Standard Conformable Antenna Diversity Techniques for OFDM and its Application to the DVB-T System

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Abstract—**In this paper, we investigate different antenna diversity concepts, which can be easily applied to orthogonal frequency division multiplexing (OFDM) systems. The focus is to provide diversity schemes, which can be implemented to already existing OFDM systems without changing the standards. It is also possible to combine these techniques in a suitable manner. The introduced diversity techniques are applied to the DVB-T system for error performance investigations, which were done by simulation.**

I. Introduction

Future mobile wireless systems are required to provide high data rate services in a spectral efficient manner due to the high costs of bandwidth resources, e.g. $\approx 400,000,000$ €/MHz for UMTS in Germany. In terms of power efficiency — especially for mobiles — and electromagnetic pollution it is required to keep the isotropic radiated power as low as possible. Particularly the electromagnetic exposure becomes more and more important for the acceptance of wireless systems in society. Wireless systems have to operate in different environments. So a mobile is expected to work reliably in scenarios like rural, urban, indoor, outdoor, etc.

Mobile communication systems mainly suffer from time-varying multipath fading with extremely different multipath intensity profiles [1]. For systems, which have to work in multipath environments, an improvement in error performance may become very difficult. Already a slight improvement in the bit error rate can necessitate a huge amount of additional transmitter power, which contradicts the aforementioned item of an economically use of transmission power. It is an enormous challenge to design wireless communication systems, which are capable to deal with these varying scenarios.

Orthogonal frequency division multiplexing (OFDM) [2] is a suitable technique for broadband transmission in multipath fading environments and is implemented in some new broadcast standards like digital audio broadcasting (DAB) [3] or terrestrial digital video broadcasting (DVB-T) [4] as well as wireless local area network (WLAN) standards [5] such as HIPERLAN/2 or IEEE 802.11a.

Because of the poor error performance of OFDM in multipath environments, it is necessary for wireless communications systems to use techniques like interleaving and channel coding in addition to OFDM. These techniques add redundancy and diversity in time and frequency direction. For many scattering environments, spatial diversity is another effective way to improve the error performance of wireless radio systems. In [6], [7] space-time-coding is proposed in order to get the benefits of channel coding in combination with spatial (antenna) diversity. Unfortunately space-time-coding is not suitable for extending existing systems, because this would make non standard conformable modifications necessary. Therefore for standardized systems additional spatial diversity techniques can only be implemented, if this modifications keep the systems standard compatible. In [8] such techniques for the transmitter side are proposed.

In this paper we will investigate standard conformable antenna diversity techniques, which are well suited for the extension of existing standardized OFDM systems. Section II introduces diversity techniques for both transmitter and receiver. At this, the main idea is to increase the frequency selectivity of the resulting channel transfer function by specific cyclic delays at the transmitter and/or receiver antennas. The transmitter sided delay diversity is also investigated in combination with receiver sided maximum ratio combining. It is shown in Section III how the mentioned diversity techniques are applicable to the DVB-T system in order to improve the error performance in multipath environments. In Section IV the DVB-T system and transmission parameters as well as the used channel models are described. Finally, simulation results for the bit error rates are presented for the DVB-T system in indoor and outdoor environments.

II. Spatial Antenna Diversity with OFDM

In this section we will introduce *Cyclic Delay Diversity (CDD)* and *Maximum Ratio Combining (MRC)* as two antenna diversity techniques, which can easily be applied to existing OFDM system standards. A combination of this diversity techniques is easily possible. Thus, we can hindsight implement (scalable) systems with best tradeoff between complexity and performance.

A. Cyclic Delay Diversity (CDD)

In [8] several diversity techniques were investigated. This section will deal with cyclic delay diversity (CDD), which can be shown to be equal to phase diversity (PD) and conditionally equal to delay diversity (DD) as it is described in [8].

A.1 Transmitter Sided CDD

Fig. 1 shows the block diagram of an M-transmitter-antenna OFDM system with CDD. The OFDM modulated signal is transmitted over M antennas, whereas the particular signals

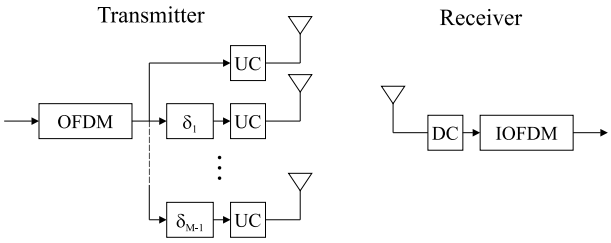

Fig. 1. OFDM system with transmitter sided cyclic delay diversity

only differ in an antenna specific cyclic shift. After cyclic shifting, an additional cyclic prefix as guard interval may be inserted. The functional blocks "UC" and "DC" stand for up-conversion resp. downconversion of the signals from the base-band into RF-band and vice versa. Note, that in case of CDD $\delta_\ell, \ell = 1, \ldots, M-1$, denote cyclic shifts.

The equivalence between PD and CDD is a property of the Discrete Fourier Transformation (DFT) and can directly be seen from the length N IDFT definition

$$s(t) = \frac{1}{\sqrt{N}} \sum_{f=0}^{N-1} S(f) \cdot e^{j\frac{2\pi}{N}ft} \quad (1)$$

$$\underbrace{s((t-\delta) \bmod N)}_{CDD\ signal} = \frac{1}{\sqrt{N}} \sum_{f=0}^{N-1} \underbrace{e^{-j\frac{2\pi}{N}f\delta} S(f)}_{PD\ signal} \cdot e^{j\frac{2\pi}{N}ft} \quad (2)$$

where t, f, $s(t)$ and $S(f)$ denote the discrete time, frequency and the complex-valued signals in time- and frequency-domain respectively with $t, f = 1 \ldots N-1$. δ stands for a cyclic time shift. Note, that δ actually can take any integer value, but due to the characteristics $(\delta + k \cdot N) \bmod N = \delta$ and $e^{j\frac{2\pi}{N}(\delta + k \cdot N)} = e^{j\frac{2\pi}{N}\delta}$, the range of values for δ can reasonably be restricted to $\delta = 0 \ldots N-1$.

As it can be seen from (2), the operation for PD has to be done *before* OFDM modulation. So for an M-antenna PD system, M OFDM transformations have to be done. Because of only one OFDM transformation in case of CDD, the implementation of CDD is much more efficient.

For DD Fig. 1 is also valid. In the sense of DD, $\delta_\ell, \ell = 1, \ldots, M-1$, mean a simple time shift of the original signal in samples. It is important to add a cyclic prefix *before* the delay operations. So CDD is equal to DD if

$$\delta_\ell \leq N_g - \tau_{max}, \quad \ell = 1, \ldots, M-1, \quad (3)$$

holds. Where δ_ℓ now denotes a simple time delay in samples. N_g and τ_{max} are the guard interval length and maximum channel delay in samples respectively. If that condition does not hold, intersymbol interference (ISI) occurs.

CDD and PD are independent of the existence of a cyclic prefix (guard interval) and do not increase the *overall* maximum channel delay spread in the sense of ISI occurrence because these operations are done *before* guard interval insertion

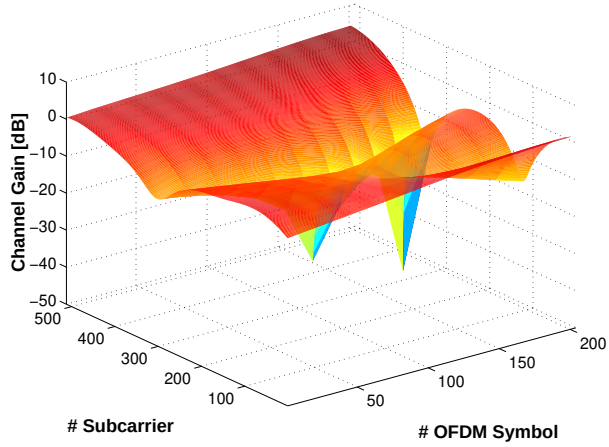

Fig. 2. Indoor channel snapshot for single antenna system

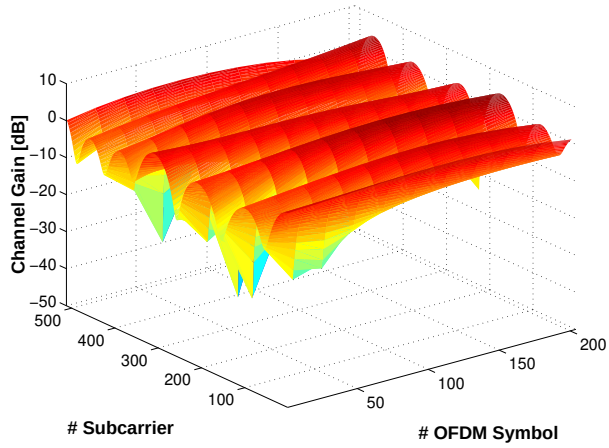

Fig. 3. Indoor channel snapshot for a CDD system

and are restricted to the OFDM symbol itself. But there is an increase of the channel delay spread *within* an OFDM symbol. The effect can be seen if we have a look on the overall channel transfer function

$$H_{res}(f,t) = \frac{1}{\sqrt{M}} \sum_{\ell=0}^{M-1} e^{-j\frac{2\pi}{N}f\delta_\ell} \cdot H_\ell(f,t), \quad (4)$$

where $H_\ell(f,t)$ denotes the channel transfer function from the ℓth transmitter antenna to the receiver antenna and δ_ℓ stands for the transmitter antenna specific (cyclic) delay ($\delta_0 = 0$). Fig. 2 shows a snapshot of $|H_{res}(f,t)|^2 = |H_0(f,t)|^2$ for a single antenna system over 512 adjacent subcarriers and the number of processed OFDM symbols.

In contrast, Fig. 3 shows a snapshot of $|H_{res}(f,t)|^2 = |H_0(f,t) + e^{-j\frac{2\pi}{N}f\delta_1} H_1(f,t)|^2$, $\delta_1 = 20$ for a 2-antenna-system in the same indoor environment.

For Fig. 2 and 3, an indoor channel model, which is described in Section IV-A, has been used. It can obviously

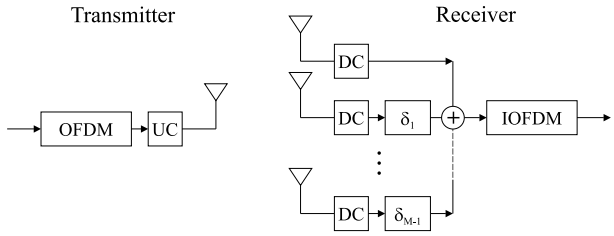

Fig. 4. OFDM system with receiver sided cyclic delay diversity

be seen, that an additional transmitter antenna with CDD increases the frequency selectivity, i.e. decreases the coherence bandwidth. In Section IV-C we will see, that a lower coherence bandwidth leads to a better error performance for coded OFDM systems in multipath environment.

In order to achieve any diversity effects, i.e. to get constructive and destructive interference within the OFDM signal bandwidth B, the delays δ_ℓ (in samples) have to fulfill

$$\delta_\ell \geq \frac{1}{B \cdot T_s}, \quad \ell = 1, \ldots, M-1, \tag{5}$$

where T_s denotes the sampling time of the OFDM time domain signal.

A.2 Receiver Sided CDD

Because of linearity, CDD, PD and DD can also be implemented at the receiver side. Fig. 4 shows the appropriate block diagram for CDD and DD. In case of CDD, first the guard interval is removed from each received signal, then the resulting signals are shifted cyclically at $\delta_\ell, \ell = 1, \ldots, M-1$, summarized and OFDM demodulated (inverse OFDM). For DD, the received signals first pass a simple time delay, denoted again as $\delta_\ell, \ell = 1, \ldots, M-1$, then the M signals are summarized. Finally the resulting signal is freed of the guard interval and OFDM demodulated. Because of the exact equivalence, all properties mentioned in Section II-A.1 are also valid for receiver sided CDD, DD and PD. Again the implementation of PD is more complex, because the diversity operation has to be done *after* the IOFDM transformation. So M IOFDM transformations have to be done in case of PD in contrast to only one transformation for CDD and DD.

B. Maximum Ratio Combining (MRC)

Fig. 5 shows the block diagram of a classical M-branch MRC-OFDM receiver. After downconversion into the baseband, the M time-signals are OFDM demodulated and combined by subcarriers, using the MRC scheme, which optimizes the signal-to-noise-ratio (SNR) for each subcarrier and is described e.g. in [1], [7].

C. Combined (CDD+MRC)

Because of the restriction of the above mentioned diversity techniques to either the transmitter or the receiver, it is easy

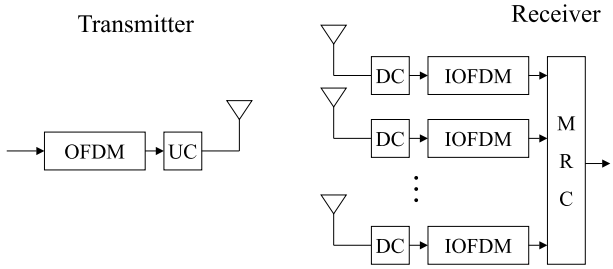

Fig. 5. OFDM system with receiver sided maximum ratio combining

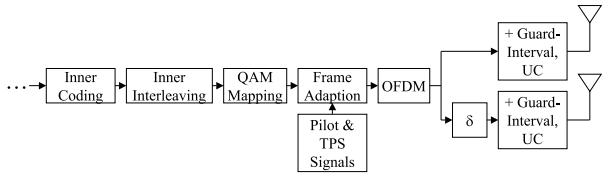

Fig. 6. Inner non hierarchical DVB-T transmitter part with CDD

to combine these techniques. A suitable combination would be transmitter sided CDD and receiver sided MRC. A block diagram can be obtained by taking the transmitter part of Fig. 1 and the receiver part of Fig. 5.

III. Application to the DVB-T System

In this section we will apply CDD and MRC, introduced in Section II, to the DVB-T system. DVB-T is basically a coded OFDM system containing an outer shortened Reed-Solomon code concatenated with an inner (punctured) convolutional code [4]. The investigations in this paper restrict to the "inner system" of DVB-T for non hierarchical transmission parameter sets, i.e. outer coding and decoding are not considered.

A. Transmitter

For the implementation of CDD at the DVB-T transmitter, only a second signal path after OFDM-modulation have to be added. Fig. 6 shows the transmitter sided inner part with transmitter CDD. After channel coding and interleaving, the bit-stream is mapped to complex-valued QAM-symbols. The functional block "Frame Adaption" is responsible for QAM-symbol interleaving, pilot insertion and transmission parameter signaling (TPS). The resulting symbol-stream is OFDM-modulated. Finally the signal is splitted, upconverted and transmitted directly on the one hand and cyclic shifted on the other hand. It is important to note that signal splitting does not increase the overall transmission power.

B. Receiver

As depicted in Section II-A, CDD may also be implemented at the receiver side. The corresponding block diagram for a non hierarchical DVB-T CDD-receiver is shown in Fig. 7. Because of the equivalence to transmitter-CDD, receiver-CDD is not investigated separately.

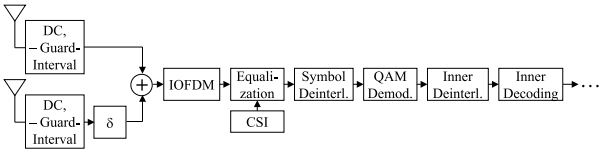

Fig. 7. Non hierarchical DVB-T CDD-receiver

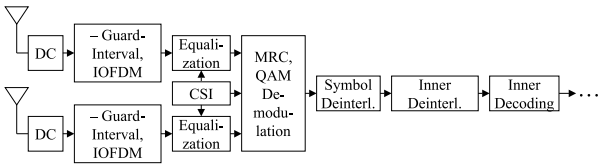

Fig. 8. Non hierarchical DVB-T MRC-receiver

Another well known standard conformable diversity technique is MRC, which may also work in combination with transmitter-CDD. A block diagram of the implementation, which is used for simulations is illustrated in Fig. 8. After downconversion and guard interval removal, the received signal is OFDM-demodulated and equalized using zero forcing. For our investigations we assume perfect knowledge of the channel state information (CSI). Both complex-valued symbol-streams are combined, using MRC, and QAM demodulated with soft-out values before symbol- and bit-deinterleaving is done. Finally the bit stream is soft-decision-maximum-likelihood (SDML) decoded.

IV. Simulations

In this section we will present simulation results for DVB-T with antenna diversity in indoor and outdoor scenarios. The simulations were done for Doppler frequencies, which follow from typical pedestrian movement in indoor environment and mobile vehicle movement in outdoor parking-lot scenarios.

A. Channel Models

Table I shows the main properties of the wide sense stationary uncorrelated scattering (WSSUS) channel models, which were used for simulations. For the individual scatterers Rayleigh fading is assumed. The mobile radio channel models were described in [9]. For the indoor scenario the "Indoor Commercial" (Channel B) channel model is used with a maximum Doppler frequency of $10Hz$, which correspond to a mobile velocity of $1.25\frac{m}{s}$ ($4.5\frac{km}{h}$) for operation in the $2.4GHz$ band. For the outdoor scenario the "Outdoor Residential – High Antenna" (Channel B) channel model is used. The maximum Doppler frequency for the outdoor scenario is chosen to be $50Hz$, which results in a mobile velocity of $6.25\frac{m}{s}$ ($22.5\frac{km}{h}$). Antenna diversity simulations were done for pairwise uncorrelated WSSUS channels $H_{k\ell}(f,t)$, where $H_{k\ell}(f,t)$ denote the propagation channel transfer function from transmitter antenna k to receiver antenna ℓ.

TABLE I
MAIN CHANNEL PROPERTIES

	indoor	outdoor
Number of Taps	7	10
max. Channel Delay τ_{max}	$750ns$	$15\mu s$
max. Doppler Frequency f_{Dmax}	$10Hz$	$50Hz$
Doppler Spectrum Form	rectangular	Jakes

B. DVB-T System Parameters

For $8MHz$ channels the DVB-T standard defines some basic parameters. In 2k-mode the OFDM symbol duration is $T_U = 224\mu s$. The number of (used) carriers is $K = 1705$. This yields a subcarrier spacing of $1/T_U = 4464Hz$ and a spacing between the spectrum edge carriers of $(K-1)/T_U = 7.61MHz$. For the 8k-mode, the basic parameters are $T_U = 896\mu s$ and $K = 6817$. Subcarrier spacing and edge carrier spacing will follow equivalently. Beside this basic parameters, there are some additional transmission parameters regarding modulation, guard interval length and channel coding. For outdoor environment simulations a guard interval length of $\tau_g = 1/8 \cdot 224\mu s = 28\mu s$ have to be chosen because of the maximum channel delay of $\tau_{max} = 15\mu s$. Table II shows the DVB-T parameters, which were chosen for simulations.

TABLE II
DVB-T SYSTEM PARAMETERS FOR SIMULATIONS

Code Rate	1/2		
Signal Constellation	4-QAM	16-QAM	64-QAM
DVB-T Mode	2k		
Hierarchy (α)	Non hierarchical		
Guard Interval	1/8		

C. Results

The Monte Carlo simulations are restricted to the "inner" DVB-T system as it is described in Section III. It is important to note, that the overall transmitted power is kept equal for all simulation runs, i.e., the transmitted power per transmit antenna decreases with increasing number of antennas. For simulations we will restrict to the use of at most 2 antennas at the transmitter respectively the receiver side. Fig. 9 shows the bit error rate vs. the signal-to-noise ratio (SNR) for the indoor channel with different diversity techniques, applied to the DVB-T system in non hierarchical 2k-mode with 4-QAM-modulation and code rate 1/2. Note that the SNR here equals E_s/N_0 (the signal energy divided by noise power spectral density). A single-antenna-system is given as a reference, i.e., for this system no spatial diversity is implemented. As it can be seen from Fig. 9, the receiver-MRC system outperforms the single-antenna system about 7.5 dB in SNR at a BER of $2 \cdot 10^{-4}$, because of additional signal power and spatial diversity at the receiver. The two propagation paths (channel

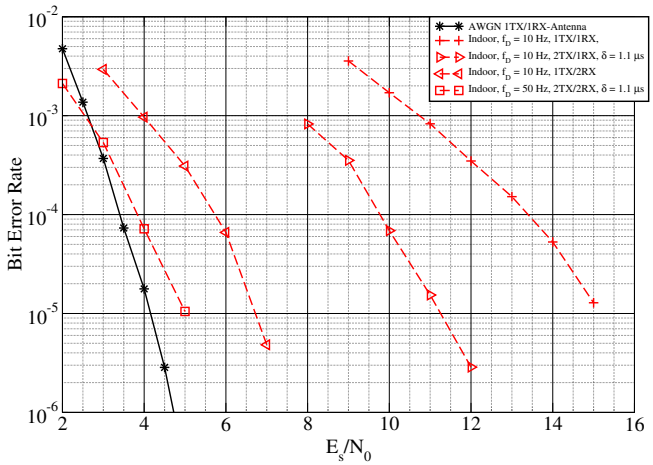

Fig. 9. BER vs. SNR for indoor environment, 2k-mode, 4-QAM, rate 1/2

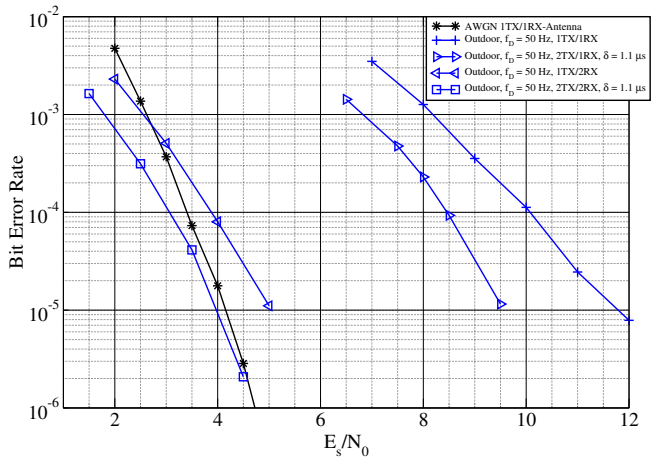

Fig. 10. BER vs. SNR for outdoor environment, 2k-mode, 4-QAM, rate 1/2

transfer functions) are assumed to be uncorrelated. So subcarriers, which are in a deep fade for receiver-antenna 1 may have good channel properties for antenna 2. With additional transmitter-CDD, some further gain can be achieved. Note, that the transmitter-CDD gain decreases with the use of MRC. As mentioned in Section 1, CDD increases channel selectivity and therefore decreases the occurrence of packet errors after demodulation. This packet errors may appear also before decoding in spite of interleaving due to extremely wide deep fades, which particularly occur in environments with small channel delay spreads, e.g. indoor scenarios.

Fig. 10 shows the bit error performance for a DVB-T system in non hierarchical 2k-mode with 4-QAM-modulation and code rate 1/2, which now operates in an outdoor environment. Again most can be gained, if MRC is used at the receiver. In general the error performance is better for all system combinations compared to the indoor scenario. Note, that the outdoor channel provides a higher maximum channel delay and therefore a higher frequency selectivity, which explains the better error performance compared to the indoor channel. It is also interesting to see, that the gains *between* the different system combinations for the outdoor channel become smaller compared to indoor channels. So the lower the maximum channel delay τ_{max} the higher is the achievable gain for additional use of CDD.

Fig. 11 shows the transmitter-CDD gain for single-antenna and 2-antenna MRC-receivers in indoor and outdoor environments vs. the cyclic delay at the 2nd transmit antenna of a 2-antenna CDD transmitter. As the results, showed in Figs. 9 and 10, signify, there is a saturation effect in terms of the cyclic diversity delay. It can be observed, that a cyclic delay of $\delta > 1.5\mu s$ results in no further improvement. Fig. 11 also shows, that the achievable gain is much higher for indoor scenarios compared to outdoor scenarios due to the extremely different maximum channel delays of $\tau_{max} = 15\mu s$ for outdoor and $\tau_{max} = 750ns$ for indoor scenarios. Different from

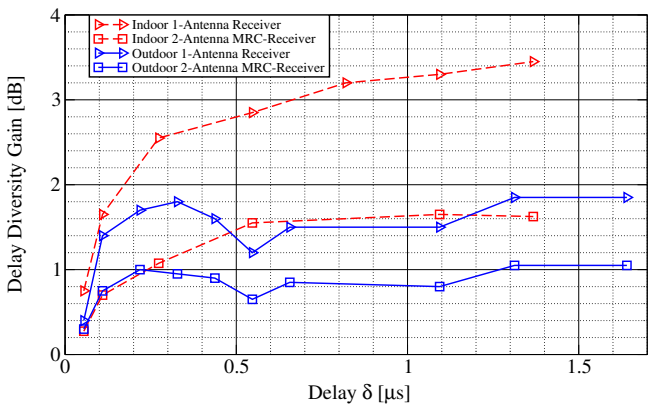

Fig. 11. Cyclic Delay Diversity Gain vs. Delay at $BER = 2 \cdot 10^{-4}$, 2k-mode, 4-QAM, rate 1/2

the indoor results, the outdoor delay diversity gain shows local minima at about $\delta = 0.55\mu s$. A reason is the structure of tap delay model for the outdoor channel. For $\delta = 0.55\mu s$, relatively poweful paths of the outdoor channel model come close to each other. This tight adjacent echos yield a reduced frequency selectivity of the overall channel transfer function $|H_{res}(f,t)|^2$.

Figs. 12-15 show the results for 16-QAM resp. 64-QAM modulation. As one can see the achievable delay diversity gains nearly remain the same as for 4-QAM. Again, local minima at $\delta = 0.55\mu s$ can be observed, which results from the outdoor channel model properties.

V. Conclusions

In this paper cyclic delay diversity (CDD) and maximum ratio combining have been presented. The (conditional) equivalence between CDD and phase diversity respectively delay diversity have been shown. It is the purpose of CDD to increase

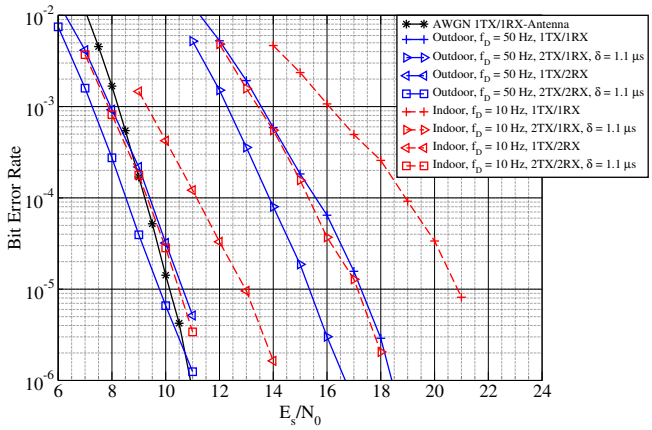

Fig. 12. BER vs. SNR, 2k-mode, 16-QAM, rate 1/2

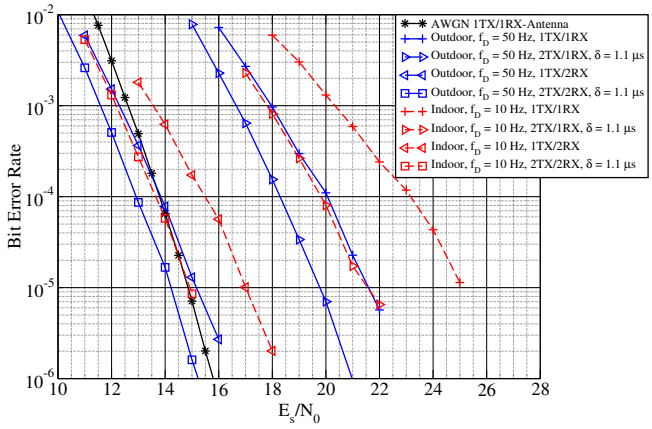

Fig. 14. BER vs. SNR for, 2k-mode, 64-QAM, rate 1/2

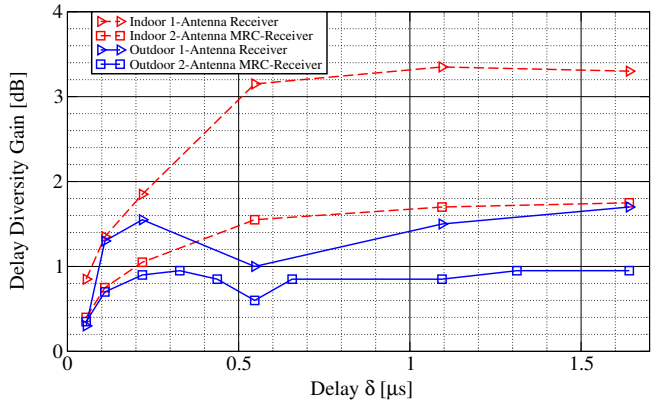

Fig. 13. Cyclic Delay Diversity Gain vs. Delay at $BER = 2 \cdot 10^{-4}$, 2k-mode, 16-QAM, rate 1/2

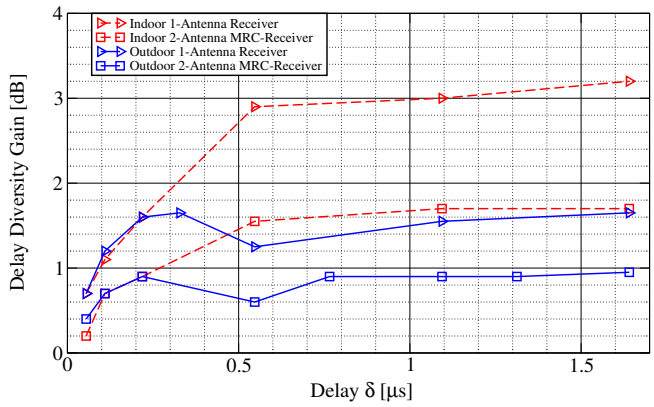

Fig. 15. Cyclic Delay Diversity Gain vs. Delay at $BER = 2 \cdot 10^{-4}$, 2k-mode, 64-QAM, rate 1/2

the frequency selectivity of the resulting channel transfer function by specific cyclic delays at the transmitter and/or receiver antennas. The presented antenna diversity schemes can easily be implemented in existing OFDM systems without changing the standards or the receivers. It was shown exemplarily, how this techniques are applicable to the DVB-T system. Simulations showed, that about $9dB$ in SNR can be gained for a DVB-T system with an 2-antenna CDD-transmitter and 2-antenna MRC-receiver compared to a single-antenna system in indoor environment.

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