Design of Unequal Error Protection for MIMO-OFDM Systems

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***Abstract*— Most multimedia source data exhibit unequal bit error sensitivity. Therefore, it is desirable to design a system where important data should be recovered with high priority. In this paper, we present unequal error protection (UEP) schemes, which exploit differences in bit error protection levels of M-ary QAM symbols, in order to support multimedia transmission in orthogonal frequency division multiplexing (OFDM) systems over frequency selective fading channels. We introduce an UEP scheme which significantly improves the performance with multiple transmit antennas. Also, we propose a receiver based on two stage Maximum Likelihood detection (MLD) schemes which can approach the performance of a full search MLD receiver with much reduced computational complexity. Simulation results show that the proposed schemes achieve a significant performance gain over the conventional equal error protection (EEP) scheme.**

I. Introduction

Next generation wireless communication systems are expected to provide reliable transmission over fading channels. Bit streams from most real world source coding algorithms for speech, audio, images and video exhibit unequal bit error sensitivity for different bits. In the conventional system, error protection needs to be designed for the most sensitive bits, because those bits determine the overall detection performance. In such equal error protection (EEP) systems, this results in the waste of resource for the protection of the least sensitive bits, since they are assigned the same protection level as the most sensitive bits.

On the other hand, the unequal error protection (UEP) receiver matches the protection level according to the system requirement, and thus can save the system resources. Since not all the bits in the bit stream require the same error protection for multimedia data, we can improve the system performance and the overall efficiency by applying UEP at no extra bandwidth requirement. The UEP schemes divide the source data into two or more groups and assign each group different protection levels. The most important bits, defined as base information, are protected with more redundancy, while the less important bits, defined as refinement information, are restored with less protection. For example, the standard services of the satellite wideband-code division multiple access (SW-CDMA) system have a target BER of 10^{-3}. It has been determined in [1] that, for good voice quality, the most sensitive bits of a G.723.1 frame cannot tolerate a BER $> 5 \times 10^{-5}$, while the remaining bits can tolerate a BER as high as 10^{-3}.

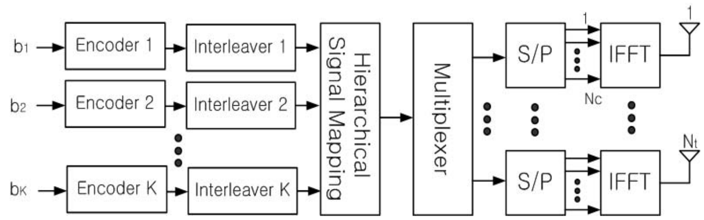

Fig. 1. Transmitter structure for the UEP

Several UEP schemes have been employed in many digital communication and broadcasting systems to jointing optimize the transmission schemes with the digital source. An UEP scheme based on rate compatible punctured convolutional (RCPC) codes has been studied in the literature [2]. The design of an error correction scheme usually consists of selecting a fixed channel code with the same error correction capability for all the data to be transmitted. A fixed code is constructed for the worst case of average channel/source conditions. To make the best use of the limited resources, it is necessary to match the error protection level provided by the channel code to the error sensitivity. The RCPC codes meet such requirements of being capable of providing a flexible UEP [2]. The punctured bits in the RCPC codes represent savings in bit rate (or bandwidth) without loss of quality. Alternatively, the gain of UEP over EEP can be converted into a power gain. Other previous approaches include nonstandard signal set signal partitioning and multistage decoding (MSD) techniques which provide good UEP capabilities [3][4]. The coding scheme is designed in such a way that the most important information bits result in a better error rate than other information bits. A hierarchical signal constellation is also a natural strategy to provide UEP capability [5].

A multi-input multi-output (MIMO) system has attracted much attention to provide high-speed data for next generation communication systems with multiple antennas both at transmitter and receiver side. Also, orthogonal frequency division multiplexing (OFDM) can be employed to cope with frequency selectivity caused by multipath fadings. In this paper, we propose an UEP scheme for MIMO-OFDM systems. The proposed UEP schemes employ a hierarchical signal mapping based on both minimum mean-square error (MMSE) and maximum likelihood (ML) detectors. We show that a significant

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performance gain for base information is achieved, compared with an EEP detection structure. We also investigate the performance of base information and refinement information in the nonuniform signal constellation.

While the maximum likelihood detection (MLD) exhibits the optimum performance, it is often too complicated to be applied in practice. Thus, much interest has been focused on reducing the complexity of the MLD method. In this paper, we propose an efficient MLD receiver operating in two stages. The proposed detector estimates the base bits using an MMSE detector first, then the remaining refinement bits are processed by MLD. Simulation results show that the proposed two stage MLD scheme can achieve a near ML performance with reduced complexity.

The organization of the paper is as follow: In Section II, we present the system overview. In Section III, we propose a new UEP scheme and a two stage MLD receiver. Section IV shows the simulation results for MIMO-OFDM. Finally, the paper is terminated with discussion in Section V.

II. System Overview

In this section, we consider an OFDM system with N_t transmit and N_r receive antennas. Figure 1 shows the transmitter configuration of the UEP system, where N_c indicates the number of subchannels. Here, K different bit sensitivity groups are transmitted through each parallel transmission substream with independent encoder and a bit-interleaver. A key function block in this structure is the hierarchical signal mapper. The hierarchical signal mapping offers different levels of protection to the transmitted bits in a message symbol according to their priorities. That is, information bits are distributed to different bit-sensitivity groups.

For a channel model, we make the following assumptions. Considering the time domain channel impulse response between the ith transmit and jth receiver antenna, the frequency selective channel can be modeled as

$$h^{i,j}(\tau) = \sum_{n=1}^{L} \bar{h}^{i,j}(n)\delta(\tau - \tau_n)$$

where the channel coefficients $\bar{h}^{i,j}(n)$ are independent complex Gaussian with zero mean, $\delta(\cdot)$ is the Dirac delta function, and L denotes the number of channel taps. It follows that the chnnel frequency response can be expressed by

$$H_k^{i,j} = \sum_{n=1}^{L} \bar{h}^{i,j}(n) e^{-j2\pi k \tau_n / N_c T_s} = \bar{\mathbf{h}}_{i,j}\mathbf{w}_k$$

where T_s represents the sampling period. Also we denote $\bar{\mathbf{h}}_{i,j} = [\bar{h}^{i,j}(1)\bar{h}^{i,j}(2)\cdots\bar{h}^{i,j}(L)]$ and $\mathbf{w}_k = [e^{-j2\pi k\tau_1/N_cT_s} e^{-j2\pi k\tau_2/N_cT_s} \ldots e^{-j2\pi k\tau_L/N_cT_s}]^T$ where T indicate transpose. Note that $|H_k^{i,j}|$ is Rayleigh distributed.

Let us define the N_t-dimensional complex transmitted signal vector $\mathbf{x}_k$, and the N_r-dimensional complex received signal vector $\mathbf{y}_k$. Then the received signal at the kth subcarrier can be written as

$$\mathbf{y}_k = \mathbf{H}_k\mathbf{x}_k + \mathbf{n}_k$$

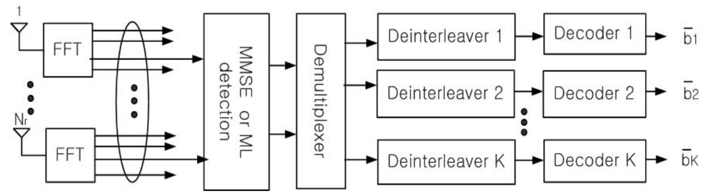

Fig. 2. Receiver structure for the proposed UEP

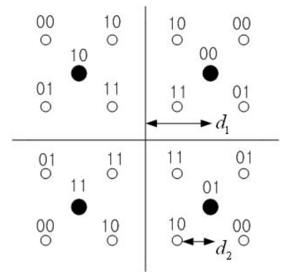

Fig. 3. The hierarchical signal mapping for 16QAM.

where

$$\mathbf{H}_k = \begin{bmatrix} H_k^{1,1} & \cdots & H_k^{N_t,1} \\ \vdots & \ddots & \vdots \\ H_k^{1,N_r} & \cdots & H_k^{N_t,N_r} \end{bmatrix}, \mathbf{n}_k = \begin{bmatrix} n_{1,k} \\ \vdots \\ n_{N_r,k} \end{bmatrix}.$$

Here, we assume that the transmitted signal power is distributed equally over N_t transmit antennas with the same variance σ_s^2. Thus the covariance matrix of $\mathbf{x}_k$ equals $E[\mathbf{x}_k\mathbf{x}_k^\dagger] = \sigma_s^2\mathbf{I}_{N_t}$ where $E[\cdot]$ and $(\cdot)^\dagger$ indicate expectation and the complex-conjugate transpose, respectively, and $\mathbf{I}_{N_t}$ denotes an identity matrix of size N_t. The additive noise terms in $\mathbf{n}_k$ are independent and identically-distributed complex zero mean Gaussian with variance σ_n^2.

III. The Proposed Structure

In this section, we first propose a UEP scheme with the hierarchical signal mapping. Then the two stage MLD scheme is also presented.

A. The Proposed UEP

Figure 2 exhibits the receiver structure for the proposed UEP. The receiver is assumed to have perfect knowledge of the channel state information. After the FFT demodulation, the received signals can be processed either by MMSE filtering or ML detection. The ML detection achieves a better performance compared to the MMSE filtering at the expense of increased complexity. After passing the demultiplexer, each parallel substream is independently bit-deinterleaved and decoded.

As an example, for 16QAM, we can consider two incoming data streams (K=2) where b_1 and b_2 denote the base information and the refinement information, respectively. In the base information stream, two bits out of four are assigned at the most important bit positions, while in the refinement information stream the rest two bits are assigned at the least important bit positions. In Figure 3, the hierarchical nonuniform signal constellation for 16QAM is presented [6][7]. The hierarchical signal constellation is designed such that the base bits are selected as one of four fictitious symbols (black circles), whereas the refinement bits are treated as one of the four symbols (empty circles), surrounding the selected fictitious symbol. In other words, the constellation consists of "clouds" of subconstellations that account for the refinement information, while the base information is represented by the position of the clouds. Thus, the incorrect detection of base information gives rise to the decision error for the refinement information. In Figure 3, $2d_1$ represents the distance between the two fictitious symbols, whereas $2d_2$ corresponds to the distance between two neighboring symbols within subconstellation. By controlling the relative distance between d_1 and d_2 with a parameter $\lambda = \frac{d_2}{d_1}$, we can change relative protection levels for the two bit streams. When λ gets smaller, the reliability of the base information increases while that of the refinement information decreases. For $\lambda = 0$, the signal constellation becomes a uniform 4QAM with no refinement information. Also, for $\lambda = \frac{1}{2}$, the signal constellation becomes a uniform 16QAM where both the base information and the refinement information have equal priority.

B. The Proposed Two Stage MLD

As will be shown in the simulation section, the base information performance of the proposed UEP based on MMSE is better than the performance of EEP based on MLD. This motivates us to consider the receiver structure which consists of the two stage MLD as shown in Figure 4. In the first stage, we process the base information based on MMSE filtering with low-complexity. Using the MMSE criterion, the equalizer matrix $\mathbf{G}_k$ is formulated to minimize the mean square values of the error. First denoting $\alpha = \frac{\sigma_n^2}{\sigma_s^2}$, the equalizer output $\mathbf{z}_k = \mathbf{G}_k\mathbf{y}_k$ is obtained with the equalizer $\mathbf{G}_k$ as

$$\mathbf{G}_k = \left(\mathbf{H}_k^\dagger\mathbf{H}_k + \alpha\mathbf{I}_{N_t}\right)^{-1}\mathbf{H}_k.$$

Defining $\mathbf{g}_t$ as the tth row of $\mathbf{G}_k$, the tth element of the equalizer output is given as

$$\begin{aligned} z_t &= \mathbf{g}_t\mathbf{H}_k\mathbf{x}_k + \mathbf{g}_t\mathbf{n}_k \\ &= \mathbf{g}_t\mathbf{h}_t x_t + \sum_{i=1, i\neq t}^{N_t}\mathbf{g}_t\mathbf{h}_i x_i + \mathbf{g}_t\mathbf{n}_k \\ &= \beta x_t + \omega \end{aligned}$$

where β and ω are defined as $\mathbf{g}_t\mathbf{h}_t$ and $\sum_{i=1,i\neq t}^{N_t}\mathbf{g}_t\mathbf{h}_i x_i + \mathbf{g}_t\mathbf{n}_k$, respectively. For analytic conveniences, we assume that the terms in ω make complex Gaussian distribution. Since those terms in ω are assumed to be independent with each other, the variance of ω is computed as

$$\sigma_w^2 = \sum_{i=1,i\neq t}^{N_t}\|\mathbf{g}_t\mathbf{h}_i\|^2\sigma_s^2 + \|\mathbf{g}_t\|^2\sigma_n^2.$$

After a biased term is properly scaled, the input of the unbiased demapper can be written as

$$\hat{x}_t = z_t/\beta = x_t + v$$

where v is the complex noise with variance $\sigma_v^2 = \sigma_w^2/\|\beta\|^2$.

In what follows, we briefly describe the Log-Likelihood Ratio (LLR) computation for soft bit information [8]. Let $\mathcal{S}_t^b$ be a set of constellation symbols and denote x_t^b as an element of the $\mathcal{S}_t^b$. We represent the ith bit of x_t^b by $b_{k,t}^{b,i}$ and define two mutually exclusive subsets of $\mathcal{S}_t^b$ as $\mathcal{S}_{t,0}^b = \{x_t^b | b_{k,t}^{b,i} = 0\}$ and $\mathcal{S}_{t,1}^b = \{x_t^b | b_{k,t}^{b,i} = 1\}$ for $i = 1, 2\cdots, m_b$ where m_b is the number of base bits. Note that $|\mathcal{S}_t^b|$ is equal to the constellation size M. Then, the LLR of base information $b_{k,t}^{b,i}$ [8] can be represented by

$$LLR(b_{k,t}^{b,i}) = \log\frac{\sum_{x_t^b\in\mathcal{S}_{t,0}^b}\exp\left(-\frac{\|\hat{x}_t - x_t^b\|^2}{\sigma_v^2}\right)}{\sum_{x_t^b\in\mathcal{S}_{t,1}^b}\exp\left(-\frac{\|\hat{x}_t - x_t^b\|^2}{\sigma_v^2}\right)}. \tag{1}$$

When the base information bits are estimated using the MMSE filtering alone, the detection performance becomes poor even if the refinement information bits are detected using ML. This indicates that the overall performance is mostly determined by the accuracy of the base information bit estimation. Thus, in order to improve the base bits detection performance, we employ a decision feedback method [9] using decoder output to enhance the accuracy of base bits at the first stage. After the base information estimates from the first stage are passed to the decoder, the decoder outputs are re-encoded. Then, the MLD utilizes the re-encoded base bits for the second stage processing. In the second stage, the refinement bits are searched using MLD in a much smaller coset.

Let us define $b_{t,k}^{r,i}$ $(i = 1, 2\cdots, m_r)$ for refinement information similarly as in $b_{t,k}^{b,i}$ above. m_r represents the number of refinement information bits. Denote the set $\mathbf{S}_{t,d}^r, d = 1$ or 0, as a set of all symbol vectors with $b_{t,k}^{r,i} = d$. Here the base bits in $\mathbf{x}_k$ are fixed with the decisions made from the first stage. The number of elements in such a set is $2^{N_t m_r - 1}$. Then the LLR for refinement information is given as

$$LLR(b_{t,k}^{r,i}) = \log\frac{\sum_{\mathbf{x}_k\in\mathbf{S}_{t,0}^r}\exp\left(-\frac{\|\mathbf{y}_k - \mathbf{H}_k\mathbf{x}_k\|^2}{\sigma_n^2}\right)}{\sum_{\mathbf{x}_k\in\mathbf{S}_{t,1}^r}\exp\left(-\frac{\|\mathbf{y}_k - \mathbf{H}_k\mathbf{x}_k\|^2}{\sigma_n^2}\right)}.$$

We summarize the proposed two stage MLD procedure in each step. For 16QAM case with $m_b = m_r = 2$ as in Figure 3.

- In the first stage, the LLR values for base information bits are computed using (1) based on MMSE filtering.
- The computed LLR values are passed to the decoder for the base bits.
- The decoder outputs are re-encoded and transferred as base information to the second stage for the MLD detection.
- In the second stage, by fixed the base information bits in the hierarchical signal mapping, the LLR values for the refinement information bits are computed with MLD.
- Finally, these LLR values are passed to decoder for refinement bits.

For example, if the re-encoded bits are 00, then we perform a local MLD for four symbols surrounding 00 in Figure 3 to

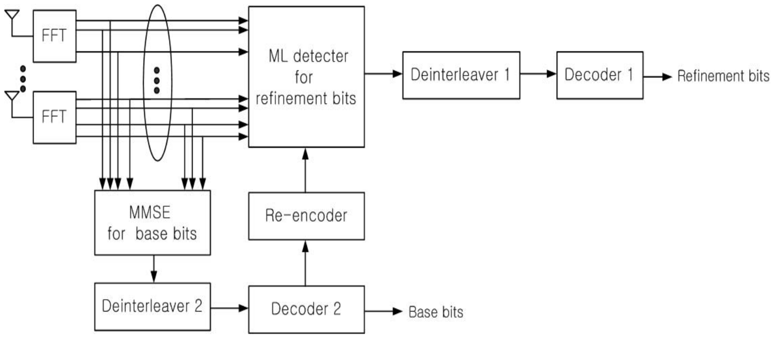

Fig. 4. Receiver structure for the two stage MLD

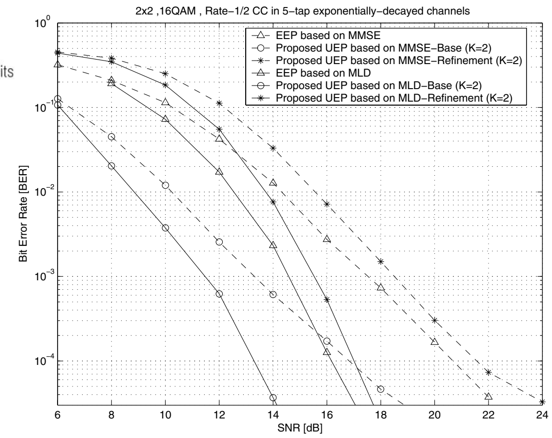

Fig. 5. Performance of the proposed UEP for 16QAM

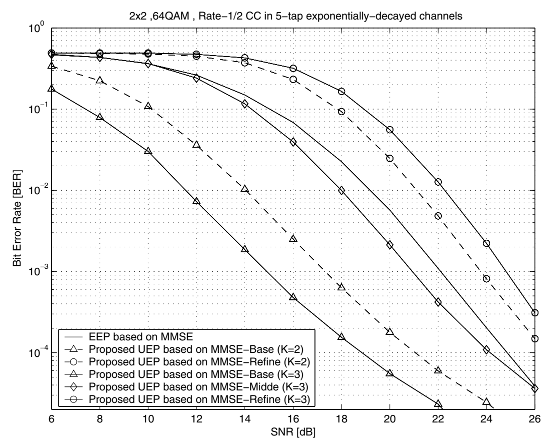

Fig. 6. Performance of the proposed UEP for 64QAM

obtain the LLR values for refinement information. It should be pointed out that the computational complexity of the two stage MLD is mainly determined by the number of refinement bits. Therefore, using the proposed two stage MLD scheme, we can reduce the number of candidate search from $2^{N_t \cdot (m_b + m_r)}$ to $2^{N_t \cdot m_r}$ for the full ML detector.

IV. Performance Evaluation

In this section, we present the simulation results to demonstrate the potential of our proposed UEP and the two stage MLD over Rayleigh fading channels. A rate $\frac{1}{2}$ binary convolutional code with polynomials $(133, 171)$ in octal notation is used throughout the simulations. The OFDM modulation defined in the 802.11a standard with 64 point FFT is assumed. One OFDM symbol duration is set to $4\mu s$ including the $0.8\mu s$ guard interval. A 5 tap multipath channel with an exponentially decaying delay profile is used in the simulations. Also two transmit antennas and two receive antennas are applied. In the simulation, we consider a system where the refinement information BER of 10^{-2} is acceptable.

In Figures 5 and 6, we present the performance comparison with uniform signal constellation between the proposed UEP and the conventional EEP scheme in terms of bit-error rate (BER). With the UEP scheme, a lower BER can be obtained for the sensitive bits, at the expense of a higher BER on the less sensitive bits. In the 16QAM case shown in Fig 5, the proposed UEP scheme assumes $m_b = m_r = 2$. For both cases of MMSE and MLD, the base information performance of the proposed UEP demonstrates more than 3 dB gain over the conventional EEP detector at a BER of 5×10^{-5}. And the refinement information performance of the proposed UEP demonstrates 1 dB loss over the conventional EEP detector at a BER of 10^{-3}. The results for the 64QAM in Fig 6 are similar to those presented in the 16QAM. In this case, the proposed UEP scheme can divide total information into two ($K = 2, m_b = m_r = 3$) or three groups ($K = 3, m_b = m_m = m_r = 2$), where m_m denotes the number of middle information bits. Note that regardless of the number of divided groups K, the complexity of MMSE or MLD remains the same. As can be seen in Fig 6, the base information performance improvement as well as the refinement information performance loss is greater when the information bits are divided by more than two groups. It should be noted that the refinement information performance is only slightly worse than the EEP case.

Figure 7 presents simulation results that show the effect of nonuniform signal constellation for 16QAM case. Here the EEP assumes the uniform signal constellation. By increasing λ, the performance of the refinement information improves at expense of the base information performance. Thus, λ can be optimized depending on the system requirement such as required BER of refinement information. Note that with $d_1 = 2$ and $d_2 = 1.13$ ($\lambda = \frac{1.13}{2}$), the refinement information performance is comparable to EEP. Also for $\lambda = \frac{1.25}{2}$, the overall performance of UEP based on MMSE is about 1dB better than the EEP based on MMSE. Here, the overall performance is computed by averaging BERs for the base information bits and the refinement information bits. Thus, this confirms that nonuniform signal constellation is quite promising for achieving a better performance.

In Figures 8 and 9, we show the performance of the proposed two stage MLD for the 16QAM case with $m_b = m_r = 2$, comparing with the conventional EEP receiver based on the MMSE and the full search MLD. The plot shows that the performance gap between the proposed two

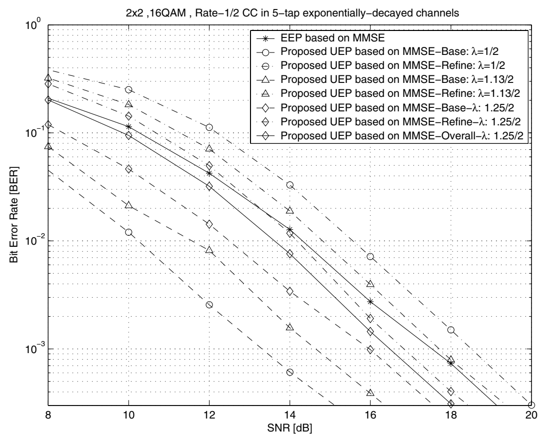

Fig. 7. Performance of the proposed UEP using the nonuniform signal constellation with different λ for 16QAM.

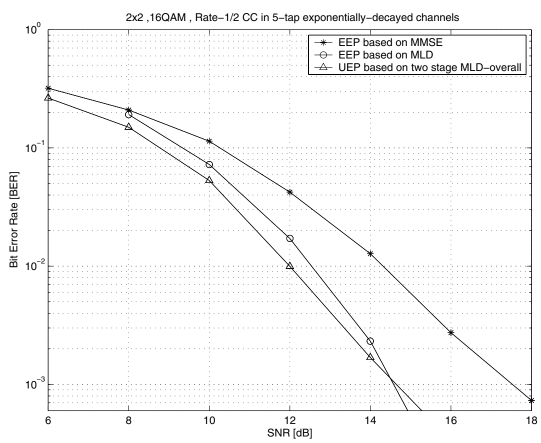

Fig. 8. Performance of the proposed two stage MLD for 16QAM

stage MLD and the full MLD is very small. In Figure 9, similar performance curves are shown for 64QAM. Here we change the number of refinement bits from two to four. In this figure, we observe that as the number of refinement bits increases, the proposed scheme performance gets closer to that of the full search MLD algorithm. Thus we conclude that the proposed two stage MLD method approaches the optimum MLD performance with significantly reduced complexity.

V. Conclusion

In this paper, a new UEP scheme has been proposed and simulated for MIMO-OFDM systems. Numerical results have demonstrated that a significant performance improvement can be obtained with the proposed new detector compared with the conventional EEP detector. Moreover, we can achieve an extra gain by optimizing λ in the nonuniform constellation. Also, we have proposed a two stage MLD to reduce the complexity. A low complexity MLD is designed by decoupling a full MLD into two stages. Although this paper focuses on 16QAM or 64QAM with two transmit antennas, it is straightforward to generalize to arbitrary signal constellations with different antenna configurations.

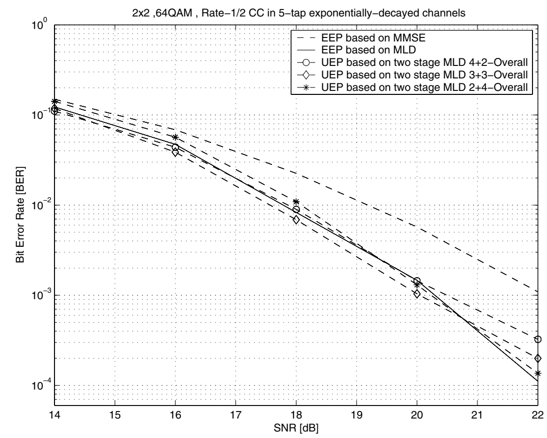

Fig. 9. Performance of the proposed two stage MLD for 64QAM

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References

[1] D. Boudreau, R. Lyon, G. Gallinaro, and R. D. Gaudenzi, "A simulation of audio and video telephony services in a satellite UMTS environment," *in Proc. IMSC*, pp. 220–225, June 1999.

[2] R. Cox, J. Hagenauer, N. Seshadri, and C.-E. Sundberg, "Subband speech coding and matched convolutional channel coding for mobile radio channels," *IEEE Transactions on Acoustics, Speech, and Signal Processing*, pp. 1717 – 1731, Aug 1991.

[3] H. Imai and S. Hirakawa, "A new multilevel coding method using error correcting codes," *IEEE Transactions on Information Theory*, vol. 23, pp. 371–377, May 1977.

[4] R. Morelos-Zaragoza, M. Fossorier, L. Shu, and H. Imai, "Multilevel coded modulation for unequal error protection and multistage decoding: I. Symmetric constellations," *IEEE Transactions on Communications*, vol. 48, pp. 204 – 213, Feb 2000.

[5] P. Vitthaladevuni and M.-S. Alouini, "A closed-form expression for the exact BER of generalized PAM and QAM constellations," *IEEE Transactions on Communications*, pp. 698–700, May 2004.

[6] P. Vitthaladevuni and M.-S. Alouini, "BER computation of 4/M-QAM hierarchical constellations ," *IEEE Transactions on Broadcasting*, vol. 47, pp. 228 – 239, Sept 2001.

[7] K. Ramchandran, A. Ortega, K. Uz, and M. Vetterli, "Multiresolution broadcast for digital HDTV using joint source/channel coding," *IEEE J. Select. Areas Commun.*, vol. 11, pp. 6 – 23, Jan 1993.

[8] F. Tosato and P. Bisaglia, "Simplified soft-output demapper for binary interleaved COFDM with application to HIPERLAN/2," in *Proc. International Conference on Communications '02*, pp. 664–668, September 2002.

[9] X. Li, H. Huang, G. Foschini, and R. A. Valenzuela, "Effects of Iterative Detection and Decoding on the Performance of BLAST," in *Proc. IEEE GLOBECOM '00*, pp. 1061–1066, Dec 2000.