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# Linear Block Precoding for OFDM Systems Based on Maximization of Mean Cutoff Rate

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**Abstract**—A new linear block precoding technique is proposed to improve the performance of orthogonal frequency division multiplexing (OFDM) communication systems. The design of our precoder is based on the maximization of the mean cutoff rate and requires only the knowledge of the average relative channel multipath powers and delays at the transmitter. Simulation results show an improved performance of the proposed precoder relative to other known linear block precoding techniques.

**Index Terms**—Cutoff rate, linear block precoding, orthogonal frequency division multiplexing (OFDM) communications.

## I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) is a promising multiuser communication scheme which enables to mitigate multiple-access interference (MAI) by means of providing each user with a nonintersecting fraction of subcarriers [1]. Due to the inverse fast Fourier transform (IFFT) at the transmitter and the fast Fourier transform (FFT) at the receiver, the frequency selective fading channel is converted by OFDM into parallel flat fading channels [2]. This greatly facilitates the equalizer design at the receiver.

However, a well known disadvantage of the OFDM scheme is that, at each subcarrier, the channel may be subject to a deep fading. This makes a reliable detection of the information-bearing symbols at this particular subcarrier very difficult and, as a result, the overall performance of the system may degrade substantially. Thus, the transceiver optimization is required.

A general transceiver optimization framework is discussed in [3]. In application to OFDM systems, a popular recent approach to improve the performance of OFDM systems in fading environments is to use linear block precoding at the transmitter [4]. For example, the minimum mean square error (MMSE) and the minimum bit error rate (MBER) precoders for zero-forcing (ZF) equalization have been proposed in [4] and [5], respectively, and the MBER precoder for MMSE equalization has been studied in

Manuscript received September 14, 2004; revised February 27, 2005. The work of A. B. Gershman was supported by the Wolfgang Paul Award Program of the Alexander von Humboldt Foundation (Germany); Discovery Grants Program of the Natural Sciences and Engineering Research Council (NSERC) of Canada; Premier's Research Excellence Award Program of the Ministry of Energy, Science, and Technology (MEST) of Ontario; and Research Partnerships Program of Communications and Information Technology Ontario (CITO). The results of this paper were presented in part at EUSIPCO'04, Vienna, Austria, September 2004, and ITG/IEEE WSA'05, Duisburg, Germany, April 2005. The associate editor coordinating the review of this manuscript and approving it for publication was Prof. Ioan Tabus.

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Digital Object Identifier 10.1109/TSP.2005.859325

[6]. Another efficient precoding technique based on the channel capacity maximization has been proposed in [7].

Unfortunately, the application of precoders [4]–[7] may be limited by the fact that they require the full channel knowledge at the transmitter. To avoid this drawback, another linear precoder has been designed in [8] based on maximization of the diversity and coding gains. In contrast to the precoders of [4]–[7], the technique of [8] requires only the knowledge of the multipath channel order at the transmitter.

Another MBER based technique that does not require any channel information has been proposed in [9]. However, the class of MBER-optimal channel independent precoders developed in [9] is limited by the case when the MMSE equalization and quadrature phase shift keying (QPSK) modulation are used. Moreover, the performance of MBER precoder with MMSE equalization can be significantly improved by combining it with a water-filling procedure [6]. However, in the latter case the full channel knowledge at the transmitter is required.

In this paper, a new linear precoder is proposed that maximizes the channel mean cutoff rate and requires the knowledge of the average relative channel multipath powers and delays at the transmitter. Our simulations show that the proposed precoder substantially outperforms the approach of [8] and several other linear precoding techniques in terms of BER.

## II. SYSTEM MODEL

For the sake of simplicity and following [4]–[7], let us consider the single-user block transmission system with  $N$  subcarriers. The extension to the multiuser case can be done straightforwardly by allocating a different group of subcarriers to each user [8]. The frequency selective wireless channel between the transmitter and the user is characterized by the path gains  $h_l$  ( $l = 1, \dots, L$ ) and the delays  $\tau_l$  ( $l = 1, \dots, L$ ), where all path gains are assumed to be independent (but not necessarily identically distributed) zero-mean complex Gaussian random variables.

Employing the cyclic prefix (CP)-based OFDM transmissions, we have the following relationship [4]–[7]:

$$\mathbf{r}(t) = \sqrt{E_s} \mathbf{D} \mathbf{y}(t) + \mathbf{v}(t) \quad (1)$$

where  $t$  is the block index,  $\mathbf{r}(t) = [r(tN), \dots, r(tN+N-1)]^T$  is the  $N \times 1$  vector of the received symbols after the FFT operation,  $E_s$  is the transmitted symbol power,  $\mathbf{y}(t) = [y(tN), \dots, y(tN+N-1)]^T$  is the  $N \times 1$  vector of the transmitted symbols without CP,  $\mathbf{v}(t) = [v(tN), \dots, v(tN+N-1)]^T$  is the  $N \times 1$  vector of white complex Gaussian noise with the covariance matrix

$E\{\mathbf{v}(t)\mathbf{v}^H(t)\} = \sigma_v^2 \mathbf{I}_N$ ,  $E\{\cdot\}$  denotes the statistical expectation,  $\sigma_v^2$  is the noise power,  $\mathbf{I}_N$  is the  $N \times N$  identity matrix,  $\mathbf{D}$  is the  $N \times N$  diagonal matrix whose  $(n, n)$ th element represents the channel impulse response of the  $n$ th subcarrier<sup>1</sup> and is given by [10]

$$[\mathbf{D}]_{n,n} = \frac{1}{\sqrt{N}} \sum_{l=1}^L h_l \exp\left(-\frac{j2\pi n\tau_l}{NT}\right) \quad (2)$$

where  $T$  is the sampling interval, and  $j \triangleq \sqrt{-1}$ .

If linear block precoding is used at the transmitter, then  $\mathbf{y}(t) = \mathbf{T}\mathbf{s}(t)$  where  $\mathbf{s}(t) = [s(tP), \dots, s(tP + P - 1)]^T$  is the  $P \times 1$  vector of the information-bearing symbols, and  $\mathbf{T}$  is the  $N \times P$  precoding matrix [4]. Below, we assume that  $P = N$  because in this case, the data rate is not sacrificed [2]. Then, (1) can be written as

$$\mathbf{r}(t) = \sqrt{E_s} \mathbf{D} \mathbf{T} \mathbf{s}(t) + \mathbf{v}(t). \quad (3)$$

### III. PROPOSED LINEAR BLOCK PRECODER

The channel cutoff rate  $R_0$  is a lower bound on the Shannon channel capacity, beyond which the sequential decoding becomes impractical [11], [12]. It also specifies an upper bound on the error rate of the optimal ML decoder and has been frequently used as a practical coding limit because it can be calculated in a simpler way than the channel capacity [12]. Therefore, the cutoff rate appears to be a proper criterion for the design of linear block precoders. Note that it has been previously used as a performance metric for OFDM systems [13], [14], and as a design criterion for transmitter optimization in multiple-input multiple-output (MIMO) channels [15].

We assume that a discrete constellation is used at the transmitter, the full channel knowledge is available at the receiver, and the maximum likelihood (ML) technique is used to detect the symbols  $\mathbf{s}(t)$  from the received data  $\mathbf{r}(t)$ . The conditional probability density function of the received data can be written as

$$f(\mathbf{r}|\mathbf{s}^{(i)}, \mathbf{T}, \mathbf{D}) = \frac{1}{(\pi\sigma_v^2)^N} \exp\left(-\frac{\|\mathbf{r} - \sqrt{E_s} \mathbf{D} \mathbf{T} \mathbf{s}^{(i)}\|^2}{\sigma_v^2}\right) \quad (4)$$

where  $\mathbf{s}^{(i)}$  is the  $i$ th member of the transmit vector constellation,  $\|\cdot\|$  denotes the Frobenius matrix norm or the Euclidean vector norm, and the dependence on  $t$  is omitted in the interest of notational brevity. To simplify the notation further, let us denote

<sup>1</sup>Note that the frequency domain channel response is often expressed through the FFT, but in this case the true multipath channel taps need to be converted to equivalent taps which have delays that are integer multiples of the sampling period. Therefore, following [10] we use the discrete Fourier transform (DFT) instead of FFT to calculate the channel impulse response in (2).

$f(\mathbf{r}|\mathbf{s}^{(i)}, \mathbf{T}, \mathbf{D})$  as  $f(i)$ . The mean cutoff rate can be calculated as [11, p. 361]

$$R_0 = -\log E_{\mathbf{D}} \left\{ \int_{\mathbf{r}} \left[ \frac{1}{M^N} \sum_{i=1}^{M^N} \sqrt{f(i)} \right]^2 d\mathbf{r} \right\} \\ = -\log \left[ \frac{1}{M^N} + \frac{1}{M^{2N}} \sum_{i=1}^{M^N} \sum_{\substack{l=1 \\ l \neq i}}^{M^N} E_{\mathbf{D}} \left\{ \int_{\mathbf{r}} \sqrt{f(i)f(l)} d\mathbf{r} \right\} \right] \quad (5)$$

where  $M$  is the constellation size. Inserting (4) into (5) we obtain the expression for the mean cutoff rate shown in (6) at the bottom of the page. Using the results of [17], the expectation of exponential quadratic form in (6) can be written as

$$E_{\mathbf{D}} \left\{ \exp\left(-\frac{E_s \|\mathbf{D} \mathbf{T} (\mathbf{s}^{(i)} - \mathbf{s}^{(l)})\|^2}{4\sigma_v^2}\right) \right\} \\ = \prod_{k=1}^{r\{\mathbf{E}_{i,l}\}} \left(1 + \frac{E_s}{4\sigma_v^2} \lambda_k\right)^{-1} \quad (7)$$

where

$$\mathbf{E}_{i,l} \triangleq E_{\mathbf{D}} \{\mathbf{D} \mathbf{T} \mathbf{e}_{i,l} \mathbf{e}_{i,l}^H \mathbf{T}^H \mathbf{D}^H\}, \quad \mathbf{e}_{i,l} \triangleq \mathbf{s}^{(i)} - \mathbf{s}^{(l)}. \quad (8)$$

$\lambda_k$  is the  $k$ th nonzero eigenvalue of the matrix  $\mathbf{E}_{i,l}$ , and  $r\{\cdot\}$  denotes the matrix rank. Substituting (7) in (6), after straightforward manipulations we obtain

$$R_0 = -\log \left[ \frac{1}{M^N} + \frac{1}{M^{2N}} \sum_{i=1}^{M^N} \sum_{\substack{l=1 \\ l \neq i}}^{M^N} \prod_{k=1}^{r\{\mathbf{E}_{i,l}\}} \left(1 + \frac{E_s}{4\sigma_v^2} \lambda_k\right)^{-1} \right]. \quad (9)$$

It is worth noting that the expression (9) for the mean cutoff rate is directly related to the expression for the Chernoff bound on the pairwise error probability (PEP). In particular, the second term under the logarithm in (9) can be seen as an average of the Chernoff bounds on PEP for all distinct pairs of symbols. In other words, the maximization of mean cutoff rate is equivalent to the minimization of averaged PEP. This observation provides further motivation to choose the mean cutoff rate as a criterion for precoder design.

It is well known that the precoder based on PEP minimization also provides maximum diversity gain [16]. Therefore, the maximization of the mean cutoff rate will achieve the maximum diversity gain under the following condition [16]:

$$\left| [\mathbf{T}]_n (\mathbf{s}^{(i)} - \mathbf{s}^{(l)}) \right| \neq 0, \quad n=1, \dots, N, \quad i, l=1, \dots, M^N \quad (10)$$

where  $[\mathbf{T}]_n$  is the  $n$ th row of  $\mathbf{T}$ .

$$R_0 = -\log \left[ \frac{1}{M^N} + \frac{1}{M^{2N}} \sum_{i=1}^{M^N} \sum_{\substack{l=1 \\ l \neq i}}^{M^N} E_{\mathbf{D}} \left\{ \exp\left(-\frac{E_s \|\mathbf{D} \mathbf{T} (\mathbf{s}^{(i)} - \mathbf{s}^{(l)})\|^2}{4\sigma_v^2}\right) \right\} \right]. \quad (6)$$

To compute the matrix  $\mathbf{E}_{i,l}$  explicitly, let us introduce the vector  $\mathbf{d} \triangleq [[\mathbf{D}]_{1,1}, \dots, [\mathbf{D}]_{N,N}]^T$ . Then,  $\mathbf{E}_{i,l}$  can be rewritten as

$$\mathbf{E}_{i,l} = \mathbf{R}_d \odot (\mathbf{T} \mathbf{e}_{i,l} \mathbf{e}_{i,l}^H \mathbf{T}^H) \quad (11)$$

where  $\odot$  stands for the Schur-Hadamard matrix product and  $\mathbf{R}_d \triangleq \mathbb{E}_d\{\mathbf{d}\mathbf{d}^H\}$ . The  $(n, k)$ th entry of  $\mathbf{R}_d$  is given by

$$[\mathbf{R}_d]_{n,k} = \frac{1}{N} \sum_{l=1}^L P_l \exp\left(-\frac{j2\pi(n-k)\tau_l}{NT}\right) \quad (12)$$

where  $P_l$  is the average power of the  $l$ th path relative to the first path.

Our task now is to design the precoding matrix  $\mathbf{T}$  which maximizes  $R_0$  in (9) subject to the power constraint  $\|\mathbf{T}\| = \sqrt{N}$ . This problem does not have any analytical solution, but it can be solved by using either algebraic number-theoretic techniques or by computer search over compact parameterizations of unitary matrices [8], [16]. In this paper, we obtain  $\mathbf{T}$  through computer search over the unitary<sup>2</sup> parameterization expressed via Givens rotation matrices. For the details of this technique, see, for example, [16]. Provided that each user occupies a moderate number of subcarriers (not more than 3 subcarriers per user), and since the precoding matrices can be designed for each user independently, we can conclude that the total number of real optimization parameters for the particular user is  $N(N-1) \leq 6$ . If the number of optimization parameters is small, full search is computationally feasible, and, thus, the design of our precoder becomes practically feasible as well.

It can be seen from (9)–(12) that only the knowledge of the average relative channel powers and delays is required at the transmitter for the design of our linear precoder. Although the channel state variations can be very fast due to small-scale fading, the power and multipath delay variations are typically much slower [1]. Therefore, a low-rate feedback can be used to convey this information to the transmitter.

#### IV. SIMULATION RESULTS

In this section, we investigate the performance of the proposed precoder in multipath indoor and outdoor channels. As an example of a Rayleigh fading outdoor channel, we choose the ETSI “Vehicular A” channel environment which has been defined for the evaluation of UMTS radio interface proposals [18]. The multipath time delays and the variances of the multipath gains of the “Vehicular A” channel are shown in Table I. Correspondingly, as an example of a multipath Rayleigh fading indoor channel we choose the HIPERLAN/2 “Model A” channel, which represents a typical office environment [19]. The Doppler frequencies for these two channels are set to be equal to 100 Hz and 50 Hz, respectively. The multipath time delays and the variances of the multipath gains of the HIPERLAN/2 “Model A” channel are shown in Table II.

Throughout the simulations, a multiuser block transmission system with 64 subcarriers is assumed. All subcarriers are al-

<sup>2</sup>Note that unitary precoders have the advantage that they do not alter the Euclidean distance between the entries of any block of information-bearing symbols [8].

TABLE I  
CHARACTERISTICS OF THE ETSI “VEHICULAR A” CHANNEL ENVIRONMENT

Tap	Time Delays (T)	Average Power (dB)
1	0	0
2	1.55	-1
3	3.55	-9
4	5.45	-10
5	8.65	-15
6	12.55	-20

TABLE II  
CHARACTERISTICS OF THE HIPERLAN/2 “MODEL A” CHANNEL ENVIRONMENT

Tap	Time Delays (T)	Average Power (dB)
1	0	0
2	0.2	-0.9
3	0.4	-1.7
4	0.6	-2.6
5	0.8	-3.5
6	1	-4.3
7	1.2	-5.2
8	1.4	-6.1
9	1.6	-6.9
10	1.8	-7.8
11	2.2	-4.7
12	2.8	-7.3
13	3.4	-9.9
14	4	-12.5
15	4.8	-13.7
16	5.8	-18.0
17	6.8	-22.4
18	7.8	-26.7

located among the users and interleaved [8] such that the subcarriers assigned to the same user are as less correlated to each other as possible. Each user is provided with  $N = 3$  subcarriers and the BPSK modulation is used.

In our first example, different precoding techniques are compared to each other when no channel coding is used. This enables us to study the net effect of precoding on the performance of OFDM systems. Six different techniques are compared: the approach where no precoding is used, the MMSE-ZF precoder of [4], the MBER-ZF precoder of [5], the MBER-MMSE precoder of [6], the Vandermonde precoder of [8], and the proposed precoder. The precoders of [4], [5], and [6] are assumed to utilize the full channel knowledge at the transmitter, whereas the precoder of [8] and the proposed precoder use only the multipath channel order and the average relative channel powers and delays, respectively. It is also important to stress that without any precoding,  $\mathbf{T} = \mathbf{I}$  and the detection of each information-bearing symbol is decoupled from the detection of any other symbols. Thus, provided that a constant-modulus constellation is used, the ML and MMSE symbol detectors are equivalent in this case.

Fig. 1 displays the channel mean cutoff rate of the aforementioned precoding schemes versus the signal-to-noise ratio (SNR) for the ETSI “Vehicular A” channel environment. It can be seen from this figure that, as expected, the proposed linear precoder has the highest mean cutoff rate among all the techniques tested.

Fig. 2 compares the BER performances of the same techniques with different symbol detectors for the ETSI “Vehic-

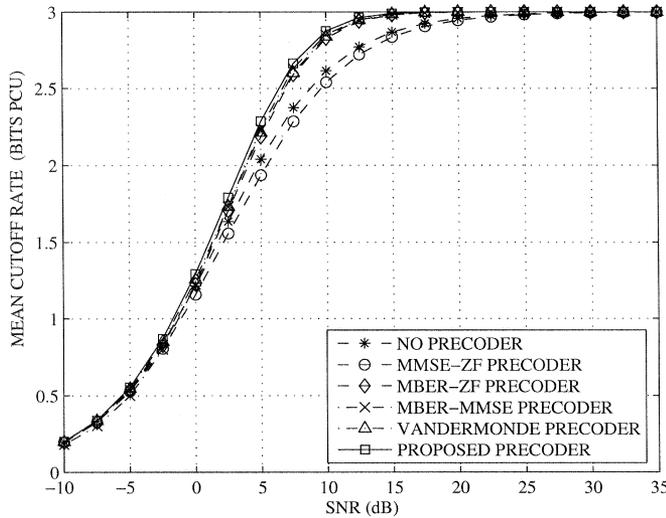


Fig. 1. Cutoff rate versus SNR. First example with the ETSI "Vehicular A" channel environment.

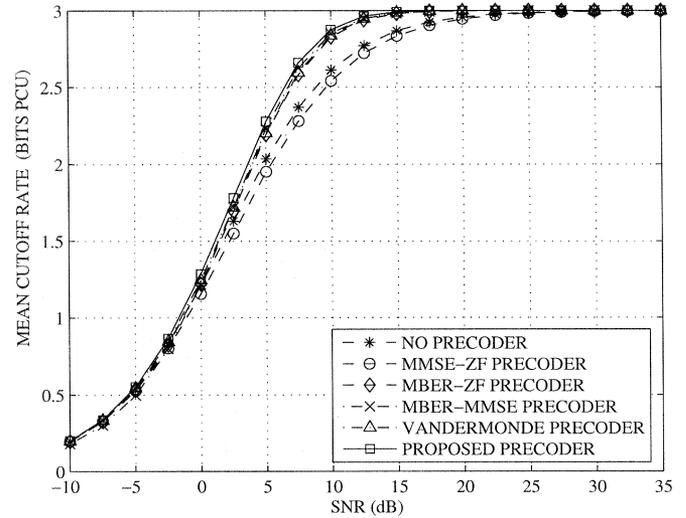


Fig. 3. Cutoff rate versus SNR. First example with the HIPERLAN/2 "Model A" channel environment.

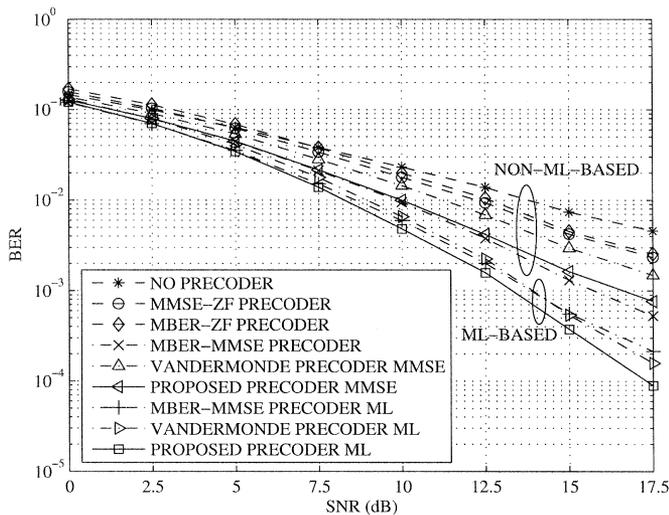


Fig. 2. BER versus SNR. First example with the ETSI "Vehicular A" channel environment.

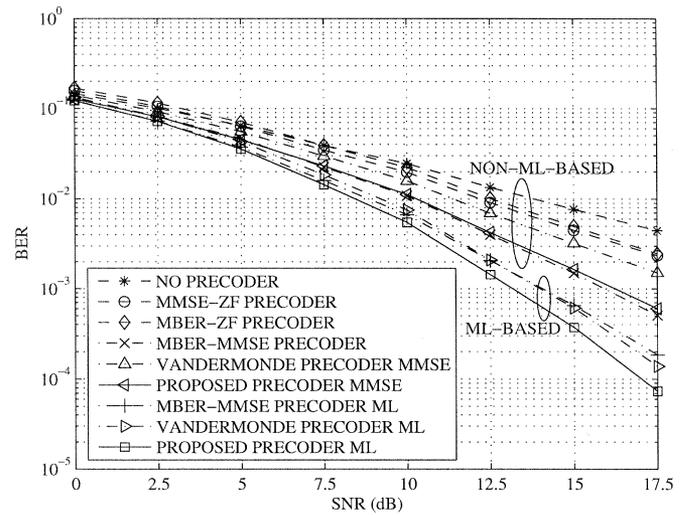


Fig. 4. BER versus SNR. First example with the HIPERLAN/2 "Model A" channel environment.

ular A" channel environment. In this figure, the performances of the proposed, MBER-MMSE, and Vandermonde precoders are displayed both in the cases when the ML and MMSE symbol detectors are used. Additionally, the BER performances of the MMSE-ZF and MBER-ZF precoders are displayed along with the BER of the standard approach where no precoding is used. All results are averaged over 1000 simulation runs.

Fig. 3 shows the channel mean cutoff rate of all aforementioned precoders versus SNR for the HIPERLAN/2 "Model A" channel environment. As in the previous case, in this indoor channel scenario the proposed linear precoder has the highest  $R_0$ .

Finally, the BER performances of all tested precoders are compared in Fig. 4 in the HIPERLAN/2 "Model A" channel case.

We can observe that our linear precoder substantially outperforms all other techniques tested in terms of BER for both the ETSI "Vehicular A" and HIPERLAN/2 "Model A" channel environments. Interestingly, this conclusion is true when the

ML-based as well as non-ML (MMSE) receivers are used, with the only exception for the MBER-MMSE precoder. In particular, the performance of the latter precoder is comparable to the performance of our precoder used with the MMSE receiver. This fact demonstrates that although the mean cutoff rate-based precoder has been proposed for the ML-based symbol detector, it also provides a good performance when applied with the simpler MMSE symbol detector.

In our second example, we consider an OFDM system with combined channel coding and linear precoding that is similar to one described in [20]. In this example, a sequence of information bits is first encoded by a rate 1/2 convolutional channel code (CC) with the generator (133, 171) that is similar to one used in the HIPERLAN/2 standard [19], [20]. The coded bits then pass through a random interleaver  $\Pi_1$  of a size corresponding to 256 OFDM symbols. The bits at the output of  $\Pi_1$  are mapped to the BPSK constellation symbols. After constellation mapping, successive blocks of  $N$  symbols  $\mathbf{s}(t) = [s(tN), \dots, s(tN + N - 1)]^T$  are linearly precoded by the matrix  $\mathbf{T}$ . As a result,

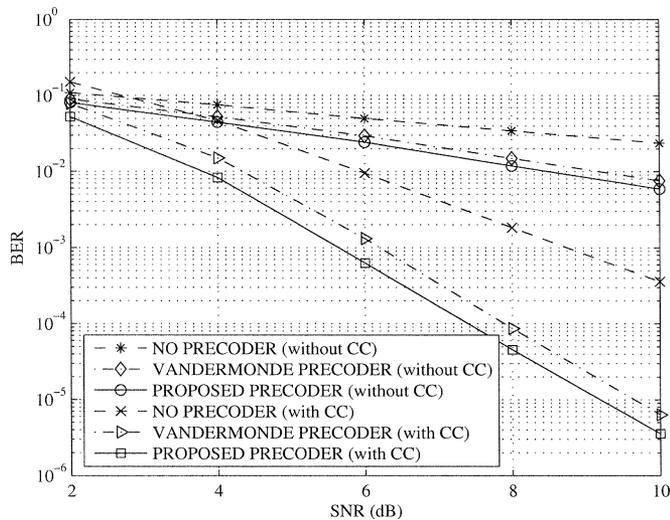


Fig. 5. BER versus SNR. Second example with the ETSI "Vehicular A" channel environment.

the precoded block  $\mathbf{y}(t) = \mathbf{T}s(t)$  is obtained. The precoded symbols are interleaved by a second interleaver  $\Pi_2$ , which is needed to decorrelate the subcarrier channel gains. The output of  $\Pi_2$  passes through the OFDM modulator, which includes serial-to-parallel conversion, IFFT, and CP insertion. The resulting output is serialized for transmission. At the receiver, after OFDM demodulation<sup>3</sup> and deinterleaving  $\Pi_2^{-1}$ , the signal is decoded by the ML decoder.

In Figs. 5 and 6, the performances of the following techniques are compared: our precoder with combined coding-precoding; Vandermonde precoder with combined coding-precoding; entirely channel coding based OFDM technique; entirely precoding based OFDM technique; and OFDM technique without coding-precoding. The comparison is given for both the ETSI "Vehicular A" channel (Fig. 5) and HIPERLAN/2 "Model A" channel (Fig. 6).

As can be seen from Figs. 5 and 6, the OFDM system with combined coding-precoding shows the best performance among the techniques tested for both the channel environments used. Moreover, the OFDM system with combined coding-precoding that uses the proposed precoder outperforms the OFDM system with combined coding-precoding that uses the Vandermonde precoder. Note that the Vandermonde precoder has been chosen because it shows the best performance among other existing precoders tested in the first example. However, the performance improvement for combined coding-precoding techniques is achieved at the price of higher decoding complexity and longer decoding delay.

## V. CONCLUSION

A new linear precoder for block OFDM transmissions has been proposed. Our precoder is based on the maximization of the channel mean cutoff rate and requires only the knowledge of the average relative channel multipath powers and delays at the transmitter. Simulation results show substantial performance improvements achieved by the proposed precoding technique

<sup>3</sup>In this simulation, hard-decision demodulation is used.

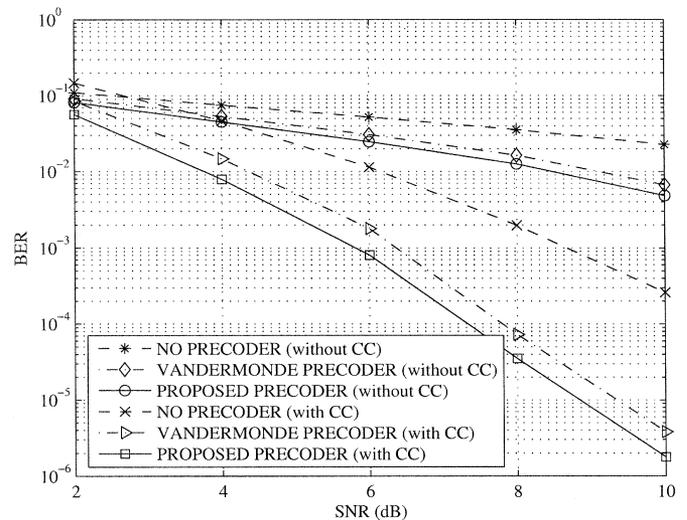


Fig. 6. BER versus SNR. Second example with the HIPERLAN/2 "Model A" channel environment.

relative to the existing linear block precoders. The proposed precoding technique can be readily generalized to the MIMO case.

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