Analysis of Cyclic Delay Diversity on DVB-H Systems over Spatially Correlated Channel

Y. Zhang, J. Cosmas, K.-K. Loo, M. Bard, and R. D. Bari

Abstract—**The objective of this work is to research and analyse the performance of Cyclic Delay Diversity (CDD) with two transmit antenna on DVB-H systems operating in spatially correlated channel. It is shown in this paper that CDD can achieve desirable transmit diversity gain over uncorrelated channel with or without receiver diversity. However, in reality, the respective signal paths between spatially separated antennas and the mobile receiver is likely to be correlated because of insufficient antenna separation at the transmitter and the lack of scattering effect of the channel. Under this spatially correlated channel, it is apparent that CDD cannot achieve the same diversity gain as obtained under the uncorrelated channel. In this paper, a new upper bound on the pairwise error probability (PEP) of the CDD with spatial correlation of two transmit antennas is derived. The upper bound is used to study the CDD theoretical error performance and diversity gain losses over a generalized spatially correlated Rayleigh channel. This theoretical analysis is validated by the simulation of DVB-H systems with two transmit antennas and the CDD scheme. Both the theoretical and simulated results give the valuable insight that the CDD ability to perform well with a certain amount of channel correlation.**

Index Terms—**Cyclic delay diversity (CDD), DVB-H, pairwise error probability (PEP), spatial correlation, transmit diversity.**

I. Introduction

DVB-H [1] is the latest development from the digital video broadcasting (DVB) [2] standard; it is on its way to provide digital multimedia broadcasting to mobile devices. Receiving devices are expected to be able to move freely while receiving the transmission at high data rates in a medium size single frequency networks (SFN). Under such adverse propagation channels between the mobile and stationary units, the DVB-H signals are prone to multipath fading in conjunction with the Doppler spread. Although the DVB-H standard is suitable for mobile reception, the error performance is solely dependant on the receiver, in particular its ability to recover a signal with poor SNR and its robustness against adjacent channel interference.

Recently, the use of multiple antennas and transmit diversity techniques have been proposed to improve error performance and capacity of wireless systems. This scheme exploits the scattering effect of the channel by mean of transmitting multiple signals in a controlled manner from different antennas and allows independently faded signals to be detected at the receiver with or without receive diversity. To date, transmit diversity is a commonly used technique for most digital mobile communications whereas it has so far not been applied in DVB systems. Much of the recent work has focused on techniques such as space-time block code and space-time trellis codes [3]–[5] which were the proposal for combining temporal and spatial diversity. Unfortunately, these techniques are not compliant to the current DVB standard i.e. DVB-T/H. Instead a simple yet elegant method known as Cyclic Delay Diversity (CDD) [6]–[8] is complaint and is applicable to DVB-T/H without modifications to its existing physical layer mainly because its signal processing is performed on the baseband OFDM [9] symbol in time-domain.

Using the CDD transmit diversity scheme, the transmit signals are cyclically shifted between respective antennas without a noticeable time-domain delay spread at the receiver. This scheme has the effect of randomizing the channel frequency response by increasing the frequency selectivity of the resulting channel transfer function at the receiver; thus reduces the likelihood of deep fading. It is shown in this paper that CDD can achieve desirable transmit diversity gain over uncorrelated channel with or without receiver diversity. However, in reality, the respective signal paths between spatially separated antennas and the mobile receiver is likely to be correlated to a certain degree because of insufficient antenna separation at the transmitter and the lack of scattering effect of the channel. This correlation results in diversity loss and performance degradation of CDD. Thus, it is apparent that CDD cannot achieve the same diversity gain as obtained under the uncorrelated channel. Performance-bound analysis has been widely adopted to study the error performance of wireless systems.

In this paper, a new upper bound on the PEP of CDD with spatial correlation of two antennas is derived. The upper bound is used to study the error performance of CDD in the existence of spatially correlated Rayleigh channel. This theoretical analysis is validated by the simulation of DVB-H systems with two transmit antennas and the CDD transmit diversity scheme. Both the theoretical and simulated results give the valuable insight that the CDD is able to obtain diversity gain to a certain extent under spatially correlated channel. The results also suggested a significant improvement in the performance of digital broadcasting systems [10], [11] in particular for mobile receiver in non-line of sight environment when CDD is applied.

This paper is organized as follows: Section II introduces the system description, CDD extension on OFDM system and

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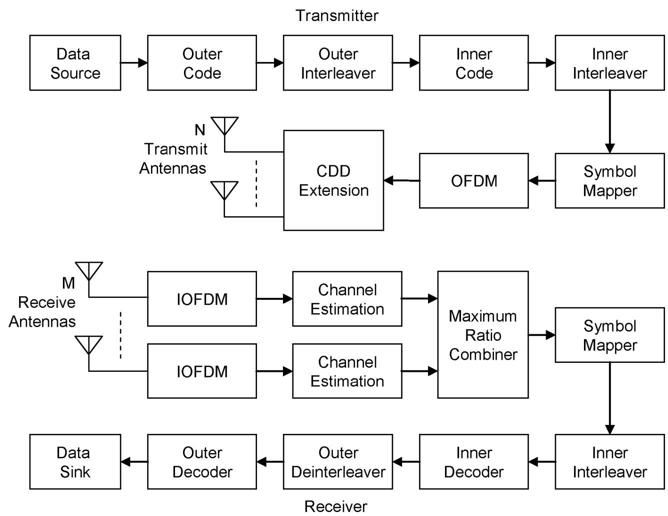

Fig. 1. System model of DVB-H with CDD extension.

MIMO channel modeling. Section III presents upper bound derivation for CDD over correlated channel. Section IV provides simulation results and discussion of DVB-H with CDD over uncorrelated and spatially correlated mobile channels. Finally, Section V concludes the findings of this paper.

II. System Description

Fig. 1 depicts the system model of DVB-H with CDD extension which is used for the BER performance simulation over the uncorrelated and spatially correlated mobile channels (Rayleigh with Doppler effect).

There is no different between DVB-T and DVB-H except additional configurations on the existing physical layer to support mobility. In DVB-H, the data source should go through an additional outer layer of forward error correction (FEC) coding called MPE-FEC (multi-protocol encapsulation forward error correction), but it was not considered in the simulation as it is not inline with the focus of this paper. Instead the data source goes through an outer Reed-Solomon code, an outer convolutional interleaver, an inner convolutional code an inner interleaver. The output of the inner convolutional code can be punctured to adjust different code rates $\mathrm{R} = 1/2$, 2/3, 3/4, 5/6 or 7/8. The DVB-H uses 4 K subcarriers ideally allow good reception of broadcast service on mobile devices in a medium size SFN. The subcarriers are modulated with QPSK, 16-QAM or 64-QAM by the OFDM followed by the guard interval insertion where the length can be 1/4, 1/8, 1/16 and 1/32. In CDD extension, the OFDM modulated symbols are cyclically shifted between respective N transmit antennas with equally distributed transmit power. At each receiver antenna, the received signals are demodulated by the inverse OFDM (IOFDM) followed by the channel estimation which is performed by minimum mean square error (MMSE) detector. For the simulation of 2×2 MIMO antennas, receiver diversity with maximum ratio combining (MRC) is applied. Thereafter, the rest of the baseband processing is the reverse operation of the transmitter.

A. Cyclic Delay Diversity Extension

CDD is a complaint diversity scheme to the existing DVB-H standard mainly because its signal processing is performed on the OFDM output symbols in time-domain. Suppose, the symbols are denoted $s(t), t = 0, \ldots, K-1$ where K is number of subcarriers. During transmission, these symbols are cyclically shifted between respective N transmit antennas. The symbols transmitted from antenna i^{th} at time t can be represented by:

$$s_i(t) = s_i\left((t - \delta^{cyc}) \mathrm{mod\ K}\right), \quad t = 0, \ldots, K-1,\ i = 0, \ldots, N-1 \tag{1}$$

where δ^{cyc} is the cyclic delay introduced between the symbols. The cyclically shifted symbols can be recorded in $N \times K$ matrix as codeword C:

$$C = \begin{pmatrix} s(0) & s(1) & \ldots & s(K-1) \\ s(K-1) & s(0) & \ldots & S(K-2) \\ \ldots & \ldots & \ldots & \ldots \\ s(K-N+1) & s(K-N+2) & \ldots & s(K-N) \end{pmatrix}_{N \times K} \tag{2}$$

which is used in Section III for the derivation of the CDD upper bound. From (2), it can be seen that i^{th} antenna is actually transmitting $s(0), s(1), \ldots, s(K-1)$ sequence that is cyclically shifted by i times seen as $s(K-i), s(K-i+1), \ldots, s(K-i-1)$. Nevertheless, the system is equivalent to the transmission of the sequence $s(0), s(1), \ldots, s(K-1)$ over a propagation channel with one transmit antenna to receiver $j, j = 1, \ldots, M$

$$h_{1j}^{e,t}(d) = \sum_{i=1}^{N} h_{ij}^{t}((d - \delta^{cyc}) \mathrm{mod\ K}), \quad j = 1, \ldots, M \tag{3}$$

and the CIR can be described as

$$h_{1,j}^{e,t} = \left[h_{1,j}^{e,t}(0), \ldots, h_{1,j}^{e,t}(K-1)\right]. \tag{4}$$

In the frequency domain, the equivalent channel transfer function is expressed as

$$H_m^e(k) = \sqrt{\frac{SNR}{N}} \sum_{n=1}^{N} e^{-j\frac{2\pi}{K}k\delta_n^{cyc}} \cdot H_{n,m}(k) \tag{5}$$

where $H_{n,m}(k)$ denotes the channel frequency response of the transmission from the n^{th} transmit antenna to the m^{th} receive antenna and δ_n^{cyc} is cyclic delay of the n^{th} transmit antenna (there is no cyclic delay at 1st transmit antenna, $\delta_1^{cyc} = 0$). Fig. 2 shows the channel frequency response of a system with and without CDD. It is obvious that CDD has transformed the spatial diversity into frequency diversity; in other words the MIMO channel is transformed into a single-input multiple-output (SIMO) with increased frequency-selectivity.

B. MIMO Channel Model

A generic multiple-input multiple-output (MIMO) channel with N transmit and M receive antennas shown in Fig. 3 is considered. The independently distributed Wide Sense Stationary Uncorrelated Scattered (WSSUS) [12] channel model is used as the basis in modeling the MIMO channel where each signal

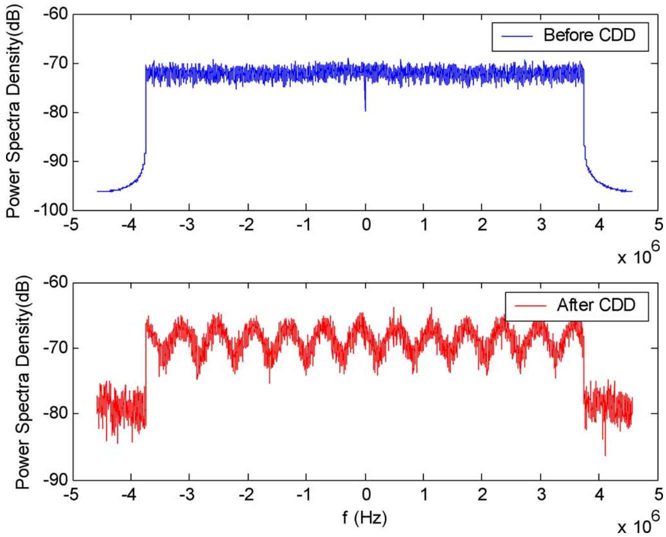

Fig. 2. Power spectra density of signals before CDD and after CDD UHF (900 MHz).

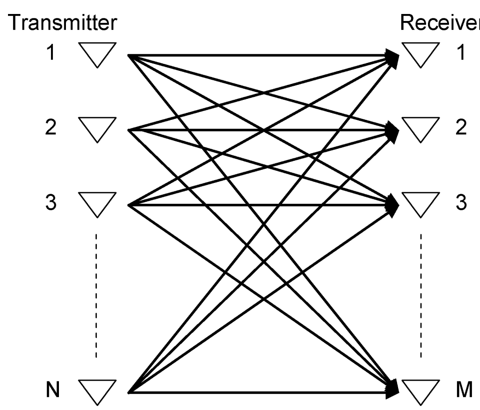

Fig. 3. M × N i.i.d MIMO channel.

path consists of a zero mean complex-valued Gaussian process in which the envelope is Rayleigh distributed. If the process does not have a zero mean, the envelope is then Rician distributed.

Suppose the channel is composed of D echoes. The instantaneous CIR function can be expressed as:

$$h(t,\tau) = \sqrt{1-\rho^2}\delta(\tau) + \frac{\rho}{\sqrt{D}}\sum_{n=1}^{D} e^{j\theta_n} e^{j2\pi fD_n t}\delta(\tau-\tau_n) \tag{6}$$

where

D	Number of realizations (echoes)
θ_n	Random phases
fD_n	Random Doppler frequencies
τ_n	Random delays
ρ	Fading amplitude

Consider a discrete baseband signal processing, at sample index t, the complex symbol, $s(t)$ sent by N transmit antennas and detected by k^{th} receive antenna is denoted as $r_k(t)$. The fading coefficient h_{ij} for each channel is the complex path gain from the transmit antenna j to the receive antenna i. h_{ij} is independent complex circular symmetric Gaussian according to

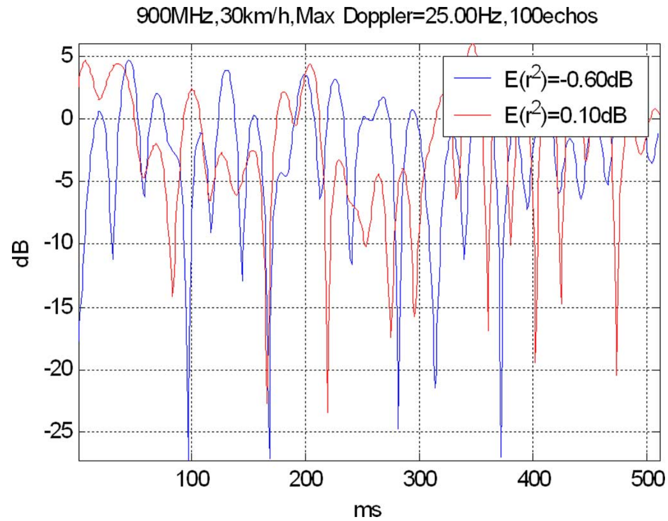

Fig. 4. Time domain of 2 uncorrelated Rayleigh fading channels.

(6). h_{ij} is assumed to be known to the receiver, thus the receive signal in time-domain can be represented as

$$r_k(t) = \sqrt{\frac{SNR}{N}}\sum_{i=1}^{N} h_{ki}(t)\otimes s_i(t) + v_k(t) \tag{7}$$

where SNR is the average SNR at each receive antenna, $v_k(t)$ is the complex zero-mean spatially and temporally additive white Gaussian noise (AWGN) with variance $N_0/2$ per dimension. In frequency domain, the receive signal is represented as:

$$R_k(m) = \sqrt{\frac{SNR}{N}}\sum_{i=1}^{N} H_{ki}(m)S_i(m) + V_k \tag{8}$$

where $R_k(m)$, $H_{ki}(m)$, $S_i(m)$, V_k denote the frequency domain representations of m^{th} subcarrier of the received signal for the k^{th} antenna, complex channel gains between i^{th} antenna, and noise signal for the k^{th} receive antenna, respectively. Finally, the MIMO channel model is expressed as

$$\tilde{R}(m) = \sqrt{\frac{SNR}{N}}\tilde{H}(m)\tilde{S}(m) + \tilde{V}(m) \tag{9}$$

where

$$\begin{aligned}
\tilde{R}(m) &= [R_1(m), R_2(m), \ldots, R_M(m)]^T,\\
\tilde{S}(m) &= [S_1(m), S_2(m), \ldots, S_M(m)]^T,\\
\tilde{H}(m) &= \begin{bmatrix} H_{11}(m) & H_{12}(m) & \ldots & \ldots & H_{1N}(m)\\ H_{21}(m) & \ldots & \ldots & \ldots & H_{2N}(m)\\ \ldots & \ldots & \ldots & \ldots & \ldots\\ \ldots & \ldots & \ldots & \ldots & \ldots\\ H_{M1}(m) & \ldots & \ldots & \ldots & H_{MN}(m) \end{bmatrix}
\end{aligned} \tag{10}$$

For the uncorrelated MIMO channel, it is assumed the channel is sufficiently spatially separated to ensure decorrelation of the multipath signals. The benefits of MIMO transmit diversity are realized when the fading on received signals from the two channels are decorrelated, that is, the signals are faded independently as shown in Fig. 4.

III. Cyclic Delay Diversity Over Correlated Channel

This section derives the pairwise error probability (PEP) in order to obtain the upper bound to study the worst-case error performance and the diversity gain losses of the CDD over generalized spatially correlated Rayleigh channel. Before discussing the PEP derivation, the spatially correlated channel is derived using the uncorrelated mobile channel in (10) and the correlation of two transmit antennas as follows [13]:

$$H = (h_1\, h_2\, \ldots h_N), \quad H = H_G S^{\frac{1}{2}}, \quad S = S^{\frac{1}{2}} S^{\frac{H}{2}}$$
$$S = \begin{bmatrix} 1 & \rho_{12} & \cdots & \rho_{1N} \\ \rho_{21} & 1 & \cdots & \rho_{2N} \\ \cdots & \cdots & \cdots & \cdots \\ \rho_{N1} & \rho_{N2} & \cdots & 1 \end{bmatrix} \tag{11}$$

where $H_G = [w_1\, w_2\, \ldots w_N]$ is N independent Rayleigh fading channel matrix, S is the transmit correlation matrix, and ρ_{ij} is the correlation coefficient between the i^{th} and the j^{th} transmit antennas.

Consider all dimensions of the signal that are expressed in codeword C, as in (2), to be transmitted, with maximum likelihood detection at the receiver, the probability of an erroneous codeword e having transmitted C conditioned on a spatially correlated channel can be expressed in closed form using the Gaussian error function:

$$P(C \to e) = Q\left(\sqrt{\frac{d^2(c,e;H)}{2\sigma^2}}\right), \tag{12}$$

where σ^2 is the variance of the complex-valued noise process, H is the spatially correlated channel and $d^2(c,e;H)$ is the distance of the two transmit signals at the receiver, which is given by

$$d^2(c,e;H) = \sum_{t=0}^{K-1} \left\| H(c_l - e_l) \right\|^2 \tag{13}$$

Assuming perfect channel state information (CSI) at the receive antenna, the probability $P(C \to e)$ in Gaussian error function (12) can be tightly upper bounded by using the Chernoff bound, thus yields [14]:

$$P(C \to e) \le \exp\left(-\frac{E_s}{4N_0} d^2(c,e;H)\right) \tag{14}$$

where the mean value will be give an upper bound for the ergodic PEP. In order to arrive to a meaningful expression for this upper bound, (13) is further developed into

$$\begin{aligned} d^2(c,e;H) &= \sum_{t=0}^{K-1} \left| \sum_{i=1}^{N} h_i (c_l^i - e_l^i) \right|^2 \\ &= \sum_{t=0}^{K-1} \left(\sum_{i=1}^{N} h_i (c_l^i - e_l^i) \right) \left(\sum_{j=1}^{N} \overset{*}{h}_j (\overset{*}{c}{}_l^j - \overset{*}{e}{}_l^j) \right) \\ &= (h_1\, h_2\, \ldots h_N)(c-e)_{NK} \\ &\quad \times (c-e)_{NK}^{\dagger} (h_1\, h_2\, \ldots h_N)^{\dagger} \\ &= (h_1\, h_2\, \ldots h_N) \cdot A \cdot (h_1\, h_2\, \ldots h_N)^{\dagger} \end{aligned} \tag{15}$$

where $(.)^{\dagger}$ is conjugate transposition, $(.)^*$ is element wise conjugate, h is the spatially correlated channel and $A = (C-e)(C-e)^{\dagger}$ is the Hermitian matrix. According to the characteristics of the Hermitian matrix, there must be a unitary matrix U and a diagonal matrix D to get $U \cdot A \cdot U^{\dagger} = D$. The diagonal elements of D are the corresponding eigenvalues, λ_i, of A. Applying these properties into (15) yields:

$$\begin{aligned} d^2(c,e;H) &= (h_1\, h_2 \ldots h_N) \cdot (U^{\dagger} \cdot D \cdot U) \cdot (h_1\, h_2 \ldots h_N)^{\dagger} \\ &= (h_1\, h_2 \ldots h_N) U^{\dagger} \begin{bmatrix} \lambda_1 & 0 & \cdots & 0 \\ 0 & \lambda_2 & \cdots & 0 \\ 0 & \cdots & \cdots & 0 \\ 0 & 0 & \cdots & \lambda_N \end{bmatrix} \\ &\quad \times U (h_1\, h_2 \ldots h_N)^{\dagger}. \end{aligned} \tag{16}$$

where according to (11), the $(h_1\, h_2\, \ldots h_N)U^{\dagger}$ can be represented by $(w_1, w_2, \ldots, w_N) S^{1/2} U^{\dagger} = (\xi_1, \xi_2, \ldots, \xi_N)$, thus (16) can be expressed as

$$\begin{aligned} d^2(c,e;H) &= (\xi_1\, \xi_2\, \ldots \xi_N) \begin{bmatrix} \lambda_1 & 0 & \cdots & 0 \\ 0 & \lambda_2 & \cdots & 0 \\ 0 & \cdots & \cdots & 0 \\ 0 & 0 & \cdots & \lambda_N \end{bmatrix} \\ &\quad \times (\xi_1\, \xi_2\, \ldots \xi_N)^{\dagger} \\ &= \sum_{i=1}^{N} \lambda_i \xi_i \xi_i^{\dagger} \end{aligned} \tag{17}$$

Since U is unitary matrix where $\{u_1, u_2, \ldots, u_N\}$ is an orthonormal basis vector, $\xi_i \xi_i^{\dagger}$ can be derived into:

$$\begin{aligned} \xi_i \xi_i^{\dagger} &= (h_1, h_2, \ldots, h_N) \cdot u_i \cdot u_i^{\dagger} \cdot (h_1, h_2, \ldots, h_N)^{\dagger} \\ &= (w_1, w_2, \ldots, w_N) \cdot S^{1/2} \cdot S^{\dagger/2} (w_1, w_2, \ldots, w_N)^{\dagger} \\ &= (w_1, w_2, \ldots, w_N) \cdot S \cdot (w_1, w_2, \ldots, w_N)^{\dagger} \end{aligned} \tag{18}$$

Applying (18) into (17), the codeword distance is expressed as:

$$\begin{aligned} d^2(c,e;H) &= \sum_{i=1}^{N} \lambda_i \xi_i \xi_i^{\dagger} \\ &= \sum_{i=1}^{N} \lambda_i (w_1, w_2, \ldots, w_N) \\ &\quad \cdot S \cdot (w_1, w_2, \ldots, w_N)^{\dagger} \end{aligned} \tag{19}$$

where w_i is the Rayleigh fading factor and $E(w_i) = 0$. Applying (19) into (14), the PEP is expressed as:

$$\begin{aligned} P(C \to e) &\le \exp\left(-\frac{E_s}{4N_0} d^2(c,e;H)\right) \\ &\le \exp\left(-\frac{E_s}{4N_0} (\lambda_1 w_1, \lambda_2 w_2, \ldots, \lambda_N w_N) \right. \\ &\qquad \left. \times S^T (w_1, w_2, \ldots, w_N)^{\dagger}\right) \\ &\le \prod_{i=1}^{N} \left(\frac{E_s}{4N_0} \lambda_i\right)^{-1} \prod_{j=1}^{N} \lambda_j^{-1}(S) \\ &\le \left(\frac{E_s}{4N_0}\right)^{-\gamma} \prod_{i=1}^{\gamma} \lambda_i^{-1}(A) \prod_{j=1}^{\gamma} \lambda_j^{-1}(S) \end{aligned} \tag{20}$$

where λ_i are the eigenvalues of A, λ_j are the eigenvalues of A, S is the channel covariance matrix, and $\gamma = rank(S) = rank(A)$. This expression provides a figure of merit for CDD

TABLE I
POWER DELAY PROFILE OF TYPICAL URBAN (TU)

TAPS	Relative delay (us)	Fading/dB
0	0	-4
1	0.2	-3
2	0.4	0
3	0.6	-2
4	0.8	-3
5	1.2	-5
6	1.4	-7
7	1.8	-5
8	2.4	-6
9	3.0	-9
10	3.2	-11
11	5.0	-10

TABLE II
POWER DELAY PROFILE OF RURAL AREA (RA)

TAPS	Relative delay (us)	Fading/dB
0	0	0
1	0.1	-4
2	0.2	-8
3	0.3	-12
4	0.4	-16
5	0.5	-20

TABLE III
POWER DELAY PROFILE OF INDOOR-B

TAPS	Relative delay (us)	Fading/dB
0	0	0
1	0.1	-3.6
2	0.2	-7.2
3	0.3	-10.8
4	0.5	-18
5	0.7	-25.2

transmit diversity that depends on all relevant design parameters at the transmitter such as a modulation alphabet, signal scheme, transmit power and transmit covariance matrix. Applying QPSK modulation into (20), the upper bound of the PEP for QPSK transmitted signal with CDD transmit diversity over spatially correlated channel is expressed as:

$$P(C \to e) \leq \left(\frac{E_s}{4N_0}\right)^{-\gamma} \prod_{j=1}^{\gamma} \lambda_j^{-1}(S) \cdot 4d_m/N \tag{21}$$

where d_m is the distance for codeword C, N is the number of subcarriers. From (21), it can be seen that CDD cannot achieve the full diversity gain because of the channel correlation matrix, S, where the diversity gain is depending on the correlation coefficients of the MIMO channels.

IV. RESULTS AND DISCUSSIONS

This section presents the simulation results of mobile DVB-H with CDD transmit/receive diversity over the WSSUS-based uncorrelated and correlated MIMO channels for three different radio environments defined as Typical Urban (TU), Rural Area (RA) and Indoor-B in UHF band. The power delay profiles for the TU and RA are specified by COST207 [15], and Indoor-B is specified by ITU-R [16]. Tables I, II and III give the values of the tap delays and the associated mean powers of TU, RA and Indoor-B.

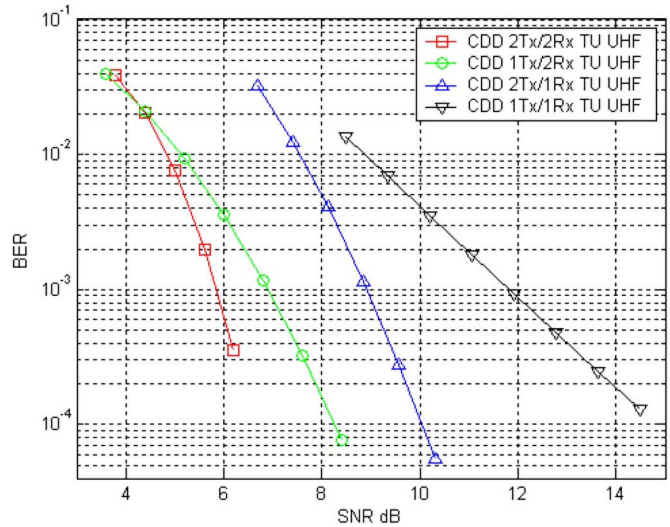

Fig. 5. Performance of CDD DVB-H in uncorrelated TU.

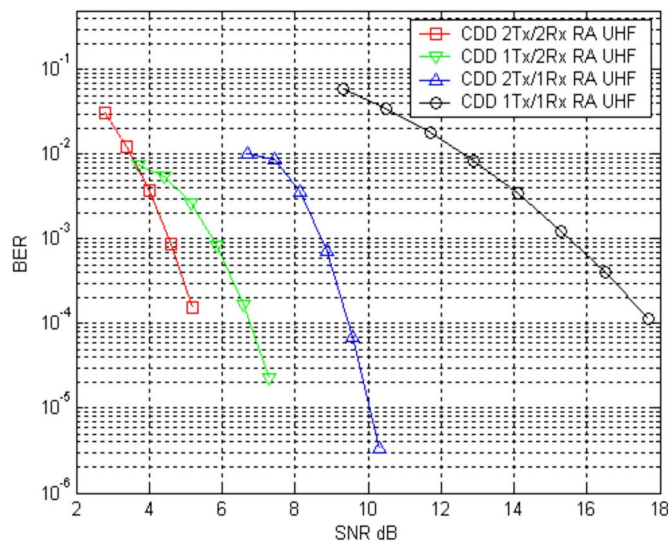

Fig. 6. Performance of CDD DVB-H in uncorrelated RA.

To enable good mobile reception, the DVB-H system is configured as follows: 4 K mode where the number of subcarriers used is 4096 in a bandwidth of 8 MHz, QPSK, code rate 1/2, and guard interval 1/8. The carrier frequency used is 900 MHz and the mobile velocity 10 meter/second in which the equivalent Doppler frequency is 30 Hz. The latter represents moving reflecting objects in which the signal is scattered. These configurations are applied to all simulations carried out in this paper.

A. Simulation Results for Uncorrelated Channels

Figs. 5, 6 and 7 show the BER performance for the TU, RA and RA radio environments with CDD transmit diversity applied to the DVB-H system in 4 K mode with QPSK modulation and code rate 1/2. The simulations are carried out with a maximum of 2-antenna transmitter (2Tx) and 2-antenna receiver (2Rx). For 2Rx with receive diversity, the signals are combined using MRC. The single-antenna (1Tx/1Rx) system in which there is no spatial diversity is simulated for reference.

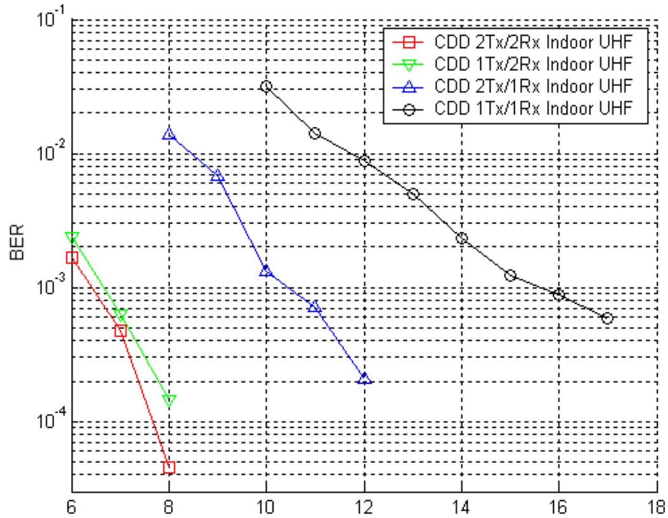

Fig. 7. Performance of CDD DVB-H in uncorrelated Indoor-B.

TABLE IV
RELATIONSHIP BETWEEN CHANNEL CORRELATION AND TRANSMITTER ANTENNAS SEPARATION

Correlation between two transmitter antennas (ρ)	Distance between two transmitter antennas (λ wavelength)
1	0
0.8	6
0.6	8
0.4	11
0.2	13
0	30

From the figures, it is observed that 2Tx/2Rx system outperforms the 1Tx/1Rx system by 6 dB for TU, 11.25 dB for RA, and 9.5 dB for Indoor-B at BER of 1×10^{-3}. The significant diversity obtained in the 2Tx/2Rx system is mainly because of additional signal power and antenna diversity at the receiver. CDD transmit diversity exploited the scattered signal propagation paths where under these conditions the channel is uncorrelated, with the receiver diversity the subcarriers that are deep faded at one receiver may have good channel properties at the other receiver. When combining these signals, the receiver has overall good signal condition. This effect is further demonstrated via the comparison between 2Tx/2Rx and 2Tx/1Rx where the gain achieved by the MRC-receiver is about 3.25 dB for TU, 4 dB for RA and 4.25 dB for Indoor-B at 2×10^{-4}. Note that the 1Tx/2Rx also performed particularly well compared to 1Tx/1Rx where the gain achieved by the MRC-receiver is about 5.5 dB for TU, 10.65 dB for RA and 10 dB for Indoor-B at BER of 2×10^{-4} (or 4.9 dB for TU, 10 dB for RA and 9 dB for Indoor-B at BER of 1×10^{-3}). From this discussion, it is interesting to see that the CDD transmit diversity (CDD-2Tx) gain decreases with the use of MRC at the receiver despite good BER performance.

When comparing 2Tx/1Rx to 1Tx/1Rx, it is observed that the diversity gain achieved in RA is the highest with about 8 dB, followed by Indoor-B (6 dB) and TU (4.5 dB). Note that the TU is frequency-selective and the channel does not undergo deep fading due to high maximum channel delay of 5 us. The impairment caused by this channel is inherently mitigated by the OFDM system itself; thus CDD has a little improvement. On the other hand, RA channel is non-frequency-selective (flat fading) with a severe deep fading because of rather short maximum channel delay of 500 ns; in this case, CDD increases the frequency-selectivity which explains higher diversity gain compared to TU. It is noticed that CDD works well especially when the channel undergoing deep fading which is usually caused by shorter maximum channel delay e.g. RA and Indoor-B.

B. Simulation Results for Correlated Channels

As evident in the previous section that systems with antenna diversity and MRC at the receiver (i.e. 2Tx/2Rx and 1Tx/2Rx) has the benefit to offer good BER performance and diversity gain where the received signals are likely to experience decorrelated fading at each spatially separated receive antenna, and that the MRC-receiver can effectively process the signals to its advantage. However, IST project PLUTO [17] aims to improve the DVB-T/H transmission in one way to increase channel capacity by exploiting diversity technique that is low cost and compatible to the existing network (fixed or mobile), and that the improved system should be able to serve existing consumers; therefore, it is anticipated that transmit diversity (i.e. 2Tx/1Rx) is more practical than receive diversity due to the difficulty of locating two receive antennas far enough apart in a small mobile device or at the roof of the consumers' residence. Moreover, MRC-receiver is not compatible to the existing DVB-T/H hardware equipment since its signal processing is carried out after the IOFDM. Thus, it is not practical to add additional antennas and change the receiver hardware at the consumers' end.

It is well understood that 2Tx/1Rx system exploits the statistical nature of fading in the scattered channel and reduces the likelihood of deep fading by providing a diversity of transmit signals; and that the signals at the receiver should have experienced uncorrelated fading in order to achieve the diversity gain and therefore BER improvement. However, in reality, the respective signal paths between spatially separated transmit antennas and the receiver is likely to be correlated because of insufficient antenna separation (or signal emission of certain antenna pattern) and the lack of the scattering effect of the channel. This correlation results in diversity loss of CDD and BER performance degradation.

This section presents the study of the degree of error performance degradation and diversity gain losses when CDD operates in the spatially correlated channel. Table IV shows the relationship between the channel correlation coefficient, ρ, and the transmitter antennas separation expressed in wavelength λ. The degree of channel correlation as in $\rho = \{0, 0.2, 0.4, 0.6, 0.8, 1\}$ corresponds to the distance between two transmit antennas. We assume that spatial fading correlation occurs at the transmitter and thus the correlated signals appear at the receiver. In this case, the correlated channel is expressed as $h_2 = \rho \cdot h_1 + \sqrt{1-\rho^2}w$ where h_1 and h_2 are the two uncorrelated fading channels, w is the Rayleigh fading factor and ρ is the correlation coefficient between h_1 and h_2. For example, $\rho = 0$ represents two uncorrelated fading channels. On the other hand, $\rho = 1$ represents a fully correlated fading channel.

Note that ρ is applied into (11) and the codeword in (2) is adjusted to record only two transmit streams such that it resem-

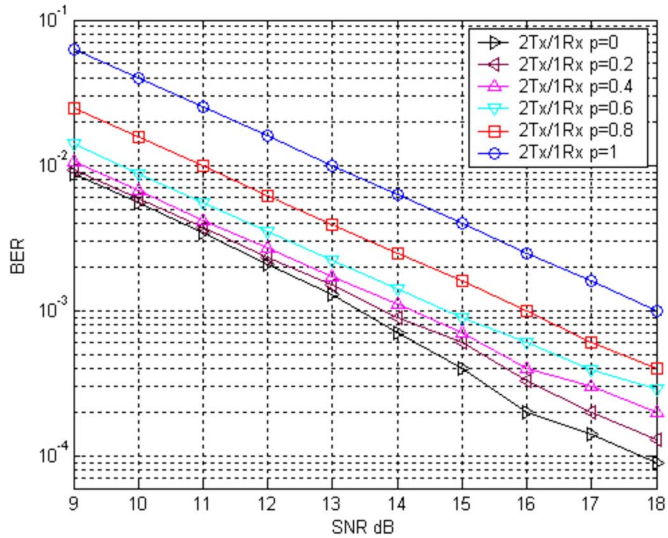

Fig. 8. Analytical curves of CDD-2Tx over different correlation coefficients.

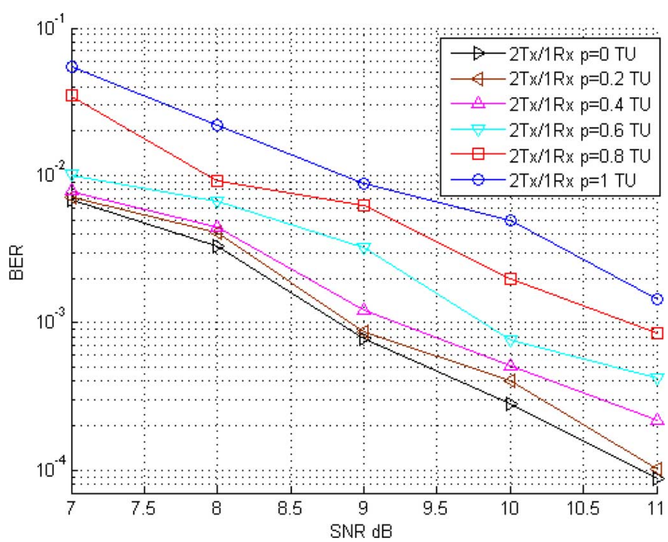

Fig. 9. BER performance of DVB-H with CDD-2Tx in correlated TU.

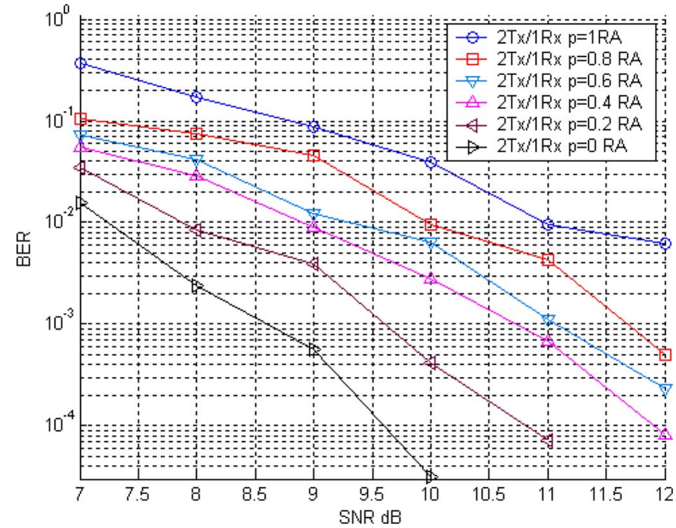

Fig. 10. BER performance of DVB-H with CDD-2Tx in correlated RA.

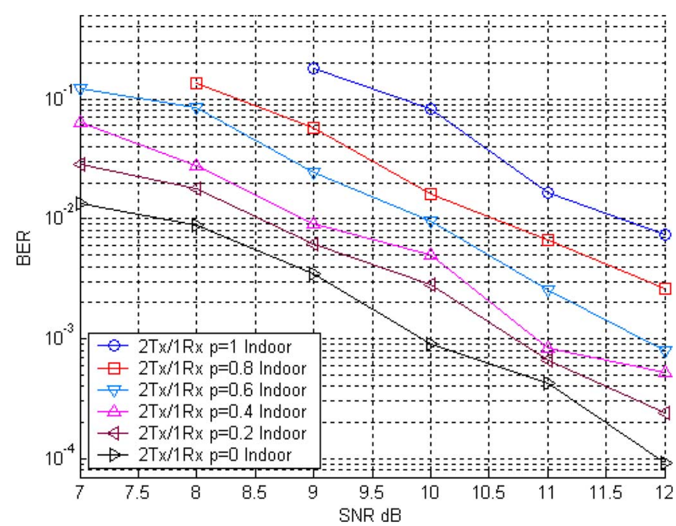

Fig. 11. BER performance of DVB-H with CDD-2Tx in correlated Indoor-B.

bles a 2Tx system. These parameters are applied into (21) to obtain the analytical curves of CDD-2Tx shown in Fig. 8. These curves indicate the worst-case error performance and diversity gain losses of the CDD over different correlation coefficients. With $\rho = 0$ as the reference bound, the figure shows that the CDD performance is degraded significantly in the fully correlated channel, $\rho = 1$, where the highest diversity gain loss of 4.6 dB is observed at BER of 1×10^{-3}. The gain loss is gradually reducing as the channel correlation reduces; for example $\rho = 0.8$, 0.6, 0.4 and 0.2 where the correspondence gain loss is 2.6 dB, 1.35 dB, 0.85 dB and 0.35 dB, respectively.

The performance of DVB-H with CDD-2Tx is simulated for three different radio environments in which the channels are spatially correlated. Figs. 9, 10 and 11 show the BER curves for TU, RA and Indoor-B with channel correlation coefficients of $\rho = \{0, 0.2, 0.4, 0.6, 0.8, 1\}$. In similar trend to the theoretical upper bound, it shows that the CDD performance is degraded significantly with fully correlated channel where the diversity gain loss is at its highest. However with a low channel correlation of $\rho = 0.2$, the performance is close to independent fading, $\rho = 0$, where the CDD diversity gain loss is about 0.1 dB for TU, 2 dB for RA and 0.9 dB for Indoor-B at BER of 2×10^{-3}. It is noticed that CDD operates in spatially correlated TU has the lowest gain loss followed by Indoor-B and the worse is RA. This effect is further demonstrated via the comparison between $\rho = 0.8$ and $\rho = 0$ where the gain loss in TU is the lowest that is about 1.5 dB followed by Indoor-B 2.9 dB and RA 3.3 dB. From these comparisons, it is clearly shown that the CDD can get better performance in the TU and Indoor-B channels that are naturally frequency-selective in the existence of high spatially correlated channel unlike as seen previously in the uncorrelated channel. We note that the CDD has less impact in RA as the channel itself is naturally flat fading and that the frequency-selectivity effect created by the CDD appears in the adjacent propagation paths due to the spatial correlation which caused the diversity to fail in randomizing the deep fading. Overall, CDD is able to perform well to a certain extent with channel correlation and the BER performance is better than the single-antenna system as shown in the previous section.

V. Conclusion

In this paper, the performance of CDD with two transmit antennas on DVB-H systems operating in spatially correlated channels has been studied via theoretical upper bound and BER simulations. The pairwise-error probability upper bound was derived resembling the two transmit antennas CDD system operating in a generalized Rayleigh fading channel with a coefficient to vary the antenna separation that corresponds to channel correlation. The theoretical results have generally shown that CDD exhibits significant performance degradation in the fully correlated channel where the diversity gain loss is highest. The gain loss is gradually reducing as the channel correlation reduces and with a low channel correlation the performance is close to independent fading. The BER simulation results suggest that CDD works well in the channel that is naturally flat fading like RA, under uncorrelated conditions where high diversity gain is obtained. Conversely, it is noted that CDD has less impact in the RA channel under correlated conditions as the frequency-selectivity effect created by the CDD appears in the adjacent propagation paths due to the spatial correlation which caused the diversity to fail in randomizing the deep fading. As a conclusion, we showed that CDD able to perform well with a certain extent of channel correlation and the BER performance is better than the single-antenna system.

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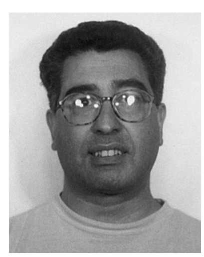

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