses

$$\widehat{h_{(n,m)}^{(tap)}} = h_{(n,m)}^{(tap)} + \varepsilon_{(n,m)}^{(tap)}. \tag{17}$$

$\varepsilon_{(n,m)}^{(tap)}$ are complex white gaussian variables with zero mean and a variance of

$$\sigma_\varepsilon^2 = \frac{p}{N_h}, \tag{18}$$

where p is given in percent of the normalized channel power.

IV. Results

The performance of STTD and M-STTD is compared using SER probability curves, plotted as a function of transmitted bit energy to received noise density (E_b/N_o). Additionally, the performances of Rake, ZF and MMSE SISO schemes as well as the approximated theoretical performance of STTD in a spatially uncorrelated flat Rayleigh fading radio channel [9] are plotted as reference cases.

Fig.3 displays Quadrature Phase Shift Keying (QPSK) SER performance in the Ped.A case, where all four paths of the power delay profile, including a 3dB Rice component, arrive within the first two chip periods. The Rake reception schemes have a tendency to reach error floors, whereas the symbol error rates of the equalized schemes continue to decline with increasing signal to noise ratio. Moreover, the equalized M-STTD schemes perform better than the other schemes. The performance benefit over the theoretical STTD scheme approximation (solid line) is due to the 3dB Rice component in the Ped.A case.

Fig.4 displays QPSK SER in the frequency selective fading Vehic.A case, where all signal power arrives through 6 fading taps within a window of 10 chips. Similar to the results in Fig.3, only the Rake reception schemes reach an error floor. However, the performance benefit of the diversity transmission schemes over their SISO scheme counter parts becomes marginal.

Introducing a channel estimation error of 4% in the Ped.A case (Fig.5), also the equalization based schemes tend to reach error floors. Nevertheless, the equalized M-STTD schemes still perform better than the others.

Fig.6 shows the symbol error performance for Gray encoded 16 QAM (Quadrature Amplitude Modulation) with and without a channel estimation error of 4%. The SISO reference cases are now omitted to improve readability. The trends are similar to the ones in Fig.3 to 5. Nevertheless, performance difference between ZF schemes and MMSE schemes are less pronounced at the plotted signal to noise ratios.

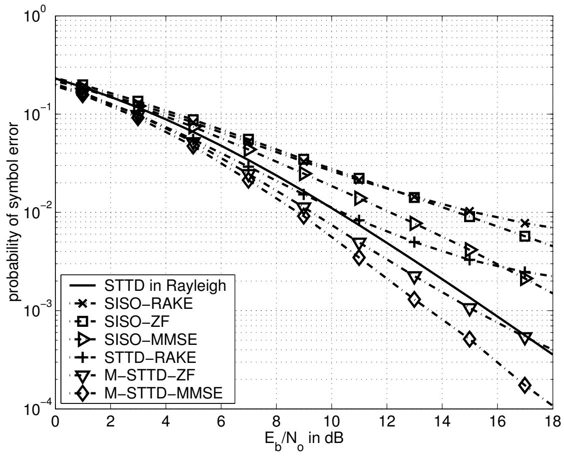

Fig. 3. QPSK SER Performance in the Ped.A Case

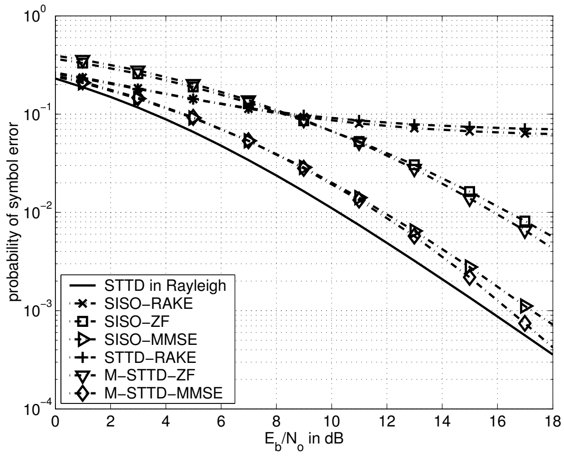

Fig. 4. SER Performance of QPSK Modulated Symbols in the Vehic.A Case

TABLE I
PROPAGATION ENVIRONMENT PARAMETERS

		Ped. A	Vehic. A
	PDP	ITU Generalised Pedestrian A	ITU Generalised Vehicular A
BS	PAS	Laplacian	Laplacian
BS	AoA	20°	20°
BS	AS	5°	5°
BS	Spacing	4λ	4λ
MS	PAS	uniform, 3dB Rice component at 22.5°	Laplacian
MS	AoA	0°	67.5°
MS	AS	360°	35°

V. Discussion

Orthogonal spreading codes are used in downlink transmission. Orthogonality, however, only exists between time aligned spreading codes. Time dispersion in the radio propagation channel is responsible for many time-shifted versions of the same data reaching the mobile. The Rake reception schemes are unable to restore orthogonality between time misaligned, spread signals, which causes IPI and MAI, and leads to the error floors in Fig.3 to Fig.6. Using ideal equalization, it is possible to counteract the effects of time dispersion. At the output of the equalizers the spread data symbols are time aligned and their orthogonality is preserved. All schemes with ideal equalization are therefore IPI and MAI free and do not reach error floors. If however channel estimation is not ideal (Fig.5 and Fig.6), it is not possible to completely counteract the effects of frequency selective fading. Reorthogonalisation of all users spread data

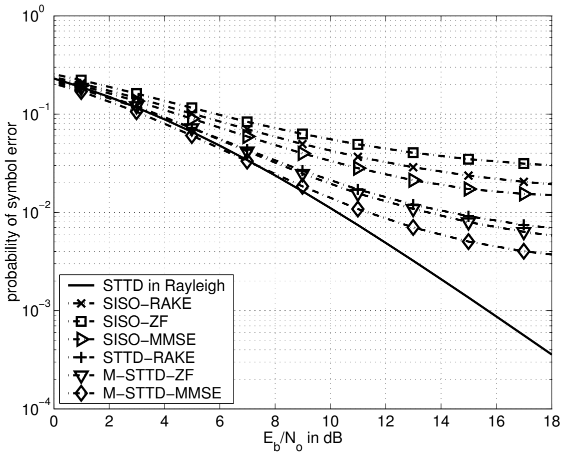

Fig. 5. SER Performance of QPSK Modulated Symbols in in the Ped.A Case with a Channel Estimation Error of 4%

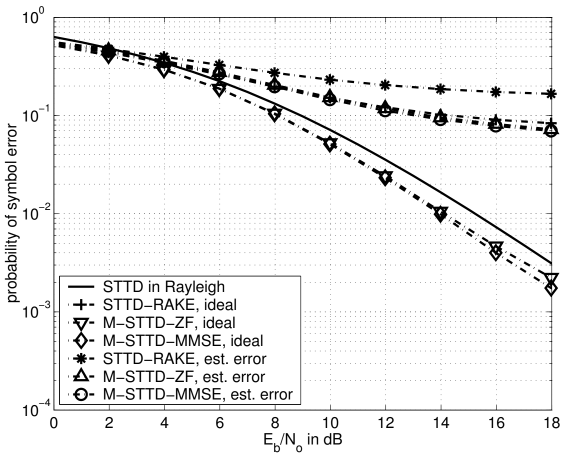

Fig. 6. SER Performance of 16 QAM Modulated Symbols in in the Ped.A Case with and without Channel Estimation Error

symbols can only partly be achieved. In these cases the equalizing schemes suffer also from MAI and IPI, and similar to their Rake reception counterparts reach error floors at higher E_b/N_os.

Furthermore, the good performance of the equalized M-STTD scheme in Ped.A case (Fig.3, Fig.5 and Fig.6) can be explained by the fact that in this relatively frequency flat fading case transmit diversity is the only available source of diversity protection. Hardly any diversity is available to the SISO schemes. On the contrary, the modest advantage of equalized M-STTD over the equalized SISO schemes in the Vehic.A case (Fig.4) is due to the fact that a considerable amount of frequency diversity can be exploited by all schemes. The additional protection of M-STTD due to transmit diversity results therefore only in a marginal benefit.

Increasing the modulation scheme order by going from QPSK to 16QAM (Fig.6), the distance between constellation points is decreasing, and it takes less perturbation due to noise or interference to cause a decoding error. That is why the Rake reception schemes reach error floors at relatively high levels and even the equalized schemes with a 4% channel estimation error struggle with a symbol error floor around 0.07.

VI. Conclusion

A space-time transmission scheme is introduced, where the STBC operation is placed at the end of the base band transmission chain. This modification enables joint space-time block decoding and linear MISO channel equalization. Assuming ideal channel knowledge at the receiver, the equalized modified STTD scheme is own cell interference free and reveals raw symbol error rate improvements compared to standard STTD and SISO schemes.

Introducing a channel estimation error of 4%, M-STTD becomes also interference limited, but continues to outperform all other schemes in the relatively frequency flat ITU Generalised Pedestrian A case.

As a considerable amount of frequency diversity can be exploited by all schemes in the spatially correlated ITU Generalised Vehicular A environment, the advantage of employing transmit diversity becomes less pronounced.

It can be concluded however that M-STTD is an interesting technique, which could for example enable higher order modulation transmit diversity communication in frequency selective fading channels.

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References

[1] Anja Klein, "Data detection algorithms specially designed for the downlink of CDMA mobile radio systems," in *IEEE 47th Vehicular Technology Conference*, 1997, vol. 1, pp. 203–207.

[2] Kari Hooli, Matti Latva-aho, and Markku Juntti, "Multiple access interferrence suppression with linear chip equalizers in WCDMA downlink receivers," in *Global Telecomunication Conference*, April 1999, vol. 1a, pp. 467–471.

[3] Siavash M. Alamouti, "A simple transmitter diversity scheme for wireless communications," *IEEE Journal on Selected Areas in Communications*, pp. 1451–1458, October 1998.

[4] 3GPP, "Physical channels and mapping of transport channels onto physical channels," Technical Specification TS 25.211 (V3.3.0), Technical Specification Group Radio Access Network, June 2000.

[5] Anand G. Dabak, Tim Schmidl, and Chaitali Sengupta, "Equalization and multi-user detection for space time block coding based transmit diversity (STTD) in frequency selective channels," in *VTC 2000*, 2000, pp. 506–513.

[6] Laurent Schumacher, Klaus I. Pedersen, and Jean-Philippe Kermoal, "A link-level MIMO radio channel simulator for evaluation of combined transmit/receive diversity concepts within the METRA project," in *Proceedings of IST Mobile Summit 2000*, Galway, October 2000, IST, pp. 515–520.

[7] Todd K. Moon and Wynn C. Stirling, *Mathematical Methods and Algorithems for Signal Processing*, Prentice Hall, Upper Saddler River, New Jersey, USA, 2000.

[8] MIMO Rapporteur, "MIMO conference call summary," TSG R1-02-0141, 3GPP, 2002.

[9] J. Lu, T.T. Tjhung, and C.C. Chai, "Error rates and tight bounds for L diversity reception of MQAM and MPSK signals in rayleigh fading," in *International Conference on Telecommunication*, Melbourne, April 97, pp. 205–210.