An Optimum Decision Algorithm for MIMO Channel Enhancement

John F An [1] and Justin Wu [2]
[1] Dept. of Communications and Guidance Eng., National Taiwan Ocean University
No.2 Penin Road, Keelung, Taiwan 20224, jfan@mail.ntou.edu.tw
[2] Quanta Computer Inc
Justin_Wu@quantatw.com

1. Introduction

Given a Space-Time Block Code (STBC) Multiple-Input Multiple-Output Spatial-Multiplexing system (MIMO-SM), both diversity gain and spatial multiplexing gain can be obtained simultaneously [1, 2]. An optimum decision algorithm is proposed to achieve an additional weight gain in the Multiple-Input Multiple-Output (MIMO) Space-Time Block Coding (STBC) wireless system design without increased hardware complexity. The optimum weight vector generated using the Bayes decision algorithm [3] maximizes *the most likely 'closest' transmitted signal power to the received vector with a minimum 'Risk' criterion* based on the first- and second-order statistics of the measured MIMO sub-channels, and then multiplies the received sub-channels respectively. Hence, this proposed scheme maximizes the received signal-to-noise ratio (*SNR*) over the spatially-correlated multipath fading channel. The simulation analyses show that our proposed design provides a system performance improvement of about 3 dB in comparison with the conventional design without adopting an optimum decision scheme.

2. System Model

Consider a complex orthogonal STBC design using N spatial diversity antennas and M receiving antennas (STBC-MIMO). In a general form, a sequence of transmit complex symbols $\{s_k\}$, k=0, 1, 2, .., N-1, is first divided into G=N/2 groups for N transmit antennas using pair-wise ML decoder of Alamouti decoding for each antenna group [4, 5]. Fig. 1 shows a typical system model for our performance analyses. The channel between each pair of TX and RX antennas is assumed to be independently and identically distributed (i.i.d.) Rayleigh fading. In reality, however, the spatially-correlated fading MIMO channel owing to inadequate scattering and/or inadequate antenna spacing results in the capacity being substantially reduced.

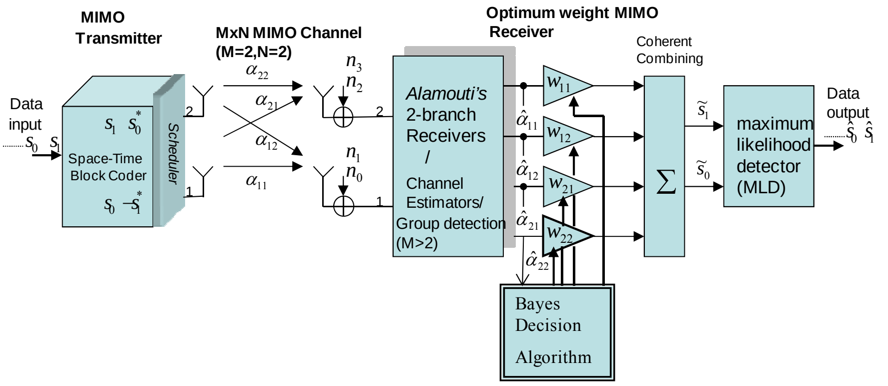

Fig.1 A typical 2x2 MIMO wireless system incorporated with an optimum weight algorithm

In order to fulfill our channel simulations, a so-called Kronecker channel model and the relative power covariance of the channel coefficients [6] is considered here. The power correlation coefficients $|\mu|^2$ and $|\rho|^2$ between the two antenna groups at the transmitter and the receiver are considered respectively, resulting a correlation matrix during our system simulation. The received signal-to-noise power ratio (*SNR*) at the r*th* sub-channel is therefore expressed as $(SNR)_r = \lambda_r \dfrac{P}{\sigma_n^2}$, where

$P = P_T / N$ for equal transmit power. The path gain power, $\sum_{m=1}^{M}\sum_{n=1}^{N}|\alpha_{mn}|^2 = \sum_{r=1}^{R}\lambda_r$, is relevant to the channel coefficients, α_{mn} , which is also expressed as the sum of the eigenvalues of channel covariance matrix with rank *R*.

In our proposed scheme, the key issue is how to generate the optimum weights that result in the received signal enhancement in the better sub-channel and less signal power in the poor sub-channel. The threshold $\tilde{\alpha}_{mn}$ is, therefore, taken from the maximum value of its corresponding optimum decision region (i.e. $\tilde{\alpha}_{mn}$= max[$R_{\alpha_{mn}}$]) for each sub-channel.

3. Optimum Decision Algorithm

The Bayes decision rule for an M-by-N MIMO system uses an extension of the *average cost* criterion [3, 7] over *M*-likelihood receiving antennas. *With a multiple channel scenario, our proposed scheme enhances more path gain in the better channels and lessens the signal power in the poor channels via optimum weigh per sub-channel.* Similar approaches, such as the water-filling algorithm [8], were presented with a closed-loop MIMO antenna structure (i.e. channel covariance known at the transmitter). In Bayes decision algorithm, the *average cost* for a decision is therefore selection of the optimal received signal range such that the *average cost* is minimized on a number of assumptions as follows:

1. A priori probabilities and conditional probability density functions

The statistical properties related to the MN-hypotheses can be categorised into the conditional probability density function, $P(\alpha / R_{ij})$, and its corresponding a priori probability, $P_r(\alpha_{ij})$,The conditional probability density function of the envelope of α_{ij}, thereafter represented by $P(\alpha / R_{ij})$, shows a Rayleigh distribution, and its a priori probability $P_r(\alpha_{ij})$ given to each channel coefficient is assumed to be equal (i.e. $P_r(\alpha_{11}) = P_r(\alpha_{21})$ =….= $P_r(\alpha_{MN})$ =*c*; *c*=1/MN).

2. Cost factors

According to Bayes costs, a zero-one cost assignment (hard decision) is considered here whereby all costs for errors are one and all costs for correct decision are zero,

$$error \quad decision : C_{kl,ij} = 1 \quad for \quad kl, ij = 11,\ 12,\ ..., MN; kl \neq ij$$

$$correct \quad decision : C_{ij,ij} = 0 \ for \quad ij = 11,\ 12,\ ..., MN$$

For the sake of generality, we extend the *average cost* for a MN-hypotheses antenna structure, $\overline{C}$, is given as

$$\overline{C} = \sum_{i=1}^{M}\sum_{j=1}^{N}\sum_{k=1}^{M}\sum_{l=1}^{N} C_{kl,ij} P_r(\alpha_{ij}) \iint ..\int_{R_{kl}} P(\alpha / R_{ij}) ds$$
$$= \iint ..\int_{R_{11}} \sum_{i=1}^{M}\sum_{j=1}^{N} C_{11,ij} P_r(\alpha_{ij}) P(\alpha / R_{ij}) + \iint ..\int_{R_{12}} \sum_{i=1}^{M}\sum_{j=1}^{N} C_{12,ij} P_r(\alpha_{ij}) P(\alpha / R_{ij}) + . \quad . \quad . \tag{1}$$
$$+ \quad \iint ..\int_{R_{21}} \sum_{i=1}^{M}\sum_{j=1}^{N} C_{21,ij} P_r(\alpha_{ij}) P(\alpha / R_{ij}) + \iint ..\int_{R_{22}} \sum_{i=1}^{M}\sum_{j=1}^{N} C_{22,ij} P_r(\alpha_{ij}) P(\alpha / R_{ij}) + . \quad .$$
$$+ \iint ..\int_{R_{MN}} \sum_{i=1}^{M}\sum_{j=1}^{N} C_{MN,ij} P_r(\alpha_{ij}) P(\alpha / R_{ij})$$

If we invoke the definition of the *average cost* function introduced in (1), the integrands can be rewritten as

$$y_{kl}(\alpha) = \sum_{i=1}^{M}\sum_{j=1}^{N} C_{kl,ij} P(\alpha_{ij}) P(\alpha / R_{ij}) \tag{2}$$

The observed channel coefficient α_{ij} can be classified into one of the signal ranges (i.e. $\alpha \in R_{ij}$). From the

cost function (1) and (2), we see that the *average cost* will be minimized if the signal regions R_{ij} , are selected when $\alpha \in R_{ij}$ if $y_{ij}(\alpha) < y_{kl}(\alpha)$ for k =1,2, .. , M and l =1,2,.., N and $ij \neq kl$. R_{ij} is the estimated optimum signal region with respect to the α_{ij} sub-channel and were identified as the intersection of MN-1 individual regions. The optimum decision regions corresponds to the sub-channel coefficients, α_{ij} 's, are selected to be mutually exclusive and exhaustive, with a measured boundary range over three times the σ_0 (i.e. $R_{11} \cap R_{12} \cap R_{13} \cap R_{14} \cap ... \cap R_{M-1,N-1} \cap R_{MN} = 3\sigma_0$), corresponding to the probability of exceeding
the Rayleigh envelope by one percentage (1%). The threshold $\tilde{\alpha}_{ij}$ is taken from the maximum value of its corresponding signal region (i.e. $\tilde{\alpha}_{ij}$ = max(R_{ij}), and then the weight factor is defined as [4]

$$w_{ij} = \frac{\tilde{\alpha}_{ij}}{\sigma_0} \quad , i=1, 2, .., M \text{ and } j=1, 2, .., N \tag{3}$$

4. Simulation Analyses and Summary

From Fig.1, these specific optimum weights are used to multiply the received signals at the received sides of every *L*-block length. The output of maximum coherent combination is therefore given by

$$\begin{aligned} \tilde{s}_0 &\cong \sum_{m=1}^{2}\sum_{n=1}^{2} w_{n,m}\left|\alpha_{n,m}\right|^2 \sqrt{1-\left|\delta_{n,m}\right|^2}\, s_0 + \sum_{n=1}^{4}\alpha_{1,n}^{*} w_{1,n}^{*}(\bar{n}_1) + \sum_{n=1}^{4}\alpha_{2,n} w_{2,n}(\bar{n}_2^{*}) \\ \tilde{s}_1 &\cong \sum_{m=1}^{2}\sum_{n=1}^{2} w_{n,m}\left|\alpha_{n,m}\right|^2 \sqrt{1-\left|\delta_{n,m}\right|^2}\, s_1 - \sum_{n=1}^{4}\alpha_{1,n} w_{1,4}(\bar{n}_2^{*}) + \sum_{n=1}^{4}\alpha_{2,n}^{*} w_{2,n}^{*}(\bar{n}_1) \end{aligned} \tag{4}$$

where $\delta_{n,m}$ is the correlation element of the Kronecker (product) matrix. The spatially-correlated fading, however, causes the received signal coupling loss to be taken into account, as a factor of $\sqrt{1-\left|\delta_{n,m}\right|^2}$. During our simulations, a maximum Doppler frequency, f_d , exists in the i.i.d. complex fading components with normalized $f_d T = 0.01$ being adopted, where *T* is the sampling data duration. Fig. 2 presents the eigen-analysis under spatially correlated 4×4 MIMO channel condition cases A ($\mu_{12} = \mu_{21} = 0.2, \mu_{34} = \mu_{43} = 0.3, and\ \rho_{12} = \rho_{21} = 0.4, \rho_{34} = \rho_{43} = 0.2$). The analysis uses various signal blocks *L* for a total of 204800 samples per channel coefficient. The cumulative distribution function (cdf) of measured eigenvalues ($\lambda_4 > \lambda_3 > \lambda_2 > \lambda_1$) validates our proposed model in which the system incorporated with the Bayes decision algorithm has remarkable *SNR* improvement than the conventional one (without adopting Bayes optimum algorithm).

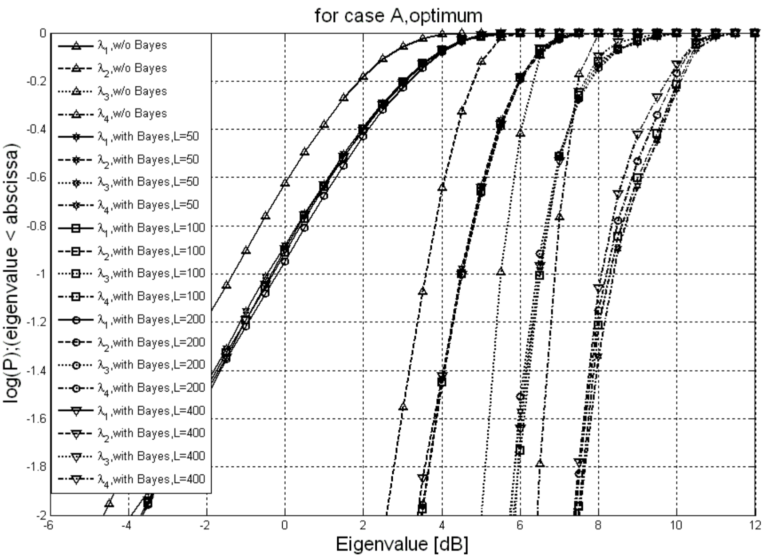

Fig. 2 The cdfs of the eigenvalues in case-C with Bayes decision algorithm for L=50, 100, 200, and 400, and comparison with conventional one (w/o Bayes); X-axis:10 $\log_{10}(\lambda)$ and Y-axis: $\log_{10}$(percentage)

For BER analysis, the QPSK modulations with Gray mapping and the block length of the information signal vector are given, corresponding to various *L* values. The BER performance with an additive white Gaussian noise (AWGN) channel is shown to serve as the performance benchmark using 1Tx/1RX antenna structure (SISO), without considering the spatially-correlated fading effects. In Fig.3, at a BER level of 10^{-3}, there is a performance improvement of about 2.2~3.0 dB compared with the conventional Alamouti two-branch model (without Bayes algorithm) under a spatially-correlated fading channel. However, the coupling loss of correlation coefficients between antennas and matches rational *SNR* values with respect to a $\sqrt{1-\left|\delta_{n,m}\right|^2}$ degradation factor is of interest for further investigation. It is also noted that a small length L suffers less Doppler effects, resulting in more precise channel covariance, but increases the iterative algorithm computational load.

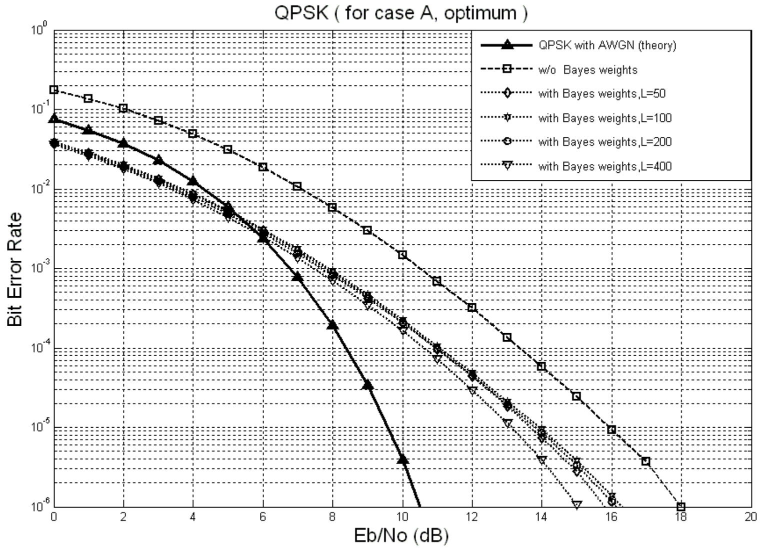

Fig.3 Average BER performance versus E_b / N_0 – comparison of QPSK simulation results with/without Bayes decision algorithm against various L in spatially-correlated case-A, vs. unconditional BER in AWGN (1-Tx/1-Rx)

References

[1] Hamid Jafarkhani, “A Quasi-Orthogonal Space-Time Block Code,” IEEE Transaction on Communications, Vol.49, No.1, pp.1-4, January 2001

[2] G. J. Foschini and M. J. Gans, “On Limits of Wireless Communications in a Fading Environment when Using Multiple Antennas”, Wireless Personal Communications, Vol.6, No.3, pp.311-335, March 1998

[3] John F. An and An-Kuo Liu, “An Optimum Sub-channel Weighting Scheme for MIMO-OFDM Wireless System”, Proceedings of 18th *IEEE* International Symposium on Personal, Indoor and Mobile Communications, PIMRC2007, Athens, Greece, September 2007

[4] S.M. Alamouti, “A simple Transmit Diversity Technique for Wireless Communications,” *IEEE JSAC*, vol.16, pp 1451-1458, Oct. 1998

[5] Lan Zhao and V.K. Dubey, “Detection Schemes for Space-Time Block Code and Spatial Multiplexing Combined System,” IEEE Communications Letters, vol.9,No.1, January 2005, pp 49-51

[6] J. Kermoal, L. Schumacher, K. Pedersen, P. Mogensen, and F. Frederiksen, “A stochastic MIMO radio channel model with experimental validation,” IEEE JSAC, vol.20, pp.1211-1226, August 2002

[7] Lonnie C. Ludeman, “Random Processes - Filtering, Estimation, and Detection”, Wiley, pp.437-452, 2003

[8] C. Chuah, D. Tse, J. Kahn and R. Valenzuela, “Capacity scaling in MIMO Wireless systems under correlated fading,” IEEE on Information Theory , Vol.48, No. 3, pp. 637-650, March 2002