Paper

Transmit Diversity in the Downlink for the TETRA-TEDS System

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Abstract—In the paper a proposal for the improvement of performance for the TETRA Enhanced Data System (TEDS) employing transmit diversity based on two antennas in the downlink is described. The key idea of the considerations relies on using the space-frequency coding algorithm. The proposal described required some relatively simple changes to the existing TEDS's Single Input Single Output (SISO) interface but the original number of payload and signaling symbols in the normal downlink burst is preserved. The simulation results obtained indicate a significant improvement in performance. The Eb/No parameter could be reduced from 5 to 8 dB with respect to Frame Error Rate (FER), compared to a single antenna transmission for the same FER = 10^{-3}.

Keywords—multiple input multiple output (MIMO), TETRA Enhanced Data System.

1. Introduction

For more than two decades worldwide huge development in mobile digital communication systems with continually improving performance can be observed, increasing throughput and an enlarging pool of services available for users. One of the known techniques employed in this progress is Multiple Input Multiple Output (MIMO) due to its ability to form different routes for the transmission of signals over the radio fading channel.

The TETRA Enhanced Data System (TEDS) [1]–[3] with its radio interface based on filtered multitone modulation (FMT) [4] is suitable for the implementation of MIMO. However, due to the relatively small dimensions of a mobile terminal, the simple version of MIMO – called Multiple Input Single Output (MISO) – is reasonable for TEDS with two transmit antennas at the base station and a single receive antenna at the mobile terminal.

Furthermore, such a MISO technique is simple and does not require significant changes in the radio interface.

2. Implementation of MISO in the TEDS Radio Interface

In the following, the implementation of MISO based on the Alamouti algorithm [5], [6], for the TEDS radio interface, is described. With FMT modulation the baseband time-continuous signal in TEDS is given by [1], [2]:

$$s(t) = \sum_{n=0}^{N-1} \sum_{k=0}^{K-1} a_n^{(k)} g(t-nT) e^{j(2\pi/T)nk\zeta t}, \qquad (1)$$

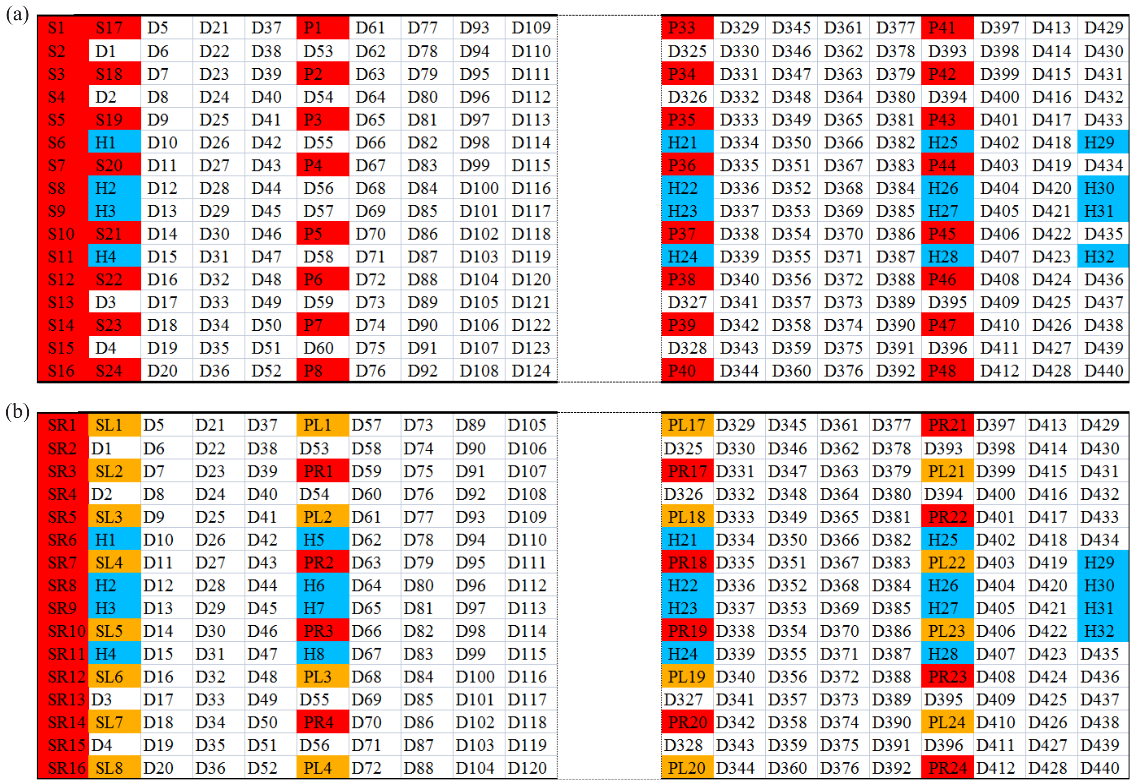
where n is the generic symbol within the burst, k is the index of the subcarrier, and K and N are the number of subcarriers and the transmitted (multicarrier) symbols, respectively.

Moreover, the impulse response of the square-root raised cosine filter has roll-off $\alpha = 0.2$, the signaling interval is $T = 1/2400$ s and the subcarrier spacing is $\Delta f = \zeta/T = 2700$ Hz, Thus, $\zeta = \Delta f \cdot T = 1.125$ is the measure of interference between neighboring subcarriers. This means that the frequency occupancy of each subcarrier is $(1+\alpha)/T = 2880$ s. As is known, the TEDS interface can be used for channels having the bandwidths: 25 kHz, 50 kHz, 100 kHz and 150 kHz. In order to explain the Double Input Single Output (DISO) for the downlink in the TEDS interface, the structure of Normal Downlink Burst (NDB), for each of the channels should be considered. As an example, in the following the NDB with $K = 16$ subcarriers (50 kHz channel) is described (see Fig. 1a). The burst contains the payload symbols (D marks), the header symbols (H marks), the pilot symbols (P marks) and the synchronization symbols (S marks).

To allow for the reception of signals transmitted by two antennas it is necessary to adequately locate the payload, pilot and synchronization symbols in the two symbol streams. Now, the pilot symbols should enable the receiver to carry out the estimation of channel characteristics for each of the symbol streams, and the synchronization symbols should provide efficient receiver synchronization. Of course, the number of payload and signaling symbols in the DISO and SISO schemes must remain identical.

Thus, it is not an easy task to comply with these requirements. A proposal for NDB burst adapted for DISO in the downlink is shown in Fig. 1b [7]. The pilot and synchronization symbols transmitted by the first and second antenna are marked by R and L, respectively. In the original Alamouti algorithm two symbols representing a pair are transmitted in two consecutive symbol times. Since the symbol time is small compared to the coherence time of the channel, the authors assume that, in practice, the channel characteristics are almost the same for both symbols.

Of course, in the multicarrier system the channel characteristic is a function of time and frequency [8]. However, if the channel characteristic is quasi-stationary in the small time interval (for two symbols), it is also quasi-stationary in the small frequency spacing concerning two neighboring subcarriers. As a result, a pair of symbols in the Alamouti algorithm may be represented either in frequency or time domain. The Alamouti algorithm in the proposal consid-



(a)

S1	S17	D5	D21	D37	P1	D61	D77	D93	D109	P33	D329	D345	D361	D377	P41	D397	D413	D429
S2	D1	D6	D22	D38	D53	D62	D78	D94	D110	D325	D330	D346	D362	D378	D393	D398	D414	D430
S3	S18	D7	D23	D39	P2	D63	D79	D95	D111	P34	D331	D347	D363	D379	P42	D399	D415	D431
S4	D2	D8	D24	D40	D54	D64	D80	D96	D112	D326	D332	D348	D364	D380	D394	D400	D416	D432
S5	S19	D9	D25	D41	P3	D65	D81	D97	D113	P35	D333	D349	D365	D381	P43	D401	D417	D433
S6	H1	D10	D26	D42	D55	D66	D82	D98	D114	H21	D334	D350	D366	D382	H25	D402	D418	H29
S7	S20	D11	D27	D43	P4	D67	D83	D99	D115	P36	D335	D351	D367	D383	P44	D403	D419	D434
S8	H2	D12	D28	D44	D56	D68	D84	D100	D116	H22	D336	D352	D368	D384	H26	D404	D420	H30
S9	H3	D13	D29	D45	D57	D69	D85	D101	D117	H23	D337	D353	D369	D385	H27	D405	D421	H31
S10	S21	D14	D30	D46	P5	D70	D86	D102	D118	P37	D338	D354	D370	D386	P45	D406	D422	D435
S11	H4	D15	D31	D47	D58	D71	D87	D103	D119	H24	D339	D355	D371	D387	H28	D407	D423	H32
S12	S22	D16	D32	D48	P6	D72	D88	D104	D120	P38	D340	D356	D372	D388	P46	D408	D424	D436
S13	D3	D17	D33	D49	D59	D73	D89	D105	D121	D327	D341	D357	D373	D389	D395	D409	D425	D437
S14	S23	D18	D34	D50	P7	D74	D90	D106	D122	P39	D342	D358	D374	D390	P47	D410	D426	D438
S15	D4	D19	D35	D51	D60	D75	D91	D107	D123	D328	D343	D359	D375	D391	D396	D411	D427	D439
S16	S24	D20	D36	D52	P8	D76	D92	D108	D124	P40	D344	D360	D376	D392	P48	D412	D428	D440

(b)

SR1	SL1	D5	D21	D37	PL1	D57	D73	D89	D105	PL17	D329	D345	D361	D377	PR21	D397	D413	D429
SR2	D1	D6	D22	D38	D53	D58	D74	D90	D106	D325	D330	D346	D362	D378	D393	D398	D414	D430
SR3	SL2	D7	D23	D39	PR1	D59	D75	D91	D107	PR17	D331	D347	D363	D379	PL21	D399	D415	D431
SR4	D2	D8	D24	D40	D54	D60	D76	D92	D108	D326	D332	D348	D364	D380	D394	D400	D416	D432
SR5	SL3	D9	D25	D41	PL2	D61	D77	D93	D109	PL18	D333	D349	D365	D381	PR22	D401	D417	D433
SR6	H1	D10	D26	D42	H5	D62	D78	D94	D110	H21	D334	D350	D366	D382	H25	D402	D418	D434
SR7	SL4	D11	D27	D43	PR2	D63	D79	D95	D111	PR18	D335	D351	D367	D383	PL22	D403	D419	H29
SR8	H2	D12	D28	D44	H6	D64	D80	D96	D112	H22	D336	D352	D368	D384	H26	D404	D420	H30
SR9	H3	D13	D29	D45	H7	D65	D81	D97	D113	H23	D337	D353	D369	D385	H27	D405	D421	H31
SR10	SL5	D14	D30	D46	PR3	D66	D82	D98	D114	PR19	D338	D354	D370	D386	PL23	D406	D422	H32
SR11	H4	D15	D31	D47	H8	D67	D83	D99	D115	H24	D339	D355	D371	D387	H28	D407	D423	D435
SR12	SL6	D16	D32	D48	PL3	D68	D84	D100	D116	PL19	D340	D356	D372	D388	PR23	D408	D424	D436
SR13	D3	D17	D33	D49	D55	D69	D85	D101	D117	D327	D341	D357	D373	D389	D395	D409	D425	D437
SR14	SL7	D18	D34	D50	PR4	D70	D86	D102	D118	PR20	D342	D358	D374	D390	PL24	D410	D426	D438
SR15	D4	D19	D35	D51	D56	D71	D87	D103	D119	D328	D343	D359	D375	D391	D396	D411	D427	D439
SR16	SL8	D20	D36	D52	PL4	D72	D88	D104	D120	PL20	D344	D360	D376	D392	PR24	D412	D428	D440

Fig. 1. Burst structures in the downlink for: (a) SISO and (b) DISO.

ered has been adapted to the format of the transmitted signal in such way that the time-dependent variable in the original algorithm has been replaced with frequency-dependent variable. In this way a space-frequency coding has been achieved and the dependence on the speed of changes in channel characteristics has been minimized. Figure 2 shows the fragment of the burst in which the data symbols $a_n^{(k)}$ presented in Eq. (1) are assigned to the suitable antennas and subcarriers [7].

As can be seen in Fig. 2, it is possible to separate $\left(a_n^{(k)}, a_{n+1}^{(k+1)}\right)$, $\left(a_{n+2}^{(k+2)}, a_{n+3}^{(k+3)}\right)$ etc., which appear in both channels on the subcarriers having the same index. Furthermore, the symbols in each pair are assigned to neighboring subcarriers. Since the frequency spacing between the neighboring subcarriers is small, one can assume that each channel associated with a given antenna produces almost the same effect on both symbols of a pair.

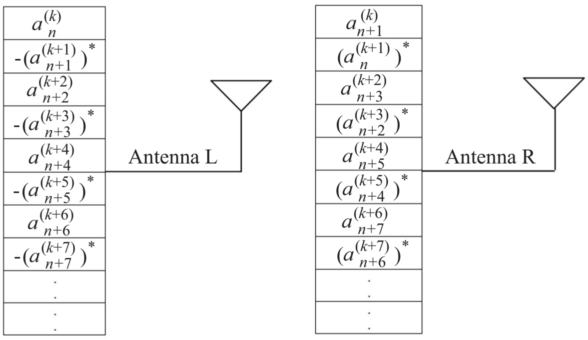


Fig. 2. Assignment of successive data symbols (indicated by the subscripts) in the burst to the subcarriers (indicated by the superscripts) and antennas, where symbol $()^*$ is the complex conjugate of the argument.

The received pair of symbols can be written as the sum of transmitted symbols multiplied by the corresponding channels' frequency responses (channel coefficients) and the noise samples:

$$\begin{aligned} r_{n,n+1}^{(k)} &= H_L^{(k,k+1)} a_n^{(k)} + H_R^{(k,k+1)} a_{n+1}^{(k)} + z_1 \\ r_{n,n+1}^{(k+1)} &= -H_L^{(k,k+1)} \left(a_{n+1}^{(k)}\right)^* + H_R^{(k,k+1)} \left(a_n^{(k)}\right)^* + z_2 \end{aligned} , \qquad (2)$$

where $r_{n,n+1}^{(k)}$, $r_{n,n+1}^{(k+1)}$ represent the combination of both symbols n and $n+1$ received on the subcarriers k and $k+1$, respectively, while $H_L^{(k,k+1)}$ and $H_R^{(k,k+1)}$ are the channels' coefficients associated with the first and second antenna and evaluated jointly for both neighboring subcarriers, whereas z_1 and z_2 are the noise samples. The symbols are evaluated as mean values separately for a real part and an imaginary part of each of the channel coefficients on both neighboring subcarriers.

However, as can be seen in Fig. 1b, some payload symbols do not appear on neighboring subcarriers and they are separated by synchronization or pilot symbols. In such cases the authors assume a channel coefficient for a pair of symbols corresponding to the synchronization or pilot

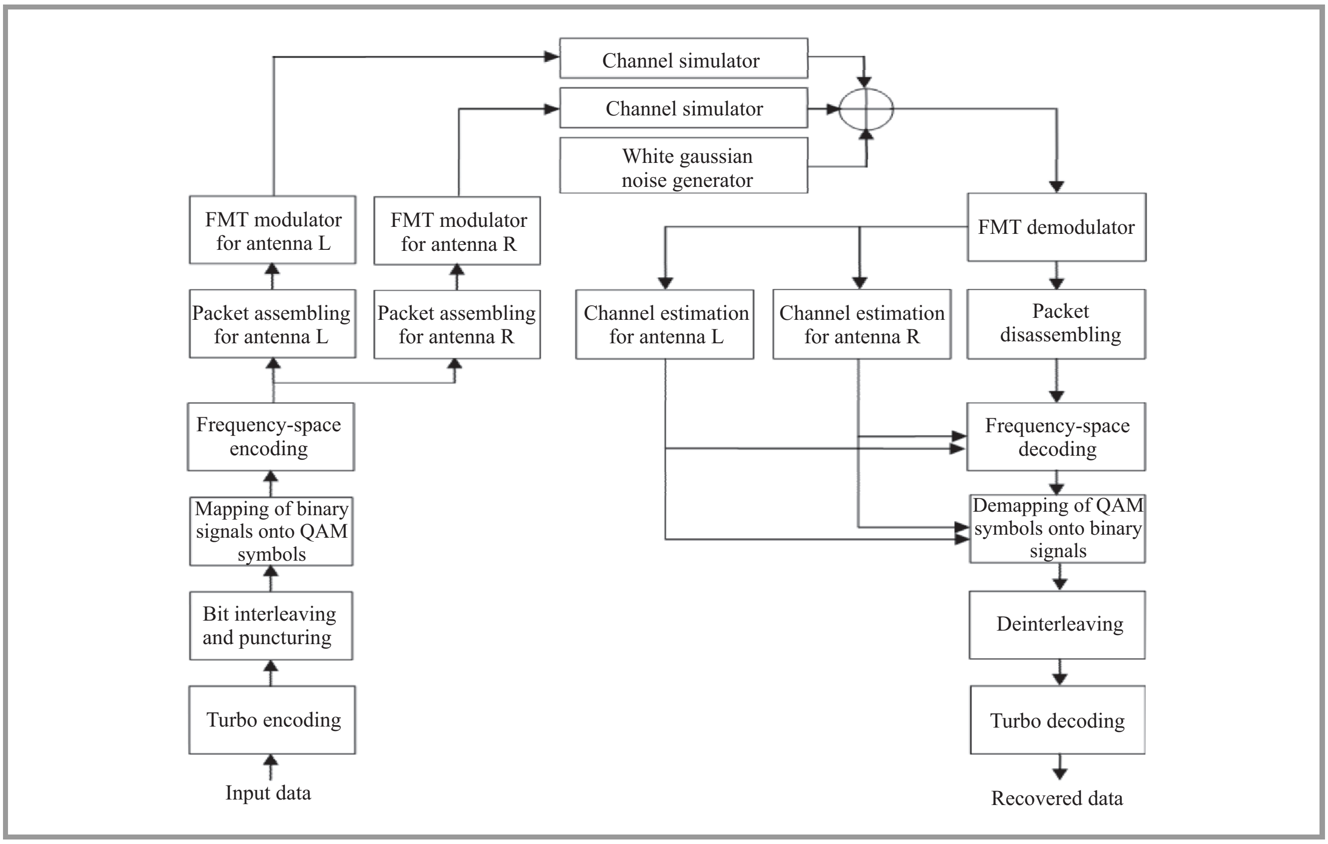


Fig. 3. Block schematic of the simulated system.

symbol. To improve the reception performance, the number of these events is reduced by the adequate distribution of payload, pilot and synchronization symbols within the burst.

The reception rule is based on Maximum-Ratio-Combining (MRC) [5], [6]:

$$\begin{aligned} \hat{r}_n^{(k)} &= \left(H_L^{(k,k+1)}\right)^* r_{n,n+1}^{(k)} + H_R^{(k,k+1)} r_{n,n+1}^{(k+1)} \\ \hat{r}_{n+1}^{(k+1)} &= \left(H_R^{(k,k+1)}\right)^* r_{n,n+1}^{(k)} - H_L^{(k,k+1)} r_{n,n+1}^{(k+1)} \end{aligned} , \quad (3)$$

In the Eqs. given by (3), a soft value for each received bit $b_{\hat{r}}$ of each symbol is calculated provided that a given bit b was transmitted. To obtain the likelihood of this bit such symbol $a_i^{b^{(m)}=x}$, $x \in \{0,1\}$, must be found in the constellation of symbols for QAM modulation which takes on the value x in the m-th position of a group of bits representing that symbol and minimizes the function

$$\begin{aligned} &-\log\left(p(b_{\hat{r}} \mid b^{(m)} = x)\right) \approx \\ &\min_i \left(\hat{r}_n^{(k)} - \left(\left|\widehat{H}_R^{k,k+1}\right|^2 + \left|\widehat{H}_L^{k,k+1}\right|^2\right) a_i^{b^{(m)}=x}\right)^2 . \quad (4) \end{aligned}$$

This approach is employed for each bit of each symbol in the received sequence of symbols.

3. Simulator for the Transmit Diversity of the TEDS Interface and Simulation Results

To investigate the performance of the above described transmit diversity method in the downlink the simulator shown in Fig. 3 has been developed [7]. The FMT modulator and demodulator used in the investigations are based on the overlap-add algorithm [9]. The rate 1/3 turbo encoder is formed by two recursive systematic convolutional encoders with 8 states each, separated by an interleaver [10]. The code rates 1/2 or 2/3 can be obtained by adequately puncturing the turbo code sequence. In the iterative turbo decoder the Max-Log-Map algorithm [6], [11] was used and the number of iterations is 10.

The selected results of simulations are shown in Figs. 4–7. They represent the relationships between FER and Eb/No for the system identified by: 16 subcarriers (50 kHz bandwidth), 1/2 code rate, 4QAM modulation on each subcarrier, the downlink transmission on 400 MHz in the typical urban (TU) and hilly terrain (HT) propagation profiles [3] and terminal speed of 50 km/h and 200 km/h, respectively.

The curves in the figures denoted by *TD* represent the results obtained when transmit diversity is employed, and the curves with that denotation missing correspond to the SISO operation. It can be seen from the figures that

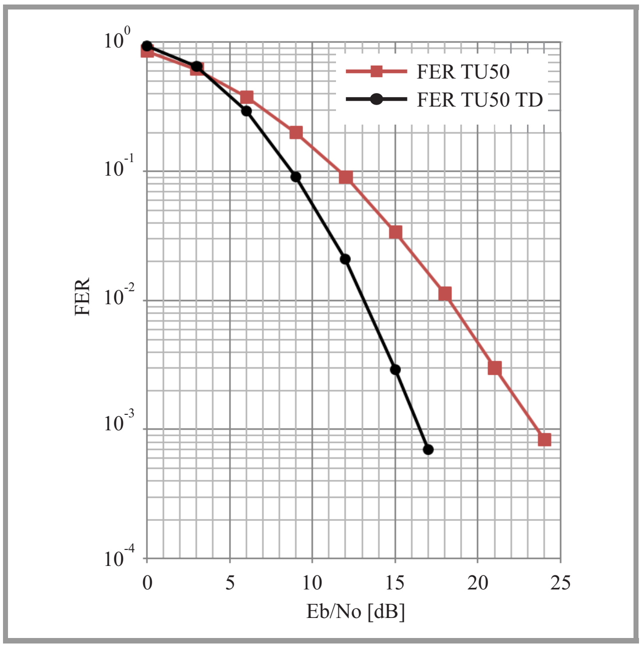


Fig. 4. Curves of FER versus Eb/No in the downlink over TU50 with 4QAM.

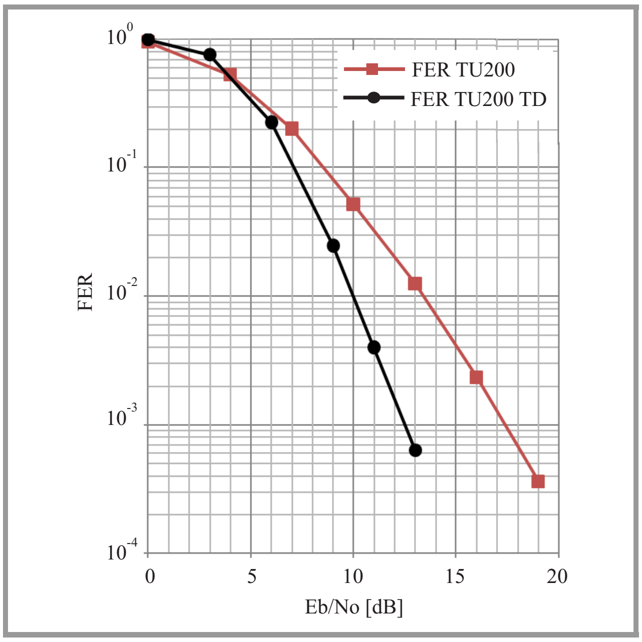


Fig. 5. Curves of FER versus Eb/No in the downlink over TU200 with 4QAM.

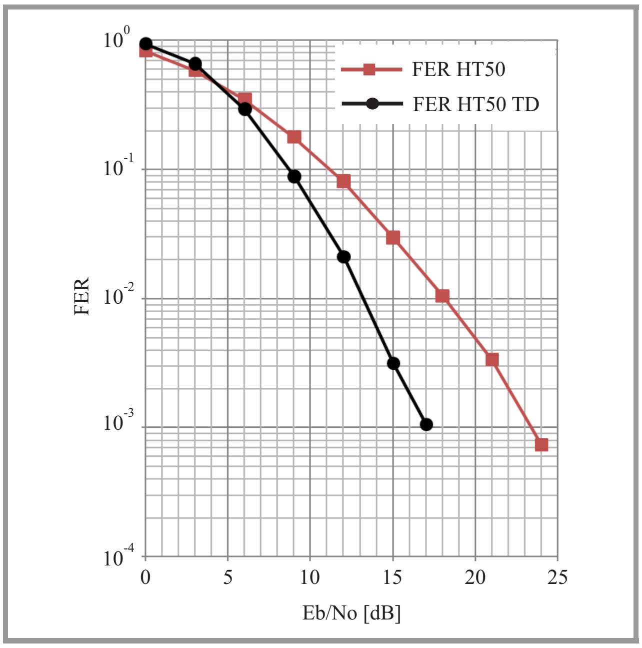


Fig. 6. Curves of FER versus Eb/No in the downlink over HT50 with 4QAM.

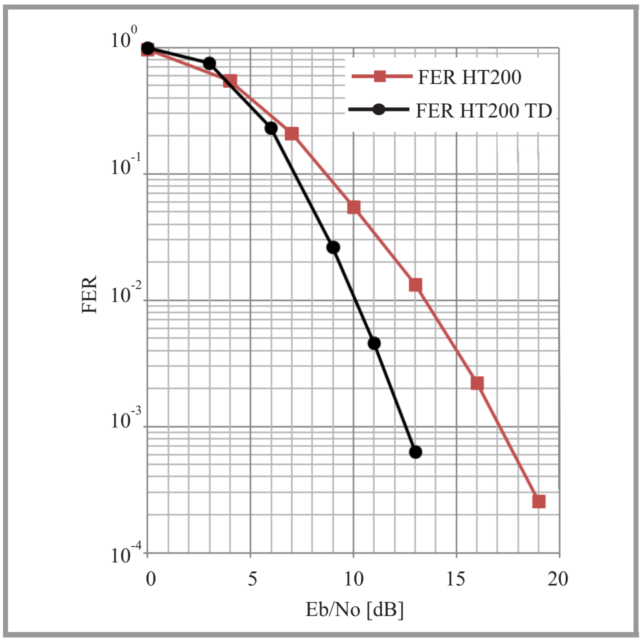


Fig. 7. Curves of FER versus Eb/No in the downlink over HT200 with 4QAM.

the proposed method provides a significant advantage. The Eb/No for the DISO configuration is reduced by 5–8 dB for FER = 10^{-3} as compared to the SISO. This transmit diversity gain is achieved irrespective of the propagation profile (TU or HT). Moreover, in both cases (with and without *TD*) one can notice that performance improves when terminal speed increases. This effect is obtained due to the decreased correlation time of fading, particularly as packet interleaving is not used. The results of the investigations also show that halving the number of symbols used for estimation of each channel characteristic with *TD*, provides sufficiently accurate channel characteristics. However, the transmit diversity gain was reduced to 5 dB for high terminal speed where the fluctuations in channel characteristics are greater. One of the reasons for this reduced gain is the reduced number of symbols used for the estimation of channel characteristics. Nevertheless, the proposed transmit diversity provides a significant improvement in system performance at the cost of an acceptable increase in complexity.

The approach presented in this paper can easily be employed for channels with other bandwidths of the TEDS radio interface.

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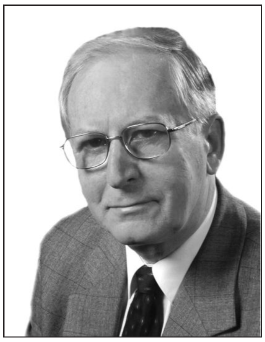

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