Diversity Gain for DVB-H by Using Transmitter/Receiver Cyclic Delay Diversity

Yue Zhang, John Cosmas, Maurice Bard, and Yong-Hua Song

Abstract—**The objective of this paper is to investigate different diversity techniques for broadcast networks that will minimize the complexity and improve received SNR of broadcast systems. Resultant digital broadcast networks would require fewer transmitter sites and thus be more cost-effective and have less environmental impact. The techniques can be applied to DVB-T, DVB-H and DAB systems that use Orthogonal Frequency Division Multplexing (OFDM). These are key radio broadcast network technologies, which are expected to complement emerging technologies such as WiMAX and future 4G networks for delivery of broadband content. Transmitter and receiver diversity technologies can increase the frequency and time selectivity of the resulting channel transfer function at the receiver. Diversity exploits the statistical nature of fading due to multipath and reduces the likelihood of deep fading by providing a diversity of transmission signals. Multiple signals are transmitted in such a way as to ensure that several signals reach the receiver each with uncorrelated fading. Transmit diversity is more practical than receive diversity due to the difficulty of locating two receive antennas far enough apart in a small mobile device. The schemes examined here comply with existing DVB standards and can be incorporated into existing systems without change. The diversity techniques introduced in this paper are applied to the DVB-H system. Bit error performance investigations were conducted by simulation for different DVB-H and diversity parameters.**

Index Terms—**CDD, CSI, DD, Diversity, DVB-H, MIMO, MRC, PD, WSSUS OFDM.**

I. Introduction

FUTURE mobile radio systems are expected to provide and serve a wide range of applications, which inherently require high data rates. Orthogonal Frequency Division Multiplexing (OFDM) [1] is a suitable technique for broadband transmission in multipath fading environments and is implemented in broadcast standards like digital audio broadcasting (DAB) [2] or terrestrial digital video broadcasting (DVB-T) [3] as well as wireless local area network (WLAN) standards [4] such as HIPERLAN/2 or IEEE 802.11a and 802.16.

Because of the poor error performance of OFDM in scattering environments, it is necessary for wireless communication systems to use spatial diversity to improve the error performance and channel capacity. Multiple antennas are an important means to improve the performance of wireless systems. It is widely understood that in a system with multiple transmit and receive antennas (multiple-input-multiple-out (MIMO) channel), the spectral efficiency is much higher than that of the conventional single-antenna channels. Traditionally, multiple antennas have been used to increase diversity to combat channel fading. Each pair of transmit and receive antennas provides a different signal path from the transmitter to the receiver. By sending signals that carry the same information through different paths, multiple independently faded replicas of the data symbol can be obtained at the receive end. It is well known that maximum diversity gain can be achieved if fading is independent across antenna pairs.

More recent work has concentrated on using multiple transmit antennas to get diversity (such as trellis-based space-time codes [5], [6] and orthogonal designs [7]). Unfortunately, the space-time-coding is not suitable for extending existing systems, because this would require non-standards compliant modifications to be made. Therefore, for standardized systems only additional spatial diversity techniques can be implemented, provided these modifications keep the systems standards compatible. A very simple and elegant method called cyclic delay diversity (CDD) [8] was proposed for broadcasting systems.

Handheld digital video broadcast (DVB-H) [9] is a new broadcast standard. DVB-H is an extension to the older DVB-T standard. DVB-H is optimized for delivery of broadcast services to mobile users, since it supports burst mode reception of broadcast transmission thus saving power on the end-user terminal and since it supports handover between broadcast cells by using the silent burst periods for channel sounding broadcast transmissions in adjacent broadcast coverage areas.

This makes it ideal for mobile phones and handheld computers to receive digital TV broadcasts over the digital TV network (without using mobile phone networks at all).

This paper is organized as follows; Section II introduces the system model; in Section III, we discuss the diversity criterion for spatial diversity and diversity scheme for both transmitter and receiver. The principle of cyclic delay diversity is explained. The main idea is to increase the frequency selectivity of the channel transfer function by specific cyclic delays at the transmitter side. Maximum Ratio Combining (MRC) is also consid-

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ered at the receiver side. Section IV shows how the discussed diversity scheme is applied to DVB-H in order to improve the bit error performance in multipath environments. In Section V, we provide the simulation results for the DVB-H system and transmission parameters. Finally, Section VI presents the discussions and conclusion of the simulations performed in this study.

II. System Model

A. OFDM System Model

OFDM is a promising technique for achieving high data rate and combating multipath fading in wireless communication. OFDM can be thought of as a hybrid of multi-carrier modulation (MCM) and frequency shift keying (FSK) modulation. Orthogonality among the carriers is achieved by separating the carriers by an integer multiple of the inverse of symbol duration of the parallel bit streams. The OFDM signal consists of N orthogonal subcarriers modulated by N parallel data streams. Denoting the frequency and complex source symbol of the k^{th} subcarrier as f_k and d_k respectively, the baseband representation of an OFDM is

$$x(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} d_k e^{j2\pi f_k t} \qquad 0 \le t \le NT \tag{1}$$

d_k is typical taken from a PSK or QAM symbol constellation, and NT is the duration of the OFDM symbol. The subcarrier frequencies are equally spaced at $f_k = k/NT$. The OFDM signal in equ-(1) (equ means math equation) can be derived by using a single IDFT operation rather than using a bank of oscillators. Therefore the OFDM symbol can be represented:

$$x[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} d_k W_N^{kn}, \qquad 0 \le n \le N-1 \tag{2}$$

where $W_N = e^{j2\pi/N}$. Assuming that the channel impulse response is shorter than the guard interval and that perfect synchronization is achieved.

B. Channel Model

We consider an equivalent baseband MIMO channel with N transmit antennas and M receiver antennas. There are MN i.i.d Rayleigh-faded channels. Each channel is Wide-Sense-Stationarity-Uncorrelated-Scattering (WSSUS) channel. According to [10], the impulse response of the channel is modeled by several paths each consisting of a zero mean complex-values Gaussian process, the envelope of which is Rayleigh-distributed. If the process does not have a zero mean, the envelope has a Rice distribution and the channel is said to be a Ricean fading channel. Suppose the channel is composed of D echoes. The instantaneous channel impulse response function can be given as:

$$h(t,\tau) = \sqrt{1-\rho^2}\delta(\tau) + \frac{\rho}{\sqrt{D}} \sum_{n=1}^{D} e^{j\theta_n} e^{j2\pi f D_n t} \delta(\tau - \tau_n) \tag{3}$$

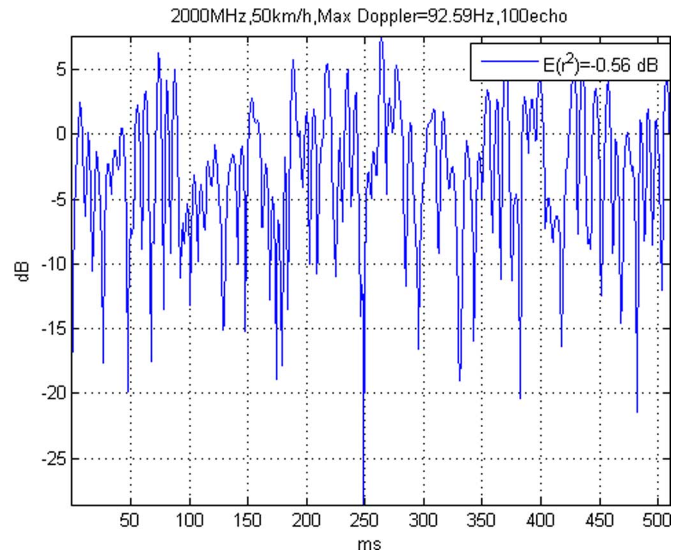

Fig. 1. Rayleigh fading channel.

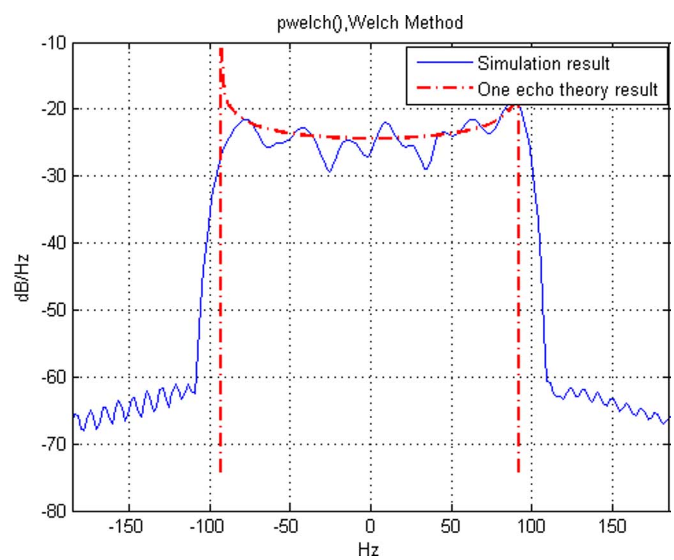

Fig. 2. Classical and simulated Doppler spectrum.

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

The transfer function for the channel model is given by

$$H(f,t) = \int_{-\infty}^{+\infty} h(\tau,t) e^{-j2\pi f\tau} d\tau \tag{4}$$

The classical Doppler Spectrum is;

$$s(f) = \frac{p}{\pi F_d \sqrt{1 - \left(\frac{f}{F_d}\right)^2}} \tag{5}$$

where P is the signal power, f is the frequency and F_d is the maximum Doppler shift.

Fig. 1 shows the simulation of the Rayleigh fading channel according to the equ-(3). Fig. 2 shows the classical Doppler spectrum according to equ-(5) and simulated Doppler spectrum. From the Fig. 2, we can see that the simulation results are almost the same as the theoretical.

As for the MIMO channel, there are N transmit antennas and M receive antennas. 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 equ-(3). h_{ij} is assumed to be known to the receiver, but not at the transmitter. Consider the discrete baseband system, at sample index l, the complex symbol, $s(l)$, sent by N transmit antennas and detected by k^{th} receive antenna is denoted as $r_k(l)$.

$$r_k(l) = \sqrt{\frac{SNR}{N}} \sum_{i=1}^{N} h_{ki}(l) \otimes s_i(l) + v_k(l) \tag{6}$$

SNR is the average SNR at each receive antenna. $v_k(l)$ are complex zero-mean spatially and temporally white Gaussian random variable with variance $N_0/2$ per dimension. In frequency domain:

$$R_k(m) = \sqrt{\frac{SNR}{N}} \sum_{i=1}^{N} H_{ki}(m) S_i(m) + V_k \tag{7}$$

where $R_k(m)$, $H_{ki}(m)$, $S_i(m)$ and 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 will be:

$$\tilde{R}(m) = \sqrt{\frac{SNR}{N}} \tilde{H}(m) \tilde{S}(m) + \tilde{V}(m) \tag{8}$$

where

$$\begin{aligned} \tilde{R}(m) &= [R_1(m), R_2(m), \ldots\ldots, R_M(m)]^T \\ \tilde{S}(m) &= [S_1(m), S_2(m), \ldots\ldots, S_M(m)]^T \\ \tilde{H}(m) &= \begin{matrix} H_{11}(m) & H_{12}(m) & \ldots & \ldots & H_{1N}(m) \\ H_{21}(m) & \ldots & \ldots & \ldots & H_{2N}(m) \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ H_{M1}(m) & \ldots & \ldots & \ldots & H_{MN}(m) \end{matrix} \end{aligned} \tag{9}$$

Given this necessary mathematical framework, we can proceed to describe the antenna diversity techniques.

III. Spatial Antenna Diversity

This section describes spatial antenna diversity including Delay diversity (DD), Cyclic Delay Diversity (CDD), Phase Diversity (PD), Maximum Ratio Combining (MRC) and the criterion for achieving full diversity with spatial antenna diversity.

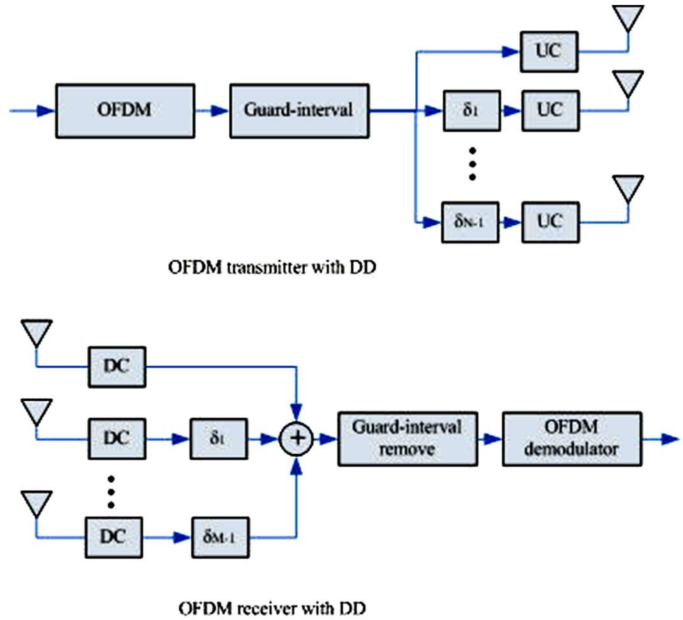

Fig. 3. Delay diversity with OFDM.

A. Diversity Criterion

Assume a MIMO channel with N transmit and M receiving antennas. A length-J space-time code will be an NJ matrix $\boldsymbol{C}$.

$$C = \begin{pmatrix} x_{00} & x_{01} & \ldots & x_{0(J-1)} \\ \ldots & \ldots & \ldots & \ldots \\ x_{(N-1)0} & x_{(N-1)1} & \ldots & x_{(N-1)(J-1)} \end{pmatrix}_{N\times J} \tag{10}$$

where x_{ln} is the code symbol to the l^{th} transmit antenna at time index n, $l = 0, \ldots, N-1$, $n = 0, \ldots, J-1$ and the x_{ln} are normalized $E[|x_{\text{ln}}|^2] = 1/N$. After codeword C is transmitted, the pairwise error probability $P(C \to e)$ is the probability that e is more likely to have been sent than C. For quasi-static flat Rayleigh fading channels, the average pairwise error probability can be represented as [6], [11], [15],

$$P(C \to e) \le \prod_{s=0}^{S-1} \prod_{l=0}^{r-1} \frac{1}{1 + \frac{E_s}{4N_0}\lambda_l} \le \left(\frac{E_s}{4N_0}\eta\right)^{-rs} \tag{11}$$

E_s is the average transmitted symbol energy from the N transmit antennas, $\eta = (\lambda_0 \ldots\ldots \lambda_{r-1})^{1/r}$ is the geometric mean of nonzero eigen values $(\lambda_0 \ldots\ldots \lambda_{r-1})$ of the MM matrix $A = (C-e)(C-e)^H$, and r is the rank of A. The diversity order is rs in (11), and the coding gain is

$$G_c = \eta = (\lambda \ldots\ldots \lambda_{r-1})^{1/r} \tag{12}$$

When full diversity has been achieved $(r = N)$, the diversity order will be NS and A has full rank. The coding gain in the case of full diversity is $G_c = (\det|A|^{1/N}$.

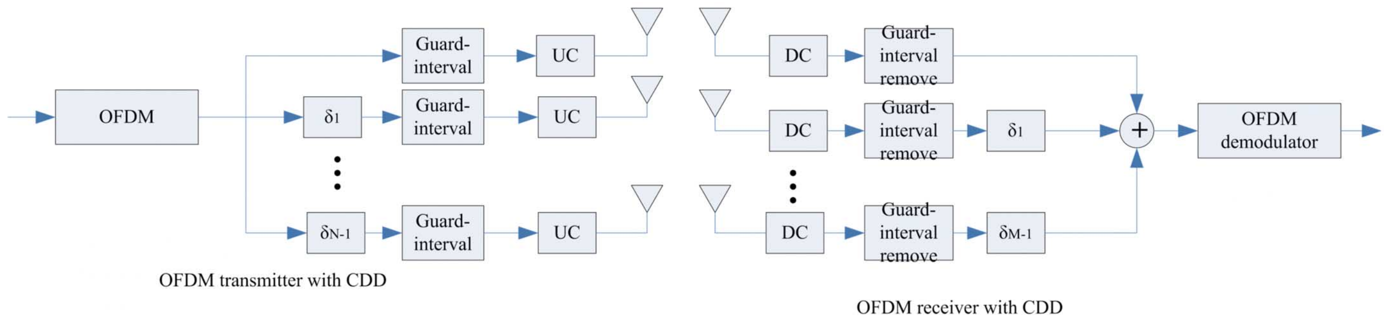

Fig. 4. Cyclic delay diversity with OFDM.

B. Full Diversity for Delay Diversity

Fig. 3 shows the block diagram of an N transmit antennas OFDM system with DD. Fig. 4 shows the block diagram of OFDM system with CDD. The difference between DD and CDD is the position of Guard-interval process. In CDD, the choice of time delay δ should not be restricted. Concurrently, no ISI can occur with CDD. The OFDM symbols of CDD signal can be generated from the reference signal OFDM symbols just by applying a cyclic time shift of δ to the reference signal OFDM symbols and subsequent insertion of the cyclic prefix.

According to equ-(2), a length K sequence modulates K sub-carriers.

$$s(l) = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} S(k) \cdot e^{j\frac{2\pi}{K}kl} \qquad l = 0, \ldots, K-1 \tag{13}$$

Now the space-time code scheme for delay diversity with N transmitter antennas will be a NK matrix. The codeword can be represented as

$$C = \begin{pmatrix} s(0) & s(1) & \ldots & s(K-1) \\ s(K-1) & s(0) & \ldots & S(K-2) \\ \ldots & \ldots & \ldots & \ldots \\ s(K-N+1) & s(K-N+2) & \ldots & s(K-N) \end{pmatrix}_{N\times K} \tag{14}$$

where the n^{th} antenna transmits sequence $(s(0), \ldots, s(K-1))$ with shift n, as $(s(K-n), s(K-n+1), \ldots, s(K-n-1))$.

To achieve full diversity for delay diversity, according to Section III-A, the matrix $A = (C-e)(C-e)^H$ should have full rank for all codeword pairs (C, e), $C \neq e$. All row vectors of $(C-e)$ should be linearly independent. According to Appendix, if $K > N$ and $S(K) \neq 0$, the $s(l)$ and its $N-1$ cyclic shifts symbols are linearly independent. Furthermore, if all pairs of sequences (C, e), $C \neq e$ differ in at least N coordinates, then CDD achieves full spatial diversity on quasi-static flat Rayleigh fading channel.

Based on the full diversity criterion, CDD with N transmit antennas can obtain full spatial diversity by using the cyclic shifts space time code C. The minimum distance in C is at least equal to N, $d_m \geqslant N$. As for the codeword C, it is easy to achieve full spatial diversity just by shifting more than N bits binary in each row in C. Therefore, according to Section III-A, the performance of CDD depends on both the diversity gain and coding gain. The coding gain is affected by the minimum coding gain $G_c = (\det|A|)^{1/N}$. The transmitted CDD space-time codeword is

$$C = \begin{pmatrix} s \\ Ds \\ \ldots \\ D^{N-1}s \end{pmatrix} \tag{15}$$

D means bit shift of the binary sequence. According to equ-(13), the sequence s can be represented as $s(l) \xrightarrow{DFT} S(k)$. The transmitted power from the N antennas is normalized at any k in frequency domain $E[S_k^2] = 1/N$. The Hamming distance in C is d_m. Therefore, the Hamming distance in $S(k)$ also is d_m. The element a_{ij} of the matrix $A = (C-e)(C-e)^H$ can be represented as [15]

$$\begin{aligned} a_{ij} &= (D^i s - D^i e)(D^j s - D^j e)^H \\ &= \sum_{n=0}^{K-1} \left(s_{(n-i)K} - e_{(n-i)K}\right)\left(s^*_{(n-j)K} - e^*_{(n-j)K}\right) \end{aligned} \tag{16}$$

And according to equ-(13), the equ-(16) will be

$$a_{ij} = \sum_{k=0}^{K-1} |S_k - E_k|^2 \, e^{-j(2\pi/K)k(i-j)} \tag{17}$$

From equ-(17), it is obvious that A is a Toeplitz matrix. According the characters of the Toeplitz matrix, the diagonal elements of matrix A are identical.

$$a_{ii} = \sum_{k=0}^{K-1} |S_k - E_k|^2 = 4d_m/N \tag{18}$$

$$\mathrm{Tr}(A) = \sum_{i=0}^{N-1} \lambda_i = 4d_m > 0 \tag{19}$$

Then

$$\left(\sum_{i=0}^{N-1} \lambda_i \Big/ N\right) \geq \left(\prod_{i=0}^{N-1} \lambda_i\right)^{1/N} \tag{20}$$

Therefore, from equ-(20), the upper bound of coding gain will be

$$G_c = (\det|A|)^{1/N} \leq 4d_m/N \tag{21}$$

When all eigen values λ_{i} are equal, the equality can be held.

$$\lambda_i = 4d_m/N, \quad i = 0 \ldots\ldots N-1.$$

Define the maximum coding gain

$$G_m = 4d_m/N \tag{22}$$

Therefore, both the spatial diversity gain and coding gain are decided by d_m.

To achieve the maximum G_m, the interleaver should be applied to the system. For a linear code C with minimum distance d_m, the optimal interleaver is an interleaver with permutes the nonzero bits of all weight d_m codewords of C such that they are uniformly distributed within the length K block. In DVB system, for a convolutional code, the conventional "write in row read in column" block interleaver is used to improve the coding gain for the whole system.

C. *The Scheme of Cyclic Delay Diversity*

According to equ-(6), the impulse response from transmit antenna n to receive antenna m at time index t can be represented as:

$$h_{ij}^t = \left[h_{ij}^t(0), h_{ij}^t(1), \ldots, h_{ij}^t(\delta_{\max}), 0, \ldots, 0\right]_{1\times k} \tag{23}$$

where $\delta_{\max}$ is the max delay per subchannel from a transmit to a receive antenna. And according to equ-(14) and (15), the transmit symbol from antenna i at time t is given by:

$$s_i(t) = s_i\left((t-\delta^{cy}) \mathrm{mod\ K}\right), \quad \mathrm{t} = 0, \ldots, \mathrm{K}-1, \quad \mathrm{i} = 0, \ldots, \mathrm{N}-1 \tag{24}$$

The system is equivalent to transmission of sequence s over a frequency selective channel with one transmit antenna and the channel impulse response will be:

$$h_{1,j}^{e,t} = \left[h_{1,j}^{e,t}(0), \ldots, h_{1,j}^{e,t}(K-1)\right] \tag{25}$$

to receive antenna j, $j = 1, \ldots, M$ and

$$h_{1,j}^{e,t}(d) = \sum_{i=1}^{N} h_{i,j}^t\left((d-\delta^{cy}) \mathrm{mod\ K}\right) \tag{26}$$

In frequency domain, the equivalent channel transfer function

$$H_{1,m}^e(k) = \sqrt{\frac{SNR}{N}} \sum_{n=1}^{N} e^{-j\frac{2\pi}{K}k\delta_n^{cy}} \cdot H_{n,m}(k) \tag{27}$$

where $H_{n,m}(k)$ denotes the channel transfer function from the n^{th} transmit antenna to the m^{th} receive antenna and δ_n^{cy} stands for the transmit antenna specific cyclic delay ($\delta_1^{cy} = 0$). For 2 transmit antenna system,

$$\left|H_m^e(k)\right|^2 = \left|H_{1,m}(k) + e^{-j\frac{2\pi}{K}k\delta_2^{cy}} H_{2,m}(k)\right|^2.$$

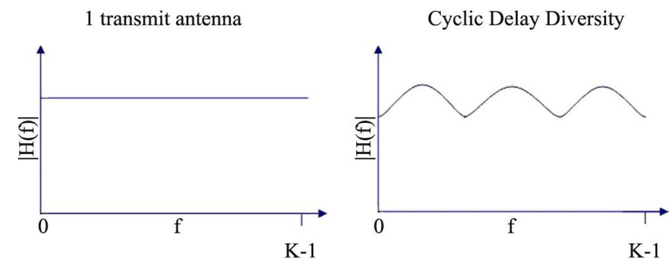

Fig. 5. Channel transfer function by cyclic delay diversity.

Therefore, CDD has transformed the MIMO channel into a single input multiple output (SIMO) channel with increased frequency selectivity. The spatial diversity is transformed into frequency diversity. The effect is illustrated in Fig. 5. The original 1 transmit antenna channel is frequency flat in order to isolate the spatial effect. In a flat fading channel, the bit error rate (BER) will be the same on each subcarrier. CDD transforms the channel into a frequency selective channel. The average BER for uncoded transmission will roughly be the same as in flat fading channel. However, the BER is not constant over the subcarriers. An outer FEC coder and decoder can use the frequency selectivity where the strong subcarriers help the weak ones. In case of a convolutional code, the maximum diversity level is determined by the d_m of the code as mentioned in Section III-B.

In order to achieve constructive and destructive superposition of the signals with bandwidth of the K subcarriers, the δ_n^{cy} has to fulfill the condition

$$\delta_n^{cy} \geq \frac{1}{B} \forall n \qquad B = K/T_s \tag{28}$$

where B is the bandwidth of the transmitted signal and T_s is the sampling rate of the transmitted signal. The different antennas have to be chosen as [8]:

$$\delta_n^{cy} = \gamma \frac{(n-1)}{B} \qquad k \geq 1, \quad n = 2, \ldots, N \tag{29}$$

where γ is a constant factor introduced for the system design which has to be chosen large enough ($\gamma \geqslant 1$) in order to guarantee the diversity gain. The parameter γ has to be determined by simulations. And $\gamma = 2$ is sufficient to achieve a promising performance improvement. This result is verified by the simulation results presented in Section V.

As for phase diversity (PD), the equivalence between PD and CDD is a property of the DFT. According to equ-(13),

$$\underbrace{s(l-\delta^{cy}) \mathrm{mod}\ K)}_{CDD\mathrm{signal}} = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} \underbrace{e^{-j\frac{2\pi}{K}\cdot k\cdot\delta^{cy}} \cdot S(k)}_{PD\mathrm{signal}} \cdot e^{j\frac{2\pi}{K}k\cdot l} \tag{30}$$

A cyclic delay δ_n^{cy} in the time domain corresponds to a phase factor of $e^{-j(2\pi/K)k\delta^{cy}}$ in the frequency domain. Equ-(30) shows that the operation for PD has to be done before OFDM modulation. There is no delay of the signals at the transmit antennas. CDD and PD are independent of the existence of

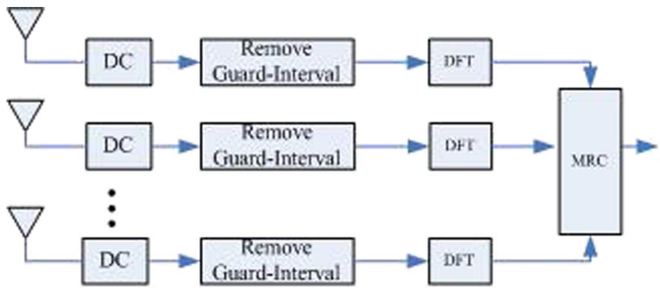

Fig. 6. OFDM receiver with MRC.

a guard interval and are capable to increase the channel frequency selectivity without increasing the overall channel delay spread because these operations are done before guard interval insertion and are restricted to the OFDM symbol itself.

D. Receive Diversity

The use of multiple antennas at the receiver, which is referred to as receive diversity, is fairly easily exploited. Maximum Ratio Combining (MRC) is used in the system. In MRC, the signals at the output of the M receive antennas are combined linearly so as to maximize the instantaneous SNR [12]. This is achieved by combining the cophased signals, which requires that the CSI (channel state information) is known at the receiver. The SNR of the combined signal is equal to the sum of the SNR of all the branch signals [13].

For an MRC system as shown in Fig. 6, the combining operations are performed at subcarrier level after the DFT operation. The received OFDM signals at different antenna branches are first transformed via M separate DFTs. Their outputs are assigned to N diversity combiners. According to equ-(7), the received signal in the m^{th} antenna is:

$$R_m(l) = H_m^e(l)S(l) + V_m \tag{31}$$

Further by assuming perfect channel state information at each antenna. MRC consists of using the linear combination prior to detection.

$$\begin{aligned} y(l) &= \sum_{m=1}^{M} H_m^E \cdot R_m(l) \\ &= \sum_{m=1}^{M} H_m^E \cdot H_m^e(l)S(l) + \sum_{m=1}^{M} H_m^E \cdot V_m \end{aligned} \tag{32}$$

H_m^E is the perfect channel estimation for the m^{th} receive antenna. The performance improvement is significant for MRC. From [7], the MRC scheme optimizes the SNR for each subcarrier.

IV. Application to the DVB-H System

The above mentioned techniques can be standard applied to broadcasting systems complying with the DAB and DVB-T/H standards. In this section, we will apply CDD and MRC to the DVB-H system. DVB-H is based on DVB-T and is a coded OFDM system containing an outer shortened Reed-Solomon (RS) code concatenated with an inner (punctured) convolutional code.

For the implementation of CDD at the DVB-H transmitter, only a second signal path after the OFDM modulation has to be added. Fig. 7 shows the transmitter and receiver with transmitter CDD. After channel coding (RS and Convolutional Code) 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 split, upconverted and transmitted directly on the one hand and cyclically shifted on the other hand.

The model uses an MRC receiver for simulations. After downconversion, synchronization and guard interval removal, the received signal is OFDM demodulated and equalized using zero forcing. We assume perfect knowledge of the channel state information. Both complex valued symbol streams are combined and QAM demodulated with soft–out values before symbol and bit-deinterleaving is computed. Finally, the bit stream is soft-decision-maximum-likelihood (SDML) decoded in a Viterbi decoder.

DVB-H is based on DVB-T. From Fig. 8, we can see that the DVB-H adds three modules to DVB-T system in the physical layer. One is 4K mode, one is DVB-H TPS and the other is in-depth interleavers. DVB-H includes a new transmission in the DVB-T physical layer using a 4096 FFT size. In additional to the 2K and 8K transmission modes provided originally by the DVB-T standard, the 4K mode brings additional flexibility in network design by trading off mobile reception performance and size of SFN networks. And DVB-H is principally a transmission system allowing reception of broadcast information on single antenna hand-held mobile devices. In the DVB-T system, the 2K transmission mode is known to provide better mobile reception performance than the 8K mode, due to the larger inter-carrier spacing it implements. However, since the duration of the 2K mode OFDM symbols, the associated guard intervals duration are very short. This makes the 2K mode only suitable for small size SFN. However, 4K OFDM symbol has a longer duration and longer guard interval than 2K mode. This makes 4K mode suitable for medium size SFN networks. It can increase the spectral efficiency for SFN networks planning. As for 8K mode, the symbol duration of 4K mode is shorter than in the 8K mode and channel estimation can be done more frequently in the demodulator. Therefore, it provides a better mobile performance than 8K mode, although not as high as with the 2K transmission mode, it is enough for the use of DVB-H scenarios. So 4K mode provides a good trade off for the two sides of the system: spectral efficiency for the DVB-H network designers and high mobility for the DVB-H consumers. According to [14], the parameters of 4K mode will be as shown in Table I.

V. Simulations

In this section we will present simulation results for DVB-H with antenna diversity in Typical Urban (TU), Bad Urban (BU) and VHF, UHF and L-band frequency carrier. The simulations included Doppler effects.

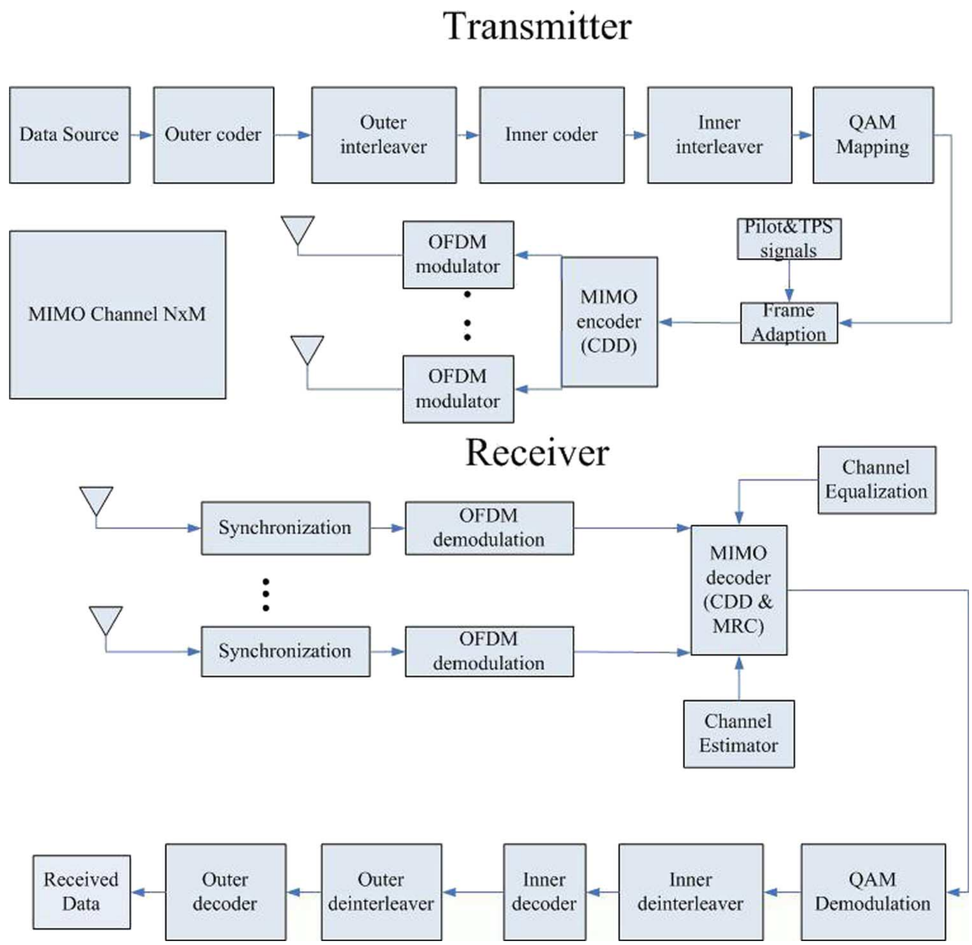

Fig. 7. DVB-H system diagram with CDD.

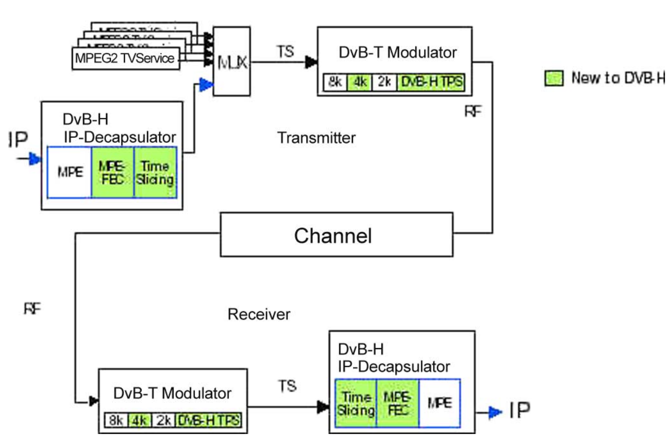

Fig. 8. A conceptual description of using a DVB-H system (sharing a MUX with MPEG-2 service).

TABLE I
OFDM PARAMETERS FOR THE 4K MODE

Parameter	4K mode			
Elementary period T	7/64 us			
Number of carriers	3409			
Value of carrier number Kmin	0			
Value of carrier number Kmax	3408			
Duration Tu	448us			
Carrier spacing 1/Tu	2232Hz			
Spacing between carriers Kmin and Kmax	7.61MHz			
Allowed guard interval Δ/Tu	1/4	1/8	1/16	1/32
Duration of symbol part Tu	4096xT 448us			
Duration of symbol interval Δ	1026xT 112us	512xT 56us	256xT 28us	128xT 14us
Symbol duration Ts=Δ+Tu	560us	504us	476us	462us

A. Channel Model

Tables II and III show the main properties of the WSSUS channel models used for simulations. For the individual scatters Rayleigh fading is assumed according to Section II-B. The mobile radio channel models are based on COST207 and described in [16]. The mobile velocity is 10 m/s. The mobile velocity will cause Doppler effects. As a rule, the higher speed causes the larger Doppler effects. The carrier frequencies are VHF (170.23 MHz), UHF (900 MHz) and L-band (1495 MHz).

TABLE II
MAIN CHANNEL MODEL PROPERTIES (TYPICAL URBAN)

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 III
MAIN CHANNEL MODEL PROPERTIES (BAD URBAN)

TAPS	Relative delay (us)	Fading/dB
0	0	-7
1	0.2	-3
2	0.4	-1
3	0.8	0
4	1.6	-2
5	2.2	-6
6	3.2	-7
7	5.0	-1
8	6.0	-2
9	7.2	-7
10	8.2	-10
11	10.0	-15

B. Results

The simulation is according to the Monte Carlo Method. The overall transmitted power is equal for all simulation runs. The total transmit power with N antennas equals to the transmit power with 1 antenna in the following analysis. The power per transmit antenna decreases with an increasing number of antennas N. The antennas are placed such that their channel transfer functions can be considered as uncorrelated. For the simulations, the most number of antennas used at the transmitter and the receiver side were two. For the BER vs. SNR simulations of the 2TX-antenna CDD systems a cyclic delay of $\delta^{cy} = 15 \cdot (T_u/K) \approx 1.6$ us $\geq (1/B) \approx 0.13$ us is chosen. Fig. 9 shows the bit error rate vs. the SNR for the Typical Urban channel with different diversity techniques, applied to the DVB-H system in 4K mode with 4-QAM modulation and code rate 1/2 in UHF (900 MHz). A single antenna system is given as a reference. For this system no spatial diversity is implemented. From Fig. 10, the receiver MRC system outperforms the single receive antenna system by about 9.5 dB in SNR at a BER of $2\mathrm{e}^{-4}$. The reason is the 2nd receiver antenna, provides the receiver with additional signal power. As for the transmit diversity, there is about 5 dB diversity gain in SNR at a BER of $2\mathrm{e}^{-4}$. The powerful channel codes also provide the additional coding gain besides diversity gain. There are two uncorrelated propagation paths in the 2 receiver antenna systems. The subcarriers that are in deep fade for receiver antenna 1 may have good channel properties for antenna 2. As shown in Section III-C, CDD increases the channel frequency selectivity and therefore decreases the occurrence of the bit errors after demodulation. The bit errors may also appear before decoding in spite of interleaving due to extremely wide deep fades. Fig. 11 shows the transmitter CDD gain for single receiver antenna and two MRC-receive antennas in typical urban environment. According to equ-(29), $\gamma = 2$ is sufficient to achieve promising performance improvement and the total gain will be 4.5 dB. Fig. 12 shows the bit error performance for DVB-H system in 4k mode with 4QAM modulation and code rate 1/2 for the Bad Urban channel in L-band. If MRC is used at the receiver, the most gain can be achieved. The Bad Urban channel provides a higher maximum channel delay and therefore the higher frequency selectivity. Fig. 13 shows the transmitter CDD gain for single antenna and 2 antenna MRC receivers in the Bad Urban environment. From Fig. 11 and Fig. 13, it is obvious that a cyclic delay $\delta > 1.2$ us results in no further improvement. Fig. 11 and Fig. 13 also show that the achievable gain of Bad Urban is much higher than for Typical Urban channels due to the extremely different maximum channel delay. For Bad Urban, the max delay is 15 us and for Typical Urban, the max delay is 5 us. In Bad Urban channel, the CDD can achieve more frequency selectivity for the channel transfer function than Typical Urban. Thus, Bad Urban can achieve more diversity gain than Typical Urban. Fig. 14 shows the simulation results for the bit error performance of a DVB-H system in 4K mode with 4QAM modulation and code rate 1/2 for Typical Urban in VHF.

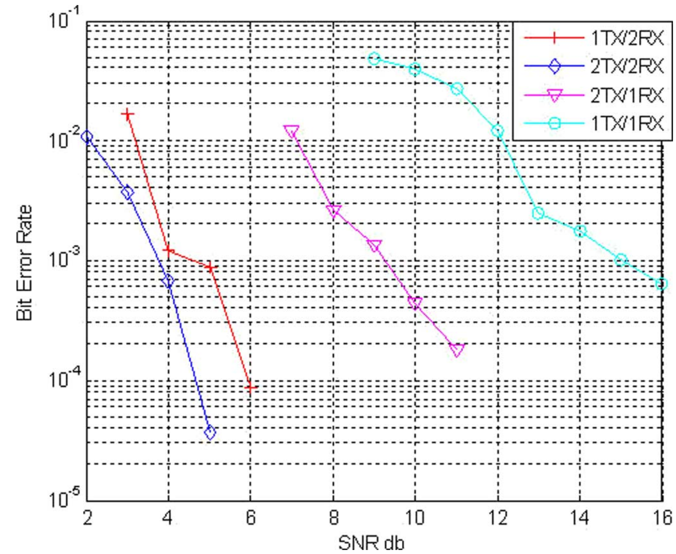

Fig. 9. BER vs. SNR for 4K mode, 4QAM, code rate 1/2, typical urban, in UHF(900 MHz).

From Fig. 10, the diversity techniques increase the frequency selectivity of the channel. For our investigation, perfect knowledge of the CSI and synchronization are assumed. Optimal interleaver and convolutional coding are applied to the system to achieve to the theoretical maximum diversity and coding gain. Fig. 15 illuminates the effect of the interleaver. There is about 5 dB code gain for the system with interleaver and 2 transmit antennas. Interleaver increases the d_m for the whole system.

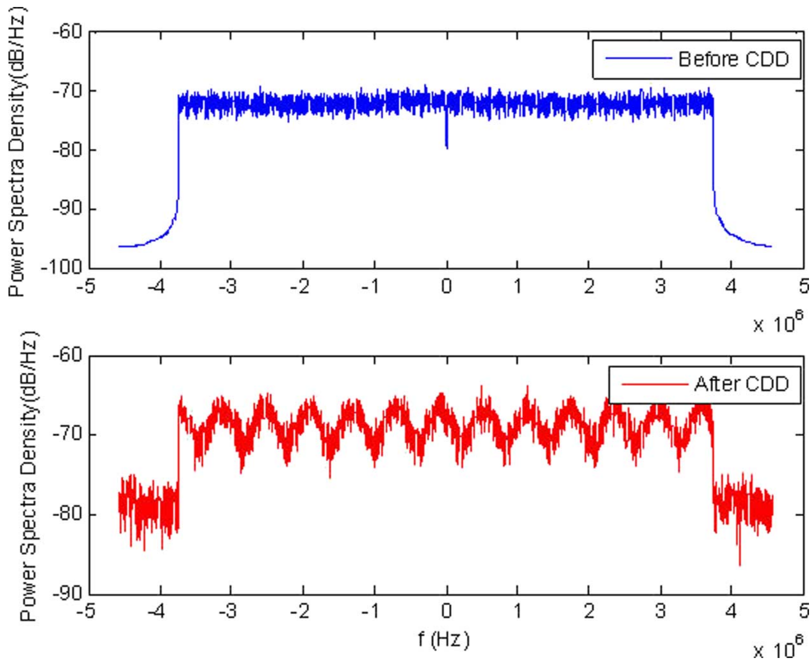

Fig. 10. Power spectra density of signals before CDD and after CDD UHF (900 MHz).

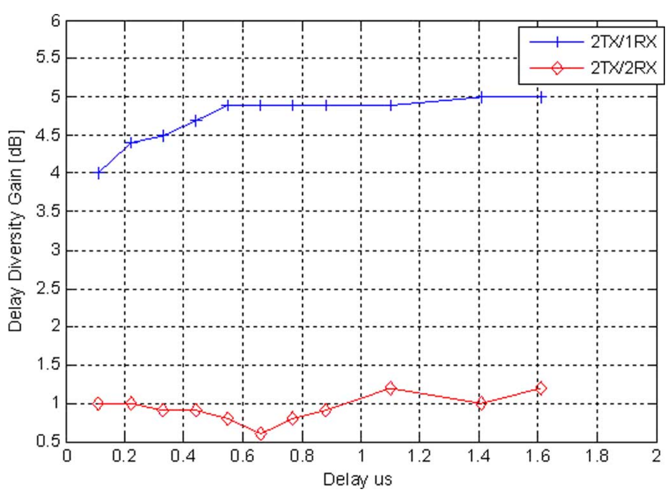

Fig. 11. Delay diversity gain vs. Delay for 4K mode, 4QAM, code rate 1/2, typical urban, in UHF (900 MHz).

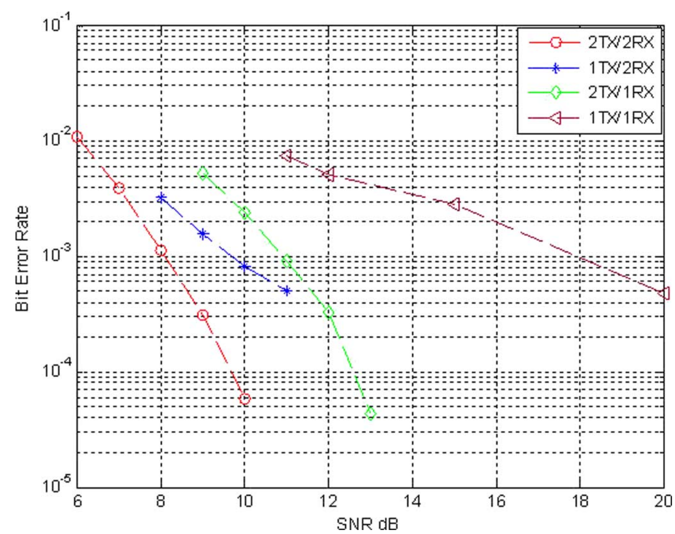

Fig. 12. BER vs. SNR for 4K mode, 4QAM, code rate 1/2, bad urban, in L-band (1495 MHz).

VI. Conclusion

In this paper, the principles of delay diversity gain and coding gain, delay diversity, cyclic delay diversity, phase diversity and maximum ratio combining have been discussed. The equivalence between cyclic delay diversity and phase diversity has been shown. The DVB-H system integrated CDD technique has been investigated. Combined with powerful channel coding, CDD can achieve desirable spatially diversity and coding gain (at least 5 dB) for SFN network planning. CDD is an elegant low cost transmit diversity technique for coded OFDM which can provide full spatially diversity and coding diversity. There are many advantages for DVB-H applications; firstly, CDD can improve BER performance; secondly, CDD can be easily implemented in existing broadcasting system without changing the standards or the receivers and finally, the number of transmit antennas is arbitrary.

Appendix

If $K > N$ and $S(K) \neq 0$, and $s(l) \xrightarrow{DFT} S(l)$.
There are $a_0, a_1, \ldots, a_{N-1}$ constants.
let $A = (a_0, a_1, \ldots, a_{N-1})$.
Then

$$A \cdot \begin{pmatrix} s \\ Ds \\ \ldots \\ D^{N-1}s \end{pmatrix} = 0$$

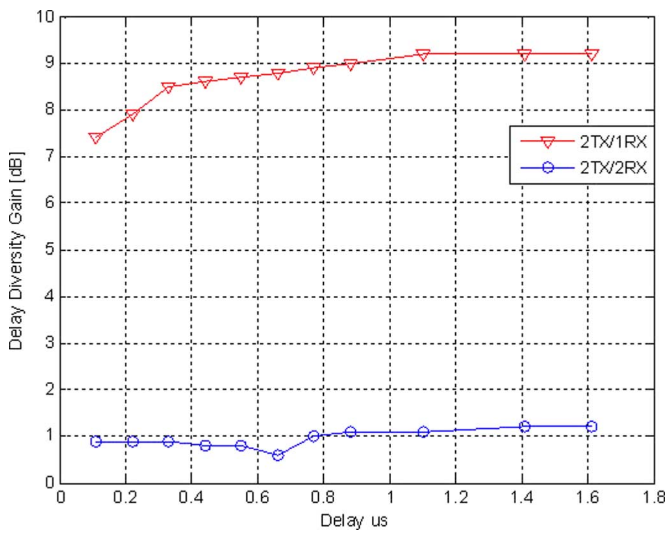

Fig. 13. Delay diversity gain vs. delay for 4K mode, 4QAM, code rate 1/2, bad urban, in L-band (1495 MHz).

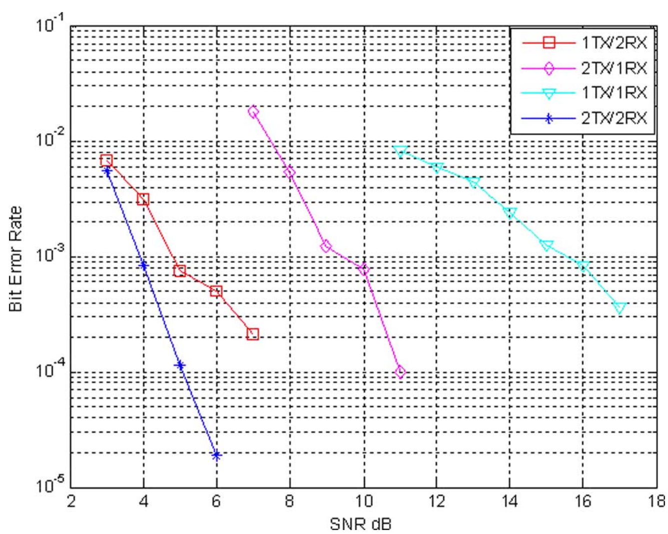

Fig. 14. BER vs. SNR for 4K mode, 4QAM, code rate 1/2, typical urban, in VHF (170.23 MHz).

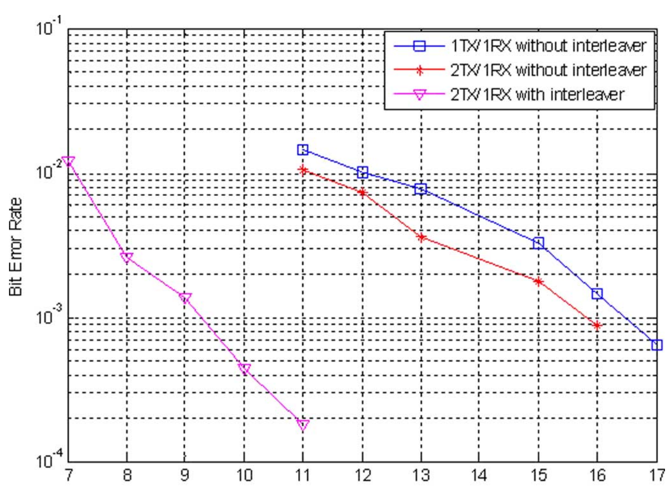

Fig. 15. BER vs. SNR for 4K mode over interleaver, 4QAM, code rate 1/2, typical urban, in UHF (900 MHz).

According to equ-(13), then

$$A \cdot \begin{pmatrix} \sum_{k=0}^{K-1} S(k) e^{j\frac{2\pi}{K}kl} \\ \sum_{k=0}^{K-1} S(k) e^{j\frac{2\pi}{K}k(l-1)} \\ \cdots \\ \sum_{k=0}^{K-1} S(k) e^{j\frac{2\pi}{K}k(l-N)} \end{pmatrix} = 0$$

Because of $S(K) \neq 0$, then $\sum_{n=0}^{N-1} a_n \sum_{k=0}^{K-1} e^{j(2\pi/K)k(l-n)} = 0$.

So $a_0 = 0, a_1 = 0, \ldots, a_{N-1} = 0$.

Then the $s(l)$ and its $N-1$ cyclic shifts symbols are linearly independent.

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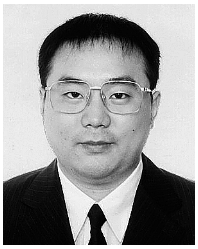

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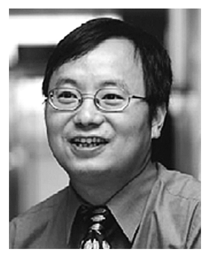

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