

MIMO Beamforming for High-Bit-Rate Transmission over Frequency-Selective Fading Channels

Huy Hoang Pham, Tetsuki Taniguchi and Yoshio Karasawa Department of Electronic Engineering,
The University of Electro-Communications
Chofugaoka 1-5-1, Chofu-shi, Tokyo, 182-8585 Japan
Email: hoang@radio3.ee.uec.ac.jp

Abstract—In recent years, smart antenna systems have been proposed and demonstrate at both the transmitter and receiver of wireless communication system, the quality of service improves at both ends. In this paper, we propose multiple-input multiple-output (MIMO) beamforming for frequency-selective fading channels to maximize the Signal-to-Noise and Interference Ratio (SINR) based on an iterative update algorithm of transmit and receive weight vectors with prior knowledge of the channel state information (CSI) at both the transmitter and receiver. We derive the necessary conditions for an optimum weight vector solution and propose an iterative weight update algorithm for an optimal SINR reception. The Maximum Signal-to-Noise (MSN) method, where noise includes the additive gaussian noise and interference signals, is used as a criterion. The proposed MIMO with $M \times N$ arrays allows the cancellation of $M + N - 2$ delayed channels. Computer simulations are presented to verify our analysis. The results show that significant improvement in performance is possible in wireless communication systems.

I. INTRODUCTION

Nowadays, communication systems have been investigated to support a wide range of service available through broadband multimedia wireless communications. These attempts have raised to improve high transmission rates of at least several megabits per second. In recent years, multiple-input multiple-output (MIMO) antenna systems, where antenna arrays are equipped at both the transmitter and receiver, have been suggested for high-bit-rate transmission. The MIMO systems must be able to cope with frequency-selective fading channels. Moreover, if single-carrier serial high-speed data is transmitted at a rate of several megabits per second, the delay spread of multipath waves exceeds 1 symbol period. Currently, two different space-time architectures have been suggested for the MIMO dispersive channel. The first architecture uses a multi-carrier MIMO system (e.g., MIMO-OFDM [orthogonal frequency division multiplexing]) to reduce the effects of multipath fading based on a parallel data transmission scheme in frequency region, in which the delay spread of multipath waves is suppressed to within 1 symbol period [1]-[3]. The second architecture is the transmit and receive beamforming, which has been studied extensively [4]-[8]. The MIMO system employs equalization techniques (e.g., MIMO-DFE [decision-feedback equalizer]) to find the optimal solution over frequency-selective fading channels without prior knowledge of channel state information (CSI). However, there are

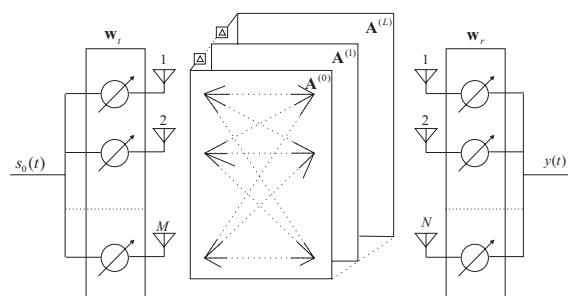


Fig. 1. Broadband MIMO channel and beamforming configuration.

practical difficulties associated with operating the equalization at several megabits per second with high-speed, compact, and low-cost hardware. Although there are many proposed schemes for MIMO frequency-selective fading channels with or without prior knowledge of CSI at the transmitter and/or receiver, these schemes do not yet point out how many delayed channels can be effectively cancelled by using adaptive beamforming by adjusting both transmitter and receiver weights in the MIMO system.

In this paper, we propose a weights determination scheme for MIMO frequency-selective fading channels with prior knowledge of CSI at both the transmitter and receiver. The scheme is then used in an iterative update algorithm of transmit and receive weight vectors for maximizing SINR. The cancellation of $M + N - 2$ delayed channels is identified by $M \times N$ MIMO beamforming. We will carry out a practical evaluation of our proposed MIMO beamforming using a conventional propagation model. It shows that the increase in the transmit and receive antenna elements improves the SINR when the delay spread of multipath waves is greater than 1 symbol period.

II. PROPAGATION MODEL OF MIMO CHANNEL

A general MIMO beamforming for single-data-stream transmission under frequency-selective fading channels is shown in Fig. 1. The number of transmit and receive antennas is M and N , respectively.

For wireless broadband systems, the MIMO propagation

channel can be modeled as

$$\mathbf{H}(\tau) = \sum_{i=0}^L \mathbf{A}^{(i)} \delta(\tau - i\Delta\tau) \quad (1)$$

$$\mathbf{A}^{(i)} = \begin{pmatrix} a_{11}^{(i)} & a_{12}^{(i)} & \cdots & a_{1M}^{(i)} \\ a_{21}^{(i)} & a_{22}^{(i)} & \cdots & a_{2M}^{(i)} \\ \cdots & \cdots & \cdots & \cdots \\ a_{N1}^{(i)} & a_{N2}^{(i)} & \cdots & a_{NM}^{(i)} \end{pmatrix} \quad (2)$$

where $i=0$, $\mathbf{A}^{(0)}$ is the channel information of the preceding wave which we regard as the desired wave in this paper, while for $i=1, \dots, L$, $\mathbf{A}^{(i)}$ is the i^{th} delayed channel information which we consider as interference waves. The notation $a_{nm}^{(i)}$ means the i^{th} delayed channel response between the m^{th} transmit antenna and n^{th} receive antenna. $\Delta\tau$ is the unit delay time, which corresponds to symbol period T_s , of the modulated signal.

III. DETERMINATION OF TRANSMIT AND RECEIVE WEIGHTS BASED ON MAXIMUM SINR CRITERION

Standard notations are used in this paper. Bold letters denote vectors and matrices. Other notation is as follows.

- $(\cdot)^*$ Conjugate
- $(\cdot)^T$ Transpose
- $(\cdot)^H$ Conjugate transpose
- $\mathbf{I}$ Identity matrix
- $\langle \cdot \rangle$ Ensemble average
- $\|(\cdot)\|$ Euclid-norm of vector $(\cdot)$

The output of array elements are linearly combined with weights to give the received signal. The received signal at the receiver is expressed as

$$y(t) = \sum_{i=0}^L \mathbf{w}_r^H \mathbf{A}^{(i)} \mathbf{w}_t s_0(t - i\Delta\tau) + \mathbf{w}_r^H \mathbf{n}(t) \quad (3)$$

where $s_0(t)$ is the source signal, $\mathbf{n} = [n_1, n_2, \dots, n_N]^T$ is the additive white gaussian noise (AWGN) vector, $\mathbf{w}_t = [w_{t1}, w_{t2}, \dots, w_{tM}]^T$ is a beamforming transmit weight vector, and $\mathbf{w}_r = [w_{r1}, w_{r2}, \dots, w_{rN}]^T$ is a beamforming receive weight vector.

Assume that each delayed signal is uncorrelated and zero-mean, thereby

$$\langle s_0^*(t - i\Delta\tau) s_0(t - j\Delta\tau) \rangle = 0 \quad \text{for } i \neq j \quad (4)$$

Let us define P_S , P_N and $1/\gamma$, which are the signal power, noise power and power ratio of the signal to noise, the SINR at the receiver is given by

$$\eta(\mathbf{w}_t, \mathbf{w}_r) = \frac{\mathbf{w}_r^H \mathbf{A}^{(0)} \mathbf{w}_t \mathbf{w}_t^H (\mathbf{A}^{(0)})^H \mathbf{w}_r}{\sum_{i=1}^L \mathbf{w}_r^H \mathbf{A}^{(i)} \mathbf{w}_t \mathbf{w}_t^H (\mathbf{A}^{(i)})^H \mathbf{w}_r + \gamma \mathbf{w}_r^H \mathbf{w}_r} \quad (5)$$

Equation (5) is a multivariable nonlinear equation. Since both the transmit and receive weight vectors are contained in the numerator and denominator, it seems difficult to find the optimal weight vectors analytically. Therefore, we propose a solution to find the optimal transmit and receive weight vectors

iteratively as follows.

Receive weight vector estimation: Assume that the transmit weight vector is given with a mandatory condition of $\mathbf{w}_t^H \mathbf{w}_t = 1$ and an optional condition of $\mathbf{w}_r^H \mathbf{w}_r = 1$. The steering vector of the desired signal is $\mathbf{A}^{(0)} \mathbf{w}_t$; thereby, the receive weight vector which is resolved by the Maximum-Signal-to-Noise (MSN) method [9], are related as follows.

$$\mathbf{A}^{(0)} \mathbf{w}_t = \frac{\mathbf{w}_r^H \mathbf{A}^{(0)} \mathbf{w}_t}{\mathbf{w}_r^H \mathbf{R}_{nr} \mathbf{w}_r} \mathbf{R}_{nr} \mathbf{w}_r \quad (6)$$

where

$$\mathbf{R}_{nr} \equiv \sum_{i=1}^L \mathbf{A}^{(i)} \mathbf{w}_t \mathbf{w}_t^H (\mathbf{A}^{(i)})^H + \gamma \mathbf{I} \quad (7)$$

Let us define $\frac{1}{\zeta} \equiv \frac{\mathbf{w}_r^H \mathbf{A}^{(0)} \mathbf{w}_t}{\mathbf{w}_r^H \mathbf{R}_{nr} \mathbf{w}_r}$, then the received weight vector is given by

$$\mathbf{w}_r = \zeta \mathbf{R}_{nr}^{-1} \mathbf{A}^{(0)} \mathbf{w}_t \quad (8)$$

In equation (7), $\mathbf{R}_{nr}$, is a Hermitian matrix, and the coefficient ζ is a complex scalar. In equation (8), since the coefficient ζ does not effect the received SINR itself, the optimal receive weight vector can be normalized and given by

$$\mathbf{w}_r^{(opt)} = \mathbf{R}_{nr}^{-1} \mathbf{A}^{(0)} \mathbf{w}_t / \|\mathbf{R}_{nr}^{-1} \mathbf{A}^{(0)} \mathbf{w}_t\| \quad (9)$$

Accordingly, the optimal receive weight vector can be determined if the transmit weight vector is given.

Transmit weight vector estimation: Assume that the receive weight vector is given, we determine the optimal transmit weight vector satisfying the condition $\mathbf{w}_t^H \mathbf{w}_t = 1$. To find the optimal transmit weight vector which maximizes SINR, we use the Lagrange multiplier method as follows.

$$\phi = \eta(\mathbf{w}_t, \mathbf{w}_r) + \lambda(1 - \mathbf{w}_t^H \mathbf{w}_t) \quad (10)$$

From equation (10), we rewrite the $\eta(\mathbf{w}_t, \mathbf{w}_r)$ expression as

$$\eta(\mathbf{w}_t, \mathbf{w}_r) = \frac{\mathbf{w}_t^H (\mathbf{A}^{(0)})^H \mathbf{w}_r \mathbf{w}_r^H \mathbf{A}^{(0)} \mathbf{w}_t}{\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \mathbf{w}_r^H \mathbf{w}_r} \quad (11)$$

where

$$\mathbf{R}_{it} \equiv \sum_{i=1}^L (\mathbf{A}^{(i)})^H \mathbf{w}_r \mathbf{w}_r^H \mathbf{A}^{(i)} \quad (12)$$

Taking the derivative of equation (11) with respect to $\mathbf{w}_t$ and setting it to zero.

$$\frac{\partial \phi}{\partial \mathbf{w}_t} = \frac{\partial \eta}{\partial \mathbf{w}_t} + \frac{\partial \lambda(1 - \mathbf{w}_t^H \mathbf{w}_t)}{\partial \mathbf{w}_t} = 0 \quad (13)$$

then,

$$\frac{\mathbf{R}_0 \mathbf{w}_t (\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)}{(\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)^2} - \frac{\mathbf{R}_{it} \mathbf{w}_t (\mathbf{w}_t^H \mathbf{R}_0 \mathbf{w}_t)}{(\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)^2} - \lambda \mathbf{w}_t = 0 \quad (14)$$

where

$$\mathbf{R}_0 \equiv (\mathbf{A}^{(o)})^H \mathbf{w}_r \mathbf{w}_r^H \mathbf{A}^{(o)} \quad (15)$$

In order to calculate the value of λ , equation (14) is multiplied by weight vector $\mathbf{w}_t^H$ as follows.

$$\begin{aligned} & \mathbf{w}_t^H \frac{\mathbf{R}_0 \mathbf{w}_t (\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)}{(\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)^2} \\ & - \mathbf{w}_t^H \frac{\mathbf{R}_{it} \mathbf{w}_t (\mathbf{w}_t^H \mathbf{R}_0 \mathbf{w}_t)}{(\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)^2} + \mathbf{w}_t^H \lambda \mathbf{w}_t = 0 \end{aligned} \quad (16)$$

Applying the condition of $\mathbf{w}_t^H \mathbf{w}_t = 1$, parameter λ is given by

$$\lambda = - \frac{\mathbf{w}_t^H (\mathbf{A}^{(o)})^H \mathbf{w}_r \mathbf{w}_r^H \mathbf{A}^{(o)} \mathbf{w}_t \gamma \|\mathbf{w}_r\|^2}{(\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2)^2} \quad (17)$$

Substituting (17) into (14) and let us define $\frac{1}{\xi} \equiv \frac{\mathbf{w}_t^H (\mathbf{A}^{(o)})^H \mathbf{w}_r}{\mathbf{w}_t^H \mathbf{R}_{it} \mathbf{w}_t + \gamma \|\mathbf{w}_r\|^2}$, then we have

$$(\mathbf{A}^{(o)})^H \mathbf{w}_r = \frac{1}{\xi} \mathbf{R}_{nt} \mathbf{w}_t \quad (18)$$

where

$$\mathbf{R}_{nt} = \mathbf{R}_{it} + \gamma \|\mathbf{w}_r\|^2 \quad (19)$$

Without detailed calculation of ξ , the optimal transmit weight vector can be given by

$$\mathbf{w}_t^{(opt)} = \mathbf{R}_{nt}^{-1} (\mathbf{A}^{(o)})^H \mathbf{w}_r / \|\mathbf{R}_{nt}^{-1} (\mathbf{A}^{(o)})^H \mathbf{w}_r\| \quad (20)$$

We conclude the determination of transmit and receive weight vectors by pointing out that the receive weight vector is estimated by equation (9) with the given transmit weight vector. Similarly, the transmit weight vector is estimated by equation (20) with the given receive weight vector. However, because the simultaneous equations become nonlinear, they seem difficult to solve analytically. Therefore, an iterative algorithm, which estimates the receive and transmit weight vectors alternately, is a simple way to reach a solution for maximizing SINR. The iterative operation ($k > 1$) is described as follows.

For the updated receive weight vector

$$\bar{\mathbf{w}}_r(k) = \mathbf{R}_{nr}^{-1}(k-1) \mathbf{A}^{(o)} \mathbf{w}_t(k-1) \quad (21)$$

where

$$\mathbf{R}_{nr}(k) \equiv \sum_{i=1}^L \mathbf{A}^{(i)} \mathbf{w}_t(k) \mathbf{w}_t^H(k) (\mathbf{A}^{(i)})^H + \gamma \mathbf{I} \quad (22)$$

and

$$\mathbf{w}_r(k) = \bar{\mathbf{w}}_r(k) / \|\bar{\mathbf{w}}_r(k)\| \quad (23)$$

$\mathbf{R}_{nr}(k-1)$ is updated based on the previous transmitted weight vector $\mathbf{w}_t(k)$ for finding the optimal receive weight vector $\mathbf{w}_r(k)$.

For the updated transmit weight vector

$$\bar{\mathbf{w}}_t(k) = \mathbf{R}_{nt}^{-1}(k) (\mathbf{A}^{(o)})^H \mathbf{w}_r(k) \quad (24)$$

where

$$\mathbf{R}_{nt}(k) \equiv \sum_{i=1}^L (\mathbf{A}^{(i)})^H \mathbf{w}_r(k) \mathbf{w}_r^H(k) \mathbf{A}^{(i)} + \gamma \mathbf{I} \quad (25)$$

and

$$\mathbf{w}_t(k) = \bar{\mathbf{w}}_t(k) / \|\bar{\mathbf{w}}_t(k)\| \quad (26)$$

Similarly, $\mathbf{R}_{nt}(k)$ is also updated based on the previous received weight vector $\mathbf{w}_r(k)$ for finding the optimal transmit weight vector $\mathbf{w}_t(k)$.

Though the problem of selection of the initial weight vector is now under examination, in this paper, the transmit weight vector $\mathbf{w}_t(0)$ for $k = 0$ is determined from MIMO system in flat fading channel environment as follows.

$$\mathbf{w}_t^{(opt)} \{= \mathbf{w}_t(0)\} = \mathbf{e}_{t,max} \quad (27)$$

The vector $\mathbf{e}_{t,max}$ is the eigenvector corresponding to the largest eigenvalue λ_{max} for the correlation matrix $(\mathbf{A}^{(o)})^H \mathbf{A}^{(o)}$ [11].

IV. SIMULATION

A. Delayed Channels Cancellation by means of MIMO beamforming

It is known that the maximum diversity gain of a MIMO system has been pointed out to be $M \times N$ when M transmitter and N receiver antennas are used [10], but how many delayed channels can be cancelled by MIMO system in a single data stream transmission is a question that could not be answered clearly until now.

Let us now consider array technology, where the optimal beamformer without tapped delay lines is used to cancel the interference or delayed channels. A Single-Input Multiple-Output (SIMO) system with a single transmitter antenna and N receiver antennas has $N - 1$ degree of freedom. In other words, the SIMO system has the capability to cancel $N - 1$ delayed channels. Similarly to a Multiple-Input Single-Output (MISO) with M transmitter antennas and a single receiver antenna, there are $M - 1$ delayed channels which can be suppressed effectively. As a result, the degree of freedom or the maximum cancellation number of delayed channels for the MIMO system becomes a certain value, which is considered to be larger than that of both SIMO and MISO systems. Through straightforward thinking, we expect that the maximum number of L may be given as

$$L = (M - 1) + (N - 1) = M + N - 2 \quad (28)$$

The maximum number of delayed channels cancellation will clearly be confirmed by computer simulation.

B. Simulation Conditions

In this section, we offer the simulation conditions to demonstrate our analysis. We assumed that the components of the channel state matrix are independent identically distributed (i.i.d) with complex Gaussian distribution, with the uniform

power delay profile (model 1) and the exponential power delay profile (model 2), respectively, given by

$$p(\tau) = \sum_{i=0}^L \delta(\tau - iT_s) \quad (\text{model 1}) \quad (29)$$

$$p(\tau) = \frac{P_R}{\sigma_\tau} \exp(-\frac{\tau}{\sigma_\tau}) \quad (\text{model 2}) \quad (30)$$

where T_s is the symbol period of the transmitted signal, and P_R is the average power of multipath waves.

For model 1, each impulse $a_{nm}^{(i)}$ is generated by the i.i.d. process while keeping $\langle |a_{nm}^{(i)}|^2 \rangle = 1$ for $n = 1 \sim N, m = 1 \sim M$ and $i = 0 \sim L$. In order to clarify the limit of the delayed channels cancellation ability, we set a sufficiently high SNR ($= P_S/P_N$), such that $P_S/P_N = 40\text{dB}$. For model 2, the exponential power profile will be used throughout this paper in order to have a more realistic channel model analyzing the performance of our proposed method. To identify the improvement of the output SINR with respect to the delay spread of multipath waves, we set the input SNR = 20dB. A BPSK (Binary Phase Shift Keying) modulation scheme has been used and the noise has been considered to be additive white Gaussian.

In order to calculate the output SINR reception, we utilize the cross-correlation coefficient defined as

$$\rho = \frac{E[y(t)s_0^*(t)]}{\sqrt{E[|y(t)|^2]E[|s_0(t)|^2]}} \quad (31)$$

The output SINR reception calculated via the cross-correlation coefficient is given by

$$\text{SINR}_{\text{out}} = \frac{|\rho|^2}{1 - |\rho|^2} \quad (32)$$

C. Results

We performed a simulation for the 4×4 MIMO system over multipath frequency-selective propagation channel of model 1 based on the iterative update algorithm of transmit and receive weight vectors, which described in the previous section. To show clearly the sufficient increase in the output SINR by using our proposed method for frequency-selective fading channels, we employ the weight vectors estimated from the SVD method for MIMO flat fading channel as the initial weight vectors. Figure 2 illustrates the averaged convergence characteristics of the system for different delays (L) with 100 times under a random channel condition. It is seen that the maximum number of delayed channels cancellation of the 4×4 MIMO system is 6, which agrees very well with the result of $M + N - 2$, given in equation (28). It should also be noted that $k = 80$ is a point having sufficient convergence for a system undergoing 6 delayed channels.

Figure 3 shows the cumulative distribution function (CDF) of SINR for the 4×4 MIMO system after 100 random trials with the signal-to-noise power ratio set to 40dB for model 1. The CDF yields the SINR probability that the proposed MIMO cancels the number of delayed channels. As can be seen in Fig. 3, the SINR curves decrease with increasing the number of delayed channels. However, the proposed MIMO

effectively cancels the 6 delayed channels seen at 50 percent of the CDF.

In order to confirm the maximum cancellation of delayed channels of the $M \times N$ MIMO system, we performed a simulation for the cases of 2×2 , 3×3 and 4×4 MIMO systems in terms of SINRs at 50 percent of the CDF. The median value of SINR versus the number of delayed channels is shown in Fig. 4. It is clear that the proposed beamforming method can effectively cancel $M + N - 2$ delayed channels, namely, 2, 4 and 6 for the 2×2 , 3×3 and 4×4 MIMO systems.

By using model 2, we are now going to show how our proposed method can apply to the frequency-selective fading channels with the delay spread of multipath waves over one symbol period. We performed a simulation for the cases of 1×1 (SISO), 2×2 , 3×3 , 4×4 and 8×8 (MIMO) systems with the delay spread σ_τ ranging from 1 to $10T_s$. The average SINR results are shown in Fig. 5. It was shown that based on our proposed method, the increase in antenna elements at both ends improves considerably the output SINR when the delay spread of multipath waves is greater than one symbol period.

Successively, we performed a simulation for the cases of 2×2 , 3×3 , 4×4 and 8×8 MIMO systems in terms of SINR at 14dB, which is considered to be an acceptable condition in wireless communication systems. The transmit and receive antennas of the $M \times N$ MIMO beamforming with $M = N$ versus the delay spread σ_τ is shown in Fig. 6. It was shown that $\sigma_\tau = (N - 1)T_s$ for $M = N$ is a maximum delay spread where a reliable communication can be achieved by the proposed beamforming method.

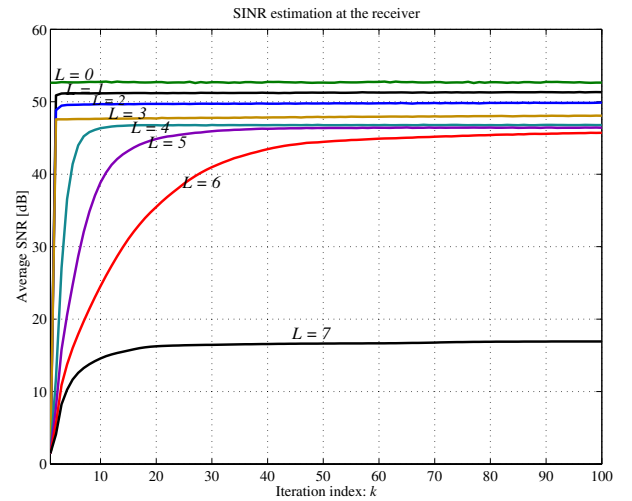


Fig. 2. Convergence characteristics for different delays. (model 1)

V. CONCLUSION

In this paper, we proposed a MIMO beamforming for high-bit-rate transmission over frequency-selective fading channels. This scheme yields a SINR improvement based on the iterative update algorithm of the transmit and receive weight vectors to enables the cancellation of $M + N - 2$ delayed channels. It also enables the adaptation of the delay spread of multipath waves

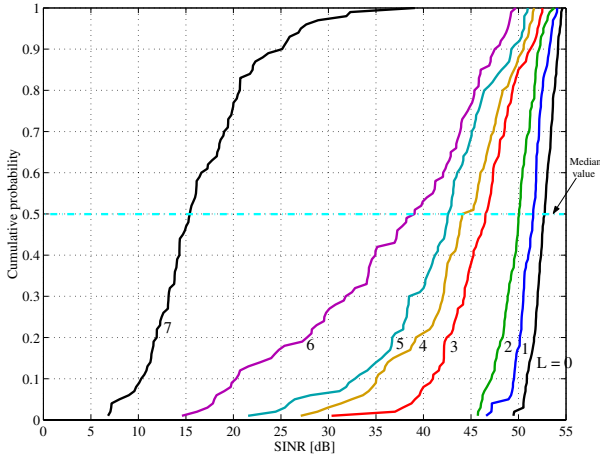


Fig. 3. CDF of SINR for the 4x4 MIMO at the receiver. (model 1)

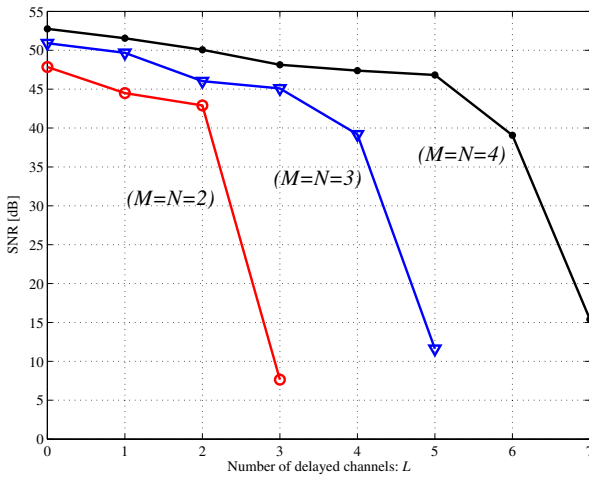


Fig. 4. Median value of SINR vs. the number of delayed channels. (model 1)

with an exponential power delay profile. The simulation result shows that the convergence speed of our proposed method is fast with knowledge of the CSI at both the transmitter and receiver.

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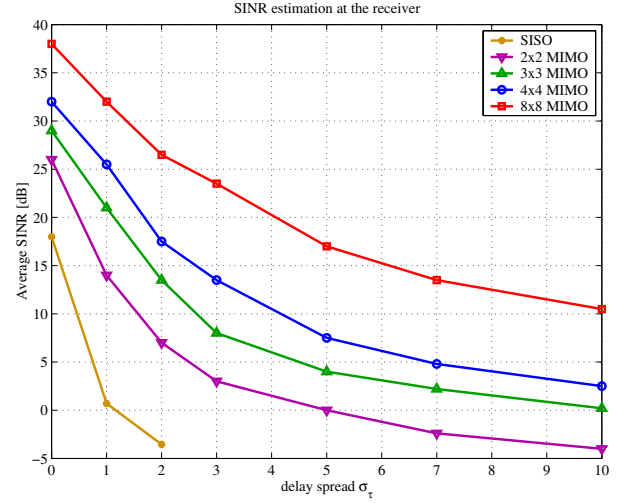


Fig. 5. Average SINR of MxN MIMO as a function of the delay spread σ_τ . (model 2)

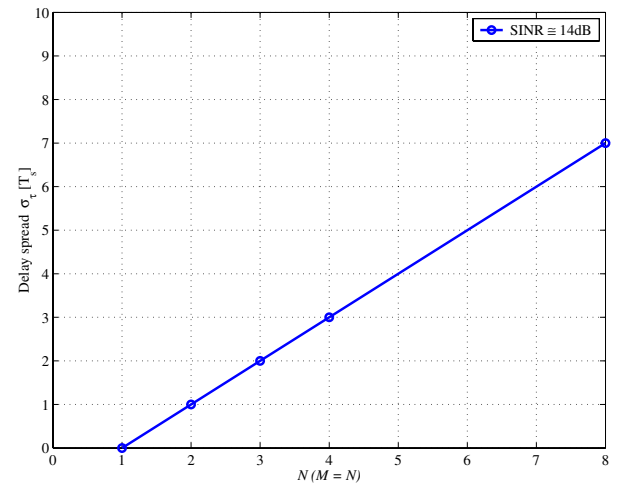


Fig. 6. The maximum delay spread realizing acceptable communication quality as a function of the number of array elements. (model 2)

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