Comparison of Transmission Approaches in Vehicular Standardization MIMO Channels

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***Abstract*— A great variety of system concepts has been proposed to exploit the potentialities of MIMO systems towards future mobile communication systems. However, most of these transmission proposals make different assumptions with respect to channel properties.**
In this paper we apply different system concepts, namely Eigenbeamforming and V-BLAST, to the realistic vehicular channel models of 3GPP standardization to obtain a fair and uniform foundation of comparison between these different approaches by means of uncoded BER.

I. INTRODUCTION

Much effort has been spent to extend existing mobile communication systems with respect to an increased performance, not only because of the growing number of subscribers, but also due to the increasing demands regarding quality of service and data rate. The use of multiple receive antennas as well as multiple transmit antennas has been accepted as promising solution to achieve the aspired goals. Such systems are denoted as *multiple-input-multiple-output* (MIMO) systems.
The proposals to exploit the potentialities of MIMO systems are very variegated (e.g. BLAST [1] and Eigenbeamforming [2] in various versions). However, just as various as the system proposals are the required properties for the MIMO propagation channel. Different transmission proposals have often been developed for different channel properties and assumptions, sometimes even for the opposite assumptions (e.g. spatial correlations). As a consequence different system concepts are difficult to compare. Basic temporal and spatial features of a realistic MIMO channel model were defined in a proposal in the standardization discussions of the 3GPP [3], [4] which provided parameters and methods associated with modeling for realistic MIMO propagation channels. Applying different transmission schemes to these realistic channel models provides a fair and uniform basis for comparison.
In this paper we apply different system concepts, namely *Eigenbeamforming* (EB) and V-BLAST, to a realistic channel model of the 3GPP standardization to obtain a fair and uniform foundation of comparison between these different approaches. Section II reviews the model of the MIMO propagation channel, Section III describes the system setup of our communication link, Section IV explains the applied signal processing strategies and Section V gives some simulation results of the performance, i.e. the uncoded BER of the investigated systems.

II. THE STANDARDIZATION MIMO CHANNEL MODELS

The MIMO channel models proposed in the 3GPP standardization are designed to be applicable to link level simulations. The parameters of the channel models are defined by the power-delay-profile and the spatial correlations at transmitter and receiver side. The power-delay-profiles are backward compatible with the existing ITU [5] channel profiles. The correlation matrices are defined by the power-azimuth spectrum with a certain azimuth-spread, angle-of-arrival and array configuration. The models are given as tap-delay model. The mathematical representation of the channel impulse response $\boldsymbol{H}(t',t)$ for a system with M transmit antennas and N receive antennas can be written as

$$\boldsymbol{H}(t',t) = \sum_{k=1}^{K} \boldsymbol{H}_k(t')\delta(t-\tau_k), \qquad (1)$$

where K denotes the number of temporal taps, each with delay τ_k and the corresponding weighting $\boldsymbol{H}_k(t') \in \mathbb{C}^{N\times M}$. Thereby, the tap weights $\boldsymbol{H}_k(t')$ are time variant and have to be chosen such, that the desired antenna correlations are fulfilled. Four cases are specified:

1) **Case I:** Modified Pedestrian Type A
2) **Case II:** Vehicular Type A
3) **Case III:** Pedestrian Type B
4) **Case IV:** Single Path (uncorrelated)

To adjust the desired correlations [3], [4] at the *base station* (NB) and at the *user equipment* (UE) for each temporal tap k we compute the weighting $\boldsymbol{H}_k(t')$ as

$$\boldsymbol{H}_k(t') = \boldsymbol{C}_{k,\mathrm{UE}}\boldsymbol{\mathcal{Z}}(t')\boldsymbol{C}_{k,\mathrm{NB}}^{\mathrm{H}}, \qquad (2)$$

where $(\bullet)^{\mathrm{H}}$ denotes conjugate transpose and $\boldsymbol{\mathcal{Z}}(t')$ is a $N \times M$ zero-mean complex Gaussian matrix with unit variance realizing the time variant fading, e.g. the Doppler spectrum. The matrices $\boldsymbol{C}_k$ emerge from the eigenvalue decomposition of the desired correlation matrices $\boldsymbol{R}_k = \boldsymbol{U}_k\boldsymbol{\Lambda}_k\boldsymbol{U}_k^{\mathrm{H}}$ as

$$\boldsymbol{C}_k = \boldsymbol{U}_k\boldsymbol{\Lambda}_k^{1/2}. \qquad (3)$$

In the following we will assume block fading, where the taps $\boldsymbol{H}_k(t')$ are constant within a block of L transmissions. Additionally we will focus on the downlink.

III. SYSTEM MODEL

A. The Transmitter

At the transmitter the bit stream $b(t)$ is multiplexed into D substreams and mapped onto complex modulation symbols $\boldsymbol{s}(t)$. Each substream is given an individual power by multiplying with a diagonal matrix $\boldsymbol{P}^{1/2}$ and then a unitary beamforming matrix $\boldsymbol{T} \in \mathbb{C}^{M\times D}$ is applied. The resulting signal $\boldsymbol{x}(t)$ is pulse shaped with a *root-raised cosine* (RRC) impulse with roll-off factor $\alpha = 0,22$. Note, that the beamforming $\boldsymbol{T}$ has no FIR structure (see Fig. 1).

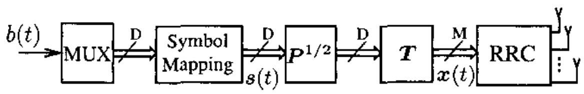

Fig. 1. System diagram for the transmitter (Base station)

B. The Receiver

The channel impulse response $\boldsymbol{H}(t',t)$ has tap delays which are no multiples of the symbol duration, cf. Eq. (1), the received signal $\boldsymbol{y}(t)$ cannot be sampled without losing information. Therefore, the received signal $\boldsymbol{y}(t)$ is matched filtered with a RRC pulse and then processed by a time-continuous *matched filter* (MF) with respect to the channel impulse response $\boldsymbol{H}(t',t)$ to produce $\boldsymbol{r}(t)$. By optimum detection theory [6], [7] the sampled sequence $\boldsymbol{r}(t)$ is a sufficient statistic containing all information about the transmitted signal $\boldsymbol{x}(t)$. The matched filter has to operate at continuous time. The output $\boldsymbol{r}(t)$ of the matched filter can be sampled at symbol rate T and consequently all subsequent steps can be regarded as time discrete. Note, that an analog *zero-forcing filter* (ZF) and an analog *Wiener filter* (WF) can always be decomposed into an analog MF with subsequent time-discrete ZF/WF [7].
The output of the analog MF is fed into the time-discrete receiver, which is either a linear receiver (MF, ZF, or WF) or a non-linear receiver (Decision-Feedback Equalizer), to produce the estimate $\hat{s}(t)$ of the transmitted signal $\boldsymbol{s}(t)$ (see Fig. 2).

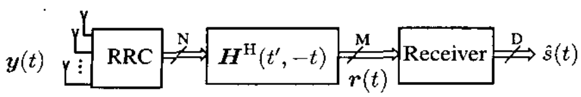

Fig. 2. System diagram for the receiver (User equipment)

C. Channel plus Matched Filter

Both parts, transmitter and receiver, are connected by the MIMO propagation channel from Eq. (1). If we include the transmit RRC filter into the channel as

$$\boldsymbol{H}_{\mathrm{RRC}}(t',t) = \sum_{k=1}^{K} \boldsymbol{H}_k(t') \cdot RRC(t-\tau_k), \qquad (4)$$

we can write

$$\boldsymbol{y}(t) = \boldsymbol{H}_{\mathrm{RRC}}(t',t) \star \boldsymbol{x}(t) + \boldsymbol{n}(t), \qquad (5)$$

where '$\star$' denotes convolution and $\boldsymbol{n}(t)$ is assumed to be additive Gaussian noise, temporally and spatially white with power σ_n^2. By combining the time-continuous channel (including the pulse-shaping) with the corresponding time-continuous MF (also including pulse-shaping) we can obtain a time-discrete block operating at symbol rate T without losing information due to the properties of a sufficient statistic. The continuous convolution of the channel $\boldsymbol{H}_{\mathrm{RRC}}(t',t)$ with the matched filter $\boldsymbol{H}^{\mathrm{H}}_{\mathrm{RRC}}(t',-t)$ reads as

$$\begin{aligned}\boldsymbol{H}_{\mathrm{MF}}(t',t) &= \boldsymbol{H}^{\mathrm{H}}_{\mathrm{RRC}}(t',-t) \star \boldsymbol{H}_{\mathrm{RRC}}(t',t) = \\ &= \sum_{k=1}^{K}\sum_{k'=1}^{K} \boldsymbol{H}^{\mathrm{H}}_k(t') \cdot \boldsymbol{H}_{k'}(t') \cdot RC(t-\tau_{k'}+\tau_k),\end{aligned} \qquad (6)$$

where $RC(t)$ is the raised-cosine impulse. Assuming block processing at the receiver we can combine the transmit signals $\boldsymbol{x}(t)$ from L time instances into one big space-time vector

$$\boldsymbol{x}_{\mathrm{st}} = \mathrm{vec}\left[\boldsymbol{x}(0),\ \boldsymbol{x}(T),\ \boldsymbol{x}(2T), \ldots, \boldsymbol{x}((L-1)\cdot T)\right]. \quad (7)$$

The same applies for the output $\boldsymbol{r}(t)$ of the MF to produce $\boldsymbol{r}_{\mathrm{st}}$. The space-time beamforming matrices are

$$\boldsymbol{T}_{\mathrm{st}} = \boldsymbol{1}^{L\times L} \otimes \boldsymbol{T}, \qquad \boldsymbol{P}_{\mathrm{st}} = \boldsymbol{1}^{L\times L} \otimes \boldsymbol{P}, \qquad (8)$$

where '$\otimes$' denotes the Kronecker-product and $\boldsymbol{1}^{L\times L}$ is the identity matrix. This produces the functional connection between the transmitted symbols $\boldsymbol{s}_{\mathrm{st}}$ and filter output $\boldsymbol{r}_{\mathrm{st}}$ in vector-matrix notation as

$$\boldsymbol{r}_{\mathrm{st}} = \boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}} \cdot \boldsymbol{x}_{\mathrm{st}} + \boldsymbol{n}_{\mathrm{st}} = \boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}} \boldsymbol{T}_{\mathrm{st}} \boldsymbol{P}^{1/2}_{\mathrm{st}} \boldsymbol{s}_{\mathrm{st}} + \boldsymbol{n}_{\mathrm{st}}, \quad (9)$$

where $\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}$ denotes a space-time block Toeplitz matrix comprising the whole time-continuous transmission elements in one block of length L. The (i,j)-th block of $\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}$ reads as

$$\begin{aligned}&\left[\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\right]_{(i,j)} = \\ &= \sum_{k=1}^{K}\sum_{k'=1}^{K} \boldsymbol{H}^{\mathrm{H}}_k(t') \cdot \boldsymbol{H}_{k'}(t') \cdot RC\left((i-j)T - \tau_{k'} + \tau_k\right)\end{aligned} \qquad (10)$$

The noise $\boldsymbol{n}_{\mathrm{st}}$ from Eq. (9) is the sampled noise after the receive matched filter. The noise $\boldsymbol{n}_{\mathrm{st}}$ is now spatially and temporally correlated. For one time instant t we have

$$\begin{aligned}\boldsymbol{n}_{\mathrm{st}}(t) &= \boldsymbol{H}^{\mathrm{H}}_{\mathrm{RRC}}(t',-t) \star \boldsymbol{n}(t) \\ &= \sum_{k=1}^{K} \boldsymbol{H}^{\mathrm{H}}_k(t') \cdot \boldsymbol{n}(t) \star RRC(\tau_k + t).\end{aligned} \qquad (11)$$

Stacking this into a space-time vector $\boldsymbol{n}_{\mathrm{st}}$ like Eq. (7) provides the space-time noise covariance matrix

$$\boldsymbol{R}^{\mathrm{st}}_{nn} = \mathrm{E}\left\{\boldsymbol{n}_{\mathrm{st}}\boldsymbol{n}^{\mathrm{H}}_{\mathrm{st}}\right\} = \sigma_n^2 \boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}. \qquad (12)$$

IV. Signal Processing Strategies

A. Transmit Processing

On the transmitter side we map D independent data streams onto M transmit antennas with the linear transformation $\boldsymbol{T}\boldsymbol{P}^{1/2}$. The choice of the number of transmitted data streams D, the power allocation $\boldsymbol{P}$ and the beamforming $\boldsymbol{T}$ is subject to the transmit processing strategy and depends on the channel properties: spatial correlations. Assuming perfect channel knowledge at the transmitter the beamforming $\boldsymbol{T}$ is chosen according to the eigenspace of the channel. The number of data streams D is chosen according to the number of dominant eigenvalues of the channel [2]. We assume no FIR beamforming.
Since the frequency selective channel $\boldsymbol{H}(t',t)$ is experiencing block fading with given average values

$$\boldsymbol{R}_{k,\mathrm{UE}} = \mathrm{E}\left\{\boldsymbol{H}_k(t')\boldsymbol{H}^{\mathrm{H}}_k(t')\right\} \qquad (13)$$

and

$$\boldsymbol{R}_{k,\mathrm{NB}} = \mathrm{E}\left\{\boldsymbol{H}^{\mathrm{H}}_k(t')\boldsymbol{H}_k(t')\right\} \qquad (14)$$

it is possible to perform the beamforming on two different time scales:

- on a short-term basis by computing the strongest eigenvectors of $[\boldsymbol{H}_{\mathrm{MF}}]_{(0,0)}[\boldsymbol{H}^{\mathrm{H}}_{\mathrm{MF}}]_{(0,0)}$ (inst. knowledge), or
- on a long-term basis by computing the strongest eigenvectors of $\sum_k \boldsymbol{R}_{k,\mathrm{NB}}$ (knowledge on average).

The eigen-spectrum in both cases allows to choose the rank D of the transmission. I.e., if the channel offers only one dominant eigenvalue it is reasonable to transmit only one scalar data stream. To maintain a constant data rate it is necessary to simultaneously increase the modulation level while decreasing the number of data streams. It is well-known, that the optimum strategy for allocating power on eigenmodes in a MIMO transmission is the water-filling approach [8]. However, this is only valid for Gaussian distributed signals. For realistic modulation signals the power allocation is much more sophisticated. For simplicity we apply uniform power allocation on all active data streams

$$\boldsymbol{P} = \frac{P_s}{D}\mathbf{1}_{D\times D}, \tag{15}$$

where P_s is the total transmit power. Note, that in the case where we do not have any channel knowledge at the transmitter we have to choose $D = M$ and $\boldsymbol{T} = \mathbf{1}$. The strategy to transmit parallel data streams over the channel is referred to as *spatial multiplexing.*

B. Linear Receive Processing

In the case of linear receive processing we will perform filtering $\boldsymbol{G}$ at the receiver with respect to the received signal $\boldsymbol{r}_{\mathrm{st}}$ in the full, $M \cdot L$-dimensional space-time to estimate the modulation symbols $\boldsymbol{x}_{\mathrm{st}}$. Thereafter we will reduce the space dimension to D by applying the reverse beamforming with $\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}$ and a scalar WF g to recover the amplitude before we perform the symbol decision. The linear filter is a block filter of one of the following: MF, ZF and WF (see Fig. 3). The resulting

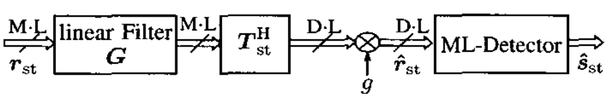

Fig. 3. System diagram of the linear receive filter.

signal $\hat{\boldsymbol{r}}_{\mathrm{st}}$ is basis for a *maximum likelihood* (ML) symbol decision (hard decision) of the actual transmitted signal in the space-only domain disregarding transmit signals at other time instants. The justification for this procedure is that the previous block filter has already removed the *inter-symbol-interference* (ISI) in the ideal case. Therefore the symbol decision can conveniently be performed in the space-only domain taking into account colored noise in the space domain. See also [9].

1) Matched Filter Receiver: The MF at the receiver has to be performed at continuous time. This time-continuous MF is placed directly after the pulse-shaping MF and consequently the linear filtering block in Fig. 3 reduces to the identity matrix $\boldsymbol{G}^{\mathrm{st}}_{\mathrm{MF}} = \mathbf{1}$. The scalar WF (see Fig. 3) computes as

$$g_{\mathrm{MF}} = \frac{\mathrm{tr}\left(\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\boldsymbol{T}_{\mathrm{st}}\boldsymbol{P}^{1/2}_{\mathrm{st}}\right)}{\mathrm{tr}\left(\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\boldsymbol{T}_{\mathrm{st}}\boldsymbol{P}_{\mathrm{st}}\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st,H}}_{\mathrm{MF}}\boldsymbol{T}_{\mathrm{st}}+\boldsymbol{R}^{\mathrm{st,MF}}_{\hat{n}\hat{n}}\right)} \tag{16}$$

and we can write

$$\begin{aligned}\hat{\boldsymbol{r}}_{st} &= g_{\mathrm{MF}}\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\left(\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\cdot\boldsymbol{x}_{\mathrm{st}} + \boldsymbol{n}_{\mathrm{st}}\right)\\ &= g_{\mathrm{MF}}\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\cdot\boldsymbol{x}_{\mathrm{st}} + \hat{\boldsymbol{n}}_{\mathrm{st}},\end{aligned} \tag{17}$$

with the new noise covariance matrix

$$\boldsymbol{R}^{\mathrm{st,MF}}_{\hat{n}\hat{n}} = g^2_{\mathrm{MF}}\mathrm{E}\left\{\hat{\boldsymbol{n}}_{\mathrm{st}}\hat{\boldsymbol{n}}^{\mathrm{H}}_{\mathrm{st}}\right\} = g^2_{\mathrm{MF}}\sigma^2_n\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\boldsymbol{T}_{\mathrm{st}}. \tag{18}$$

The result $\hat{\boldsymbol{r}}_{\mathrm{st}}$ is fed into a ML-detector which provides an estimate $\hat{\boldsymbol{s}}_{\mathrm{st}}$ of the originally transmitted modulation symbol $\boldsymbol{s}_{\mathrm{st}}$. The ML-decision for one vector-symbol $\hat{\boldsymbol{s}}(kT)$ reads as

$$\hat{\boldsymbol{s}}(kT) = \arg\min_{\boldsymbol{s}}\left\|\boldsymbol{r}(kT) - g_{\mathrm{MF}}\boldsymbol{T}^{\mathrm{H}}\boldsymbol{H}^{0}_{\mathrm{MF}}\boldsymbol{T}\boldsymbol{P}^{1/2}\boldsymbol{s}\right\|^2_{\boldsymbol{R}^{-1}_{\hat{n}\hat{n}}}, \tag{19}$$

with $\boldsymbol{H}^{0}_{\mathrm{MF}} = \boldsymbol{H}_{\mathrm{MF}}(t', t = 0)$, where only the (k, k)-th block $\boldsymbol{R}_{\hat{n}\hat{n}}$ of the space-time noise covariance matrix $\boldsymbol{R}^{\mathrm{st,MF}}_{\hat{n}\hat{n}}$ applies as noise statistic for the ML estimate.

2) Zero-Forcing Receiver: As stated in Section III, the time-continuous ZF can be decomposed into a time-continuous MF with subsequent time-discrete ZF stage $\boldsymbol{G}^{\mathrm{st}}_{\mathrm{ZF}}$. Right after the ZF stage we apply the space-time beamforming matrix from Eq. (8) and a scalar Wiener Filter g_{ZF} to recover the correct signal amplitude. This produces the filter output $\hat{\boldsymbol{r}}_{\mathrm{st}}$

$$\hat{\boldsymbol{r}}_{\mathrm{st}} = g_{\mathrm{ZF}}\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{G}^{\mathrm{st}}_{\mathrm{ZF}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\cdot\boldsymbol{x} + \hat{\boldsymbol{n}}_{\mathrm{st}} \tag{20}$$

where the zero-forcing stage $\boldsymbol{G}^{\mathrm{st}}_{\mathrm{ZF}}$ computes as

$$\boldsymbol{G}^{\mathrm{st}}_{\mathrm{ZF}} = \boldsymbol{H}^{\mathrm{st},-1}_{\mathrm{MF}} \tag{21}$$

with space-time noise covariance matrix

$$\boldsymbol{R}^{\mathrm{st,ZF}}_{\hat{n}\hat{n}} = g^2_{\mathrm{ZF}}\mathrm{E}\left\{\hat{\boldsymbol{n}}_{\mathrm{st}}\hat{\boldsymbol{n}}^{\mathrm{H}}_{\mathrm{st}}\right\} = g^2_{\mathrm{ZF}}\sigma^2_n\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{H}^{\mathrm{st},-1}_{\mathrm{MF}}\boldsymbol{T}_{\mathrm{st}} \tag{22}$$

and scalar Wiener Filter $g_{\mathrm{ZF}} = L/\,\mathrm{tr}\left(\boldsymbol{P}_{\mathrm{st}}+\boldsymbol{R}^{\mathrm{st,ZF}}_{\hat{n}\hat{n}}\right)$ to get $\hat{\boldsymbol{r}}_{\mathrm{st}}$. The result is fed into a ML estimator which provides an estimate $\hat{\boldsymbol{s}}(t)$ of the originally transmitted modulation symbol $\boldsymbol{s}(t)$. This reads, in analogy to the MF receiver, as

$$\hat{\boldsymbol{s}}(kT) = \arg\min_{\boldsymbol{s}}\left\|\boldsymbol{r}(kT) - g_{\mathrm{ZF}}\boldsymbol{P}^{1/2}\boldsymbol{s}\right\|^2_{\boldsymbol{R}^{-1}_{\hat{n}\hat{n}}}. \tag{23}$$

3) Wiener Filter Receiver: Also the time-continuous WF can be decomposed into a time-continuous MF with subsequent time-discrete WF stage $\boldsymbol{G}^{\mathrm{st}}_{\mathrm{WF}}$. Right after the WF stage we apply the space-time beamforming matrix from Eq. (8). This produces the filter output $\hat{\boldsymbol{r}}_{\mathrm{st}}$ (see Fig. 3)

$$\hat{\boldsymbol{r}}_{\mathrm{st}} = \boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{G}^{\mathrm{st}}_{\mathrm{WF}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\cdot\boldsymbol{x} + \hat{\boldsymbol{n}}_{\mathrm{st}} \tag{24}$$

where the WF stage $\boldsymbol{G}^{\mathrm{st}}_{\mathrm{WF}}$ computes as

$$\boldsymbol{G}^{\mathrm{st}}_{\mathrm{WF}} = \boldsymbol{R}^{\mathrm{st}}_{xx}\boldsymbol{H}^{\mathrm{st,H}}_{\mathrm{MF}}\left(\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\boldsymbol{R}^{\mathrm{st}}_{xx}\boldsymbol{H}^{\mathrm{st,H}}_{\mathrm{MF}} + \sigma^2_n\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\right)^{-1}, \tag{25}$$

with $\boldsymbol{R}^{\mathrm{st}}_{xx} = \boldsymbol{T}_{\mathrm{st}}\boldsymbol{P}_{\mathrm{st}}\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}$ and space-time noise covariance

$$\boldsymbol{R}^{\mathrm{st,WF}}_{\hat{n}\hat{n}} = \sigma^2_n\boldsymbol{T}^{\mathrm{H}}_{\mathrm{st}}\boldsymbol{G}^{\mathrm{st}}_{\mathrm{WF}}\boldsymbol{H}^{\mathrm{st}}_{\mathrm{MF}}\boldsymbol{G}^{\mathrm{st,H}}_{\mathrm{WF}}\boldsymbol{T}_{\mathrm{st}}. \tag{26}$$

The result $\hat{\boldsymbol{r}}_{\mathrm{st}}$ of the WF is fed into the ML estimator to provide an estimate $\hat{\boldsymbol{s}}_{\mathrm{st}}$ of the originally transmitted modulation symbol $\boldsymbol{s}_{\mathrm{st}}$ where for one symbol-vector evolves

$$\hat{\boldsymbol{s}}(kT) = \arg\min_{\boldsymbol{s}}\left\|\boldsymbol{r}(kT) - \boldsymbol{H}_{\mathrm{WF}}(k)\boldsymbol{s}\right\|^2_{\boldsymbol{R}^{-1}_{\hat{n}\hat{n}}}, \tag{27}$$

where only the (k,k)-th block $\boldsymbol{R}_{\hat{n}\hat{n}}$ of the space-time noise covariance matrix $\boldsymbol{R}_{\hat{n}\hat{n}}^{\mathrm{st,WF}}$ again applies as noise statistic for the ML estimate and $\boldsymbol{H}_{\mathrm{WF}}(k)$ denotes the (k,k)-th block of the combined space-time transfer function $\boldsymbol{H}_{\mathrm{WF}}^{\mathrm{st}} = \boldsymbol{T}_{\mathrm{st}}^{\mathrm{H}}\boldsymbol{G}_{\mathrm{WF}}^{\mathrm{st}}\boldsymbol{H}_{\mathrm{MF}}^{\mathrm{st}}\boldsymbol{T}_{\mathrm{st}}\boldsymbol{P}_{\mathrm{st}}^{1/2}$. Note, that $g_{\mathrm{WF}} \equiv 1$.

C. Non-linear Receive Processing: DFE-Receiver

We consider the *decision-feedback-equalizer* (DFE) as non-linear receiver scheme [1], [10] where we use a block ZF as feed-forward filter (ZF-DFE). In this case, the filtered and sampled receive signal $\boldsymbol{r}_{\mathrm{st}}$ will be processed iteratively to directly produce the estimated modulation symbols $\hat{s}_{\mathrm{st}}$ in the following manner:

1) Compute the complete space-time transfer function
$$\boldsymbol{H} = \boldsymbol{H}_{\mathrm{MF}}^{\mathrm{st}}\boldsymbol{T}_{\mathrm{st}}\boldsymbol{P}_{\mathrm{st}}^{1/2} \tag{28}$$
2) Compute the ZF filter $\boldsymbol{G}_{\mathrm{ZF}}^{\mathrm{st}} = \boldsymbol{H}^{+}$, where $\boldsymbol{H}^{+}$ denotes the pseudo-inverse of matrix $\boldsymbol{H}$
3) Choose the row $\boldsymbol{w}_{k_i}^{\mathrm{T}}$ of $\boldsymbol{G}_{\mathrm{ZF}}^{\mathrm{st}}$ with the smallest norm
4) Compute $\hat{r}_{k_i} = \boldsymbol{w}_{k_i}^{\mathrm{T}}\boldsymbol{r}_{\mathrm{st}}$
5) Decide modulation symbol $\hat{s}_{k_i}$
6) Update receive signal $\boldsymbol{r}_{\mathrm{st}}$ by subtracting the ISI generate by symbol $\hat{s}_{k_i}$
$$\boldsymbol{r}_{\mathrm{st}} = \boldsymbol{r}_{\mathrm{st}} - \boldsymbol{h}_{k_i}\hat{s}_{k_i}, \tag{29}$$
where $\boldsymbol{h}_{k_i}$ denotes column k_i of the channel $\boldsymbol{H}$.
7) Reduce the channel matrix $\boldsymbol{H}$ by column k_i and continue with Step 2).

V. Simulations

For the simulations we are transmitting at a fixed data rate of 8 bits per channel use over a symmetrical 4×4 MIMO system. We restrict ourselves to one channel scenario: Channel scenario II which corresponds to a MIMO channel scenario with a vehicular power-delay-profile consisting of six temporal taps and fairly high spatial correlations in the case of $\lambda/2$ antenna spacing and fairly low spatial correlations in the case of 4λ antenna spacing. We average over 500 channel realizations where we transmit $L = 60$ symbols in each channel realization.

In the following simulations the applied channel knowledge (short-term or long-term) is assumed to be perfectly known.

A. Channel Case II

Fig. 4-6 show the raw BER over transmit SNR for different transmission modes and receiver and transmitter types:

- EB corresponds to a system concept where the transmitter has knowledge about the channel either on a short-term or long-term basis. To this end, the transmitter is equipped with an antenna array with $\lambda/2$ antenna spacing. Different receiver types (MF, ZF, WF, DFE) are compared.
- V-BLAST corresponds to a system concept which tries to exploit the diversity provided by the channel. To this end, a non-linear receiver (DFE) is combined with a simple transmission approach where the transmitter has no channel knowledge. To enable a high diversity order, the transmitter is equipped with an antenna array with 4λ antenna spacing.

We compare different transmission modes: 1×8 bits, 2×4 bits and 4×2 bits, where the first two approaches correspond to EB ($\lambda/2$ antenna spacing at transmitter with channel knowledge and different receiver types), while the latter one corresponds to V-BLAST (4λ antenna spacing at transmitter with no channel knowledge and DFE receiver).

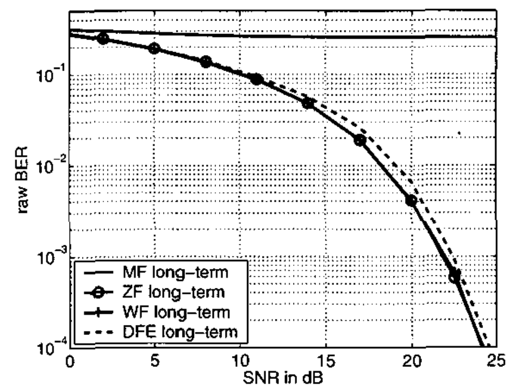

Fig. 4. Raw BER as function of the transmit SNR for Standardization MIMO channel case II transmitting one 256-QAM data stream with EB and different receiver types.

In Fig. 4 we transmit one 256-QAM data stream with EB. Consequently we do not have space diversity, only time diversity and antenna gain. The given power-delay-profile [4] offers six temporal taps, however only the first two are significant; the others are nearly negligible due to their larger attenuation ($\geq$ 10dB). Therefore the DFE receiver does not have a much steeper slope compared to the linear receivers. The ZF and WF receiver are almost identical. Obviously the ZF has enough degrees of freedom to successfully invert the channel. Short-term channel knowledge at the transmitter provides no performance gain.

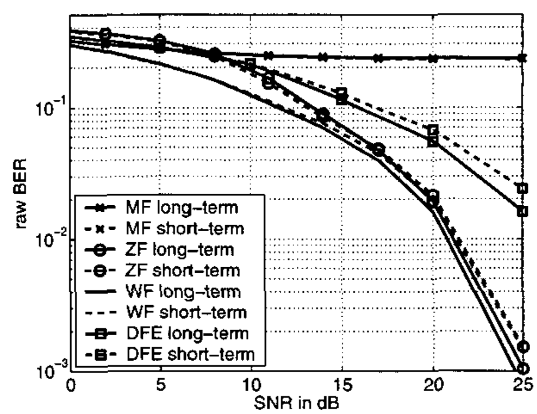

Fig. 5. Raw BER as function of the transmit SNR for Standardization MIMO channel case II transmitting two 16-QAM data streams with EB and different receiver types.

In Fig. 5 we transmit two 16-QAM data streams. We see in comparison to Fig. 4 that all curves are shifted to the right which is due to the fact that we now also use one weaker eigenmode which has worse performance. In this context we see that the DFE receiver is suffering most from this fact so we don't even see the cross-over point with the linear receiver schemes. In Fig. 5 we see that short-term processing loses at high SNR compared to long-term processing. This can be explained by the fact, that short-term processing perfectly distinguishes the data streams onto the eigenmodes contrary to long-term processing. In cases where the second strongest eigenmode is very weak, the long-term processing approach can still recover parts of the 'weak' data stream from the stronger eigenmode, contrary to the short-term processing approach. However, note that this result will change for perfect power allocating and stream adaptation (bit-loading).

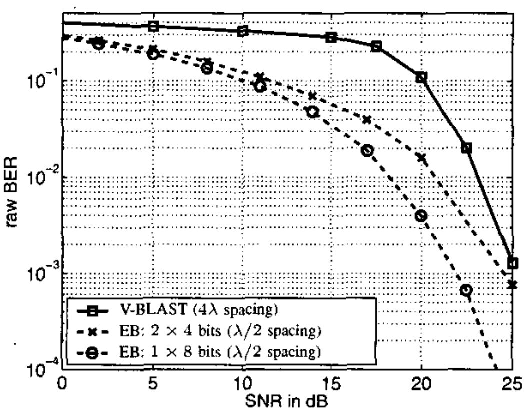

Fig. 6. Raw BER as function of the transmit SNR for Standardization MIMO channel case II comparing different system concepts and modes.

In Fig. 6 we compare the different concepts of EB and V-BLAST. Thereby we assume long-term channel knowledge and $\lambda/2$ antenna spacing at the transmitter and a WF receiver in the case of EB. In the case of V-BLAST we assume 4λ antenna spacing at the transmitter and a DFE receiver. It can be seen, that the use of channel knowledge at the transmitter (even long-term knowledge only) in combination with a linear receiver outperforms a system concept with no transmit processing and non-linear receiver exploiting diversity in this realistic channel model at moderate SNR regions.

B. Complexity Issues

1) Transmitter: Computational complexity at the transmitter is directly connected with the usage of channel knowledge. No channel knowledge at the transmitter of course requires the least effort while the application of channel knowledge at the transmitter depends on the time scale, in which this channel knowledge is attained and utilized. From the simulation results we can come to the following conclusions:

- Knowledge about long-term properties of the channel is not a tremendous burden since the long-term properties remain constant over a significant amount of time (feedback and exact estimation). Utilizing long-term knowledge already leads to a significant performance gain due to the correlations in the channel (here: channel case II).
- To obtain a performance gain with short-term channel knowledge it is not sufficient to apply only the short-term eigenbase. Only power allocation taking into account the used modulation alphabet in combination with the short-term eigenbase yields a benefit compared to long-term processing. However at much higher computational effort.

It should be mentioned that long-term channel knowledge at the transmitter can be seen as a realistic assumption due to its slow changing behavior, while the availability of short-term channel knowledge is realistic in only very few applications.

2) Receiver: Utilizing linear receiver techniques in combination with EB already leads to a good performance of the communication system (cf. Fig. 4 and 5). However, no receive diversity can be exploited with linear receiver techniques [11]. Utilizing the non-linear receiver scheme at the receiver enables the exploitation of receive diversity, however at a much higher computational burden. Note, that the DFE principle has to be applied on symbol basis to obtain best performance. Simplifying the computational amount for example by applying the DFE principle to a vector-symbol already leads to degradation of performance.

VI. CONCLUSIONS

In this paper we have applied different transmitter and receiver approaches to a realistic vehicular channel model proposed by the 3GPP standardization body. We have evaluated different system concepts with respect to the BER for different numbers of data streams while keeping the data rate constant. The results show, that for these channel types long-term knowledge at the transmitter yields a good performance which can already be achieved with linear receiver schemes.

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