PAPER *Special Section on Wideband Systems*

Broadband MIMO Communication Systems Using Spatio-Temporal Processing in Transmitter and Receiver Sides

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SUMMARY This paper presents a mathematically simple method of maximum SINR (Signal to Interference plus Noise Ratio) design of broadband MIMO (Multiple Input Multiple Output) communication systems adopting TDL (Tapped Delay Line) structure for spatio-temporal processing in both transmitter and receiver sides. The weight vectors in both ends are determined alternately, optimizing one side by fixing the other, and this operation is repeated until the SINR converges. The performance of MIMO systems using the proposed approach is investigated through computer simulations, and it is demonstrated that, though it requires high computational cost, the TDL structure brings high ability to mitigate the influence of frequency selective fading, particularly when the duration of the delay profile is long. Moreover, experimental results show that the equable distribution of the resources (weights and delay units) to both arrays is better choice than the concentration of them to one side of the transmitter or receiver.

key words: *MIMO, maximum ratio combining, frequency selective fading, spatio-temporal processing, tapped delay line*

1. Introduction

MIMO systems adopting array antennas both in transmitter and receiver sides are collecting attentions to cope with the requirement for the increase in the capacity of communications over fading channels, and a variety of schemes are proposed to improve the transmission efficiency [1], [2]. When the transmitter has no information of channel, various space-time coding techniques are proposed as represented by famous STBC schemes given in [3], [4]. For the maximum SINR (Signal to Interference plus Noise Ratio) design of MIMO systems assuming the perfect knowledge of flat fading channel state information (CSI), an effective multistream transmission based on singular value decomposition (SVD) of the channel matrix has been proposed [5].

Under the environment of frequency selective fading (FSF), however, this simple approach could not be applicable because of the existence of delayed waves, hence many schemes have been presented to avoid intersymbol interferences. One solution to this problem is the use of MIMO systems with multicarrier modulation schemes (ex., MIMO-OFDM system [6], [7]) which enables us to utilize simple SVD-based design to each subchannel. Single carrier methods are also presented based on block processing with zero padding or addition of guard interval [8], [9]. For the situation of the uninformed transmitter, a space-time coding with cyclic prefix (CP) [10] or time reversal strategy [11] are also proposed. But those approaches accompanied with the extension of data bring the degradation of the transmission rate. Multiuser method based on CDM (Code Division Multiplexing) in [12] is effective also for FSF. The method in [13] utilizing spatial filtering achieves cancellation of FSF without enlargement of data size, but the number of delayed waves is restricted by the number of array elements. MIMO systems equipped with tapped delay line structure (TDL-MIMO systems) are also considered for the case of continuous signals [14] and discrete signals [15], [16], but they are based on complicated nonlinear optimization theories. In [17], TDL adaptive array is utilized in receiver side, but the transmitter has no ability of temporal equalization. Reference [18] investigates the cascade connection of spatial filter and temporal equalizer, and [19] deals with the case using TDL-based array in the transmitter, but processing in entire spatio-tempral region is impossible using those methods.

In this paper, we propose a mathematically simple design procedure of MIMO communication systems equipped with TDL structure both in transmitter and receiver sides based on maximum SINR criterion. The performance of the proposed MIMO systems is evaluated under FSF through computer simulations. Some properties of the proposed methods and distribution problem of array resources are also discussed.

The rest of this paper is organized as follows: Sect. 2. describes the model of TDL-MIMO communication systems under the condition of FSF, then a simple maximum SINR design method of weights is proposed in Sect. 3. In Sect. 4., the performance of TDL-MIMO systems using the proposed approach is investigated, and finally, conclusions and future works are described in Sect. 5.

2. Structure of Proposed MIMO System and Frequency Selective Channel

In this section, as a foundation of the discussions in the following sections, the input-output relation of the MIMO

Manuscript received March 2, 2007.
Manuscript revised May 31, 2007.
Final manuscript received July 9, 2007.

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DOI: 10.1093/ietfec/e90–a.11.2431

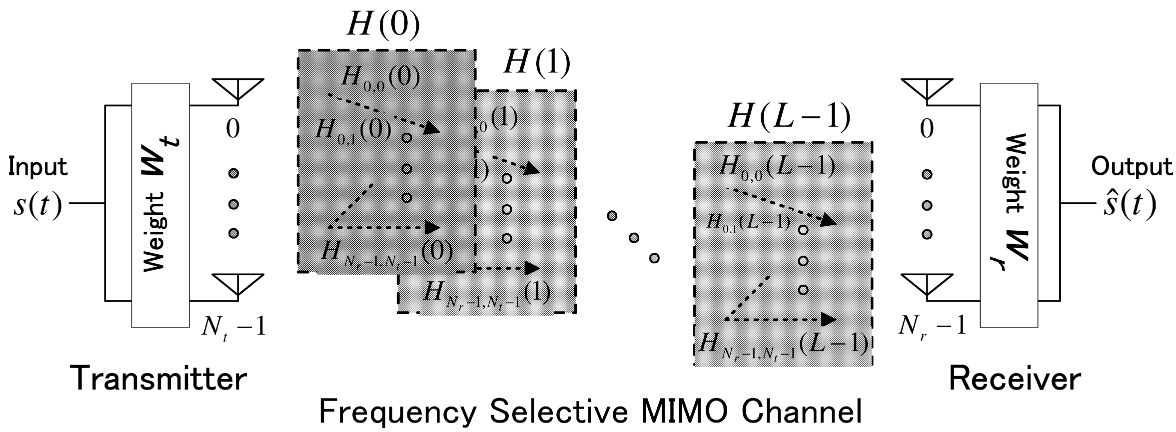
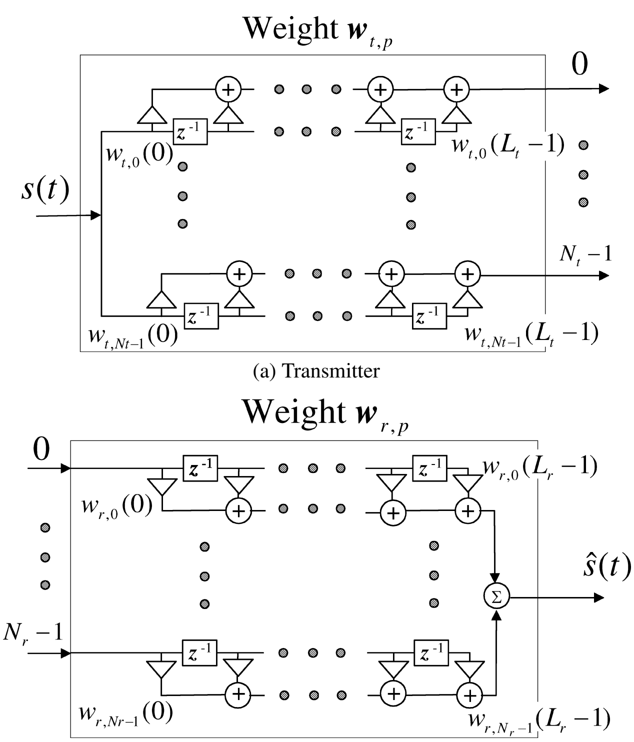


Fig. 1 Model of proposed broadband MIMO communication system under FSF.



(a) Transmitter

(b) Receiver

Fig. 2 Detailed transmitter and receiver structures of TDL-MIMO communication system shown in Fig. 1.

communication systems under FSF is formulated.

The system model considered in this study is presented in Fig. 1. The details of weights in transmitter and receiver sides are given in Fig. 2. The transmitter array consists of N_t elements each of them is preceded by L_t weights and $L_t - 1$ delay units. The receiver array has N_r elements followed by L_r weights and $L_r - 1$ delay units. The propagation channel is expressed by a sequence of N_r-by-N_t matrices $\{H(\ell),\ \ell = 0, \cdots, L-1\}$, where each element $H_{n_t,m_t}(\ell)$ denotes quasi-static fading path from the n_t-th antenna element of the transmitter array to the n_r-th antenna element of the receiver array corresponding delay of ℓT_S (T_S: symbol period) (hence the impulse response of MIMO channel is expressed by $\sum_\ell H(\ell)\delta(t - \ell Ts)$, where $\delta(t)$ denotes delta function). The input sequence $\{s(t); t \in \mathbb{Z}\}$ is multiplied by transmission weights $\{\boldsymbol{w}_t(\ell_t),\ \ell_t = 0, \cdots, L_t - 1\}$, then propagated through channel $\{H(\ell)\}$. After arriving at the receiver array, N_r-by-1 AWGN vector $\boldsymbol{n}(t)$ is added and the output signal $\hat{s}(t)$ is extracted by multiplying receiver weights $\{\boldsymbol{w}_r(\ell_r),\ \ell_r = 0, \cdots, L_r - 1\}$. Consequently, the output signal is formulated as follows:

$$\hat{s}(t) = \sum_{\ell_r=0}^{L_r-1} \boldsymbol{w}_r^H(\ell_r) \left\{ \sum_{\ell=0}^{L-1} \sum_{\ell_t=0}^{L_t-1} H(\ell)\boldsymbol{w}_t(\ell_t) \right. \\ \left. \times\, s(t - \ell_t - \ell + \ell_r) + \boldsymbol{n}(t + \ell_r) \right\}. \quad (1)$$

The detailed definition of weight vectors are as follows:

$$\begin{aligned} \boldsymbol{w}_t(\ell_t) &= [w_{t,0}(\ell_t), \cdots, w_{t,N_t-1}(\ell_t)]^T \\ \boldsymbol{w}_r(\ell_r) &= [w_{r,0}(\ell_r), \cdots, w_{r,N_r-1}(\ell_r)]^T \end{aligned} \quad (2)$$

Here, signal $s(t)$ is assumed to be self-uncorrelated ($E[s(t)s^*(t-\tau)] = P_S\delta(\tau),\ \ \tau \in \mathbb{Z}$, where P_S is symbol power), and $s(t)$ and the elements of N_r-dimensional noise vector $\boldsymbol{n}(t)$ are also uncorrelated. The equation (1) is simplified in the z-region as

$$\hat{s}(z) = \boldsymbol{w}_r^H(z)\,\{H(z)\boldsymbol{w}_t(z)s(z) + \boldsymbol{n}(z)\} \quad (3)$$

where z-transform of matrix sequence $\{H(\ell)\}$ is defined by polynomial matrix $H(z) = \sum H(\ell)z^{-\ell}$ and so on. This equation has a similar form as the case of flat fading [5], but SVD-based method could not be applied since existing canonical forms (e.g., in [20]) does not assure the orthogonality of weights, and even if $H(z)$ is decomposed in at each point, elements of left and right singular value vectors are not guaranteed to be a rational function of z (hence it could not be realize TDL structure). Therefore, in the next section, we adopt the strategy to search the best solution from the set of polynomial vectors. Those relations are depicted in Fig. 3.

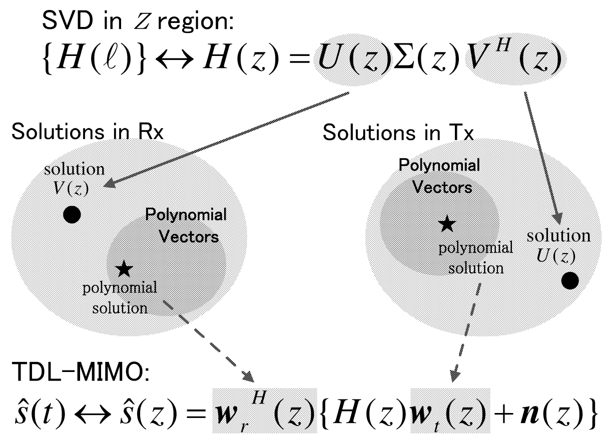


Fig. 3 SVD-based expression of channel matrix in z region and optimum solutions.

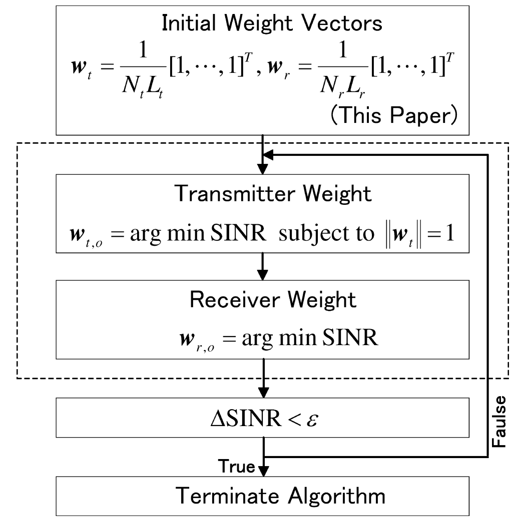


Fig. 4 Flowchart of weight optimization process.

3. Maximum SINR Design Method of TDL-MIMO Systems

While adaptive array antennas require the weight optimization only in one side of the transmitter or the receiver [21], [22], in MIMO systems, the weight design of both ends is necessary to fully utilize the CSI in the transmitter. In this section, a simple procedure for the maximum SINR design of TDL-MIMO systems based on elementary engineering mathematics is presented.

First, to simplify the notation, we introduce the stacked expressions of transmitter and receiver weights:

$$\begin{aligned} \boldsymbol{w}_t &= [\boldsymbol{w}_t^T(0),\ \cdots,\ \boldsymbol{w}_t^T(L_t-1)]^T \\ \boldsymbol{w}_r &= [\boldsymbol{w}_r^T(0),\ \cdots,\ \boldsymbol{w}_r^T(L_t-1)]^T. \end{aligned}$$

Those weights are determined to maximize the SINR of the output signal $\hat{s}(t)$ in equation (1), and shown below

$$\mathrm{SINR} = \frac{|\boldsymbol{w}_t^H H_s^H \boldsymbol{w}_r|^2 P_S}{\boldsymbol{w}_t^H R_{i,r} \boldsymbol{w}_t P_S + \|\boldsymbol{w}_r\|^2 P_N} \tag{4}$$

$$= \frac{|\boldsymbol{w}_r^H H_s \boldsymbol{w}_t|^2 P_S}{\boldsymbol{w}_r^H R_{i,t} \boldsymbol{w}_r P_S + \|\boldsymbol{w}_r\|^2 P_N} \tag{5}$$

where P_S and P_N denote the average powers of signal and noise. In above equation, two types of expression are used for the convenience of optimization. The N_rL_r-by-N_tL_t matrix H_s is commonly used in both expressions, and its (ℓ_r, ℓ_t)-th block is defined by the next N_r-by-N_t matrix

$$[H_s]_{\ell_r,\ell_t} = \begin{cases} H(\ell_t-\ell_r) & \ell_t-\ell_r \in \{0,\ \cdots,\ L-1\} \\ O_{N_r,N_t} & \ell_t-\ell_r \notin \{0,\ \cdots,\ L-1\} \end{cases} \tag{6}$$

where $O_{M,N}$ is the M-by-N null matrix. The notation $R_{i,t}$ denotes the N_rL_r-by-N_rL_r Hermitian matrix with $(\ell_{r,0},\ \ell_{r,1})$-th block represented by the following N_r-by-N_r matrix

$$\begin{aligned} [R_{i,t}]_{\ell_{r,0},\ell_{r,1}} = &\sum_{\ell_{t,0},\ell_{t,1}=0}^{L_t-1} \sum_{\ell_0,\ell_1=0}^{L-1} (1-\delta_{\ell_{t,0}+\ell_0-\ell_{r,0},0}) \\ &\times (1-\delta_{\ell_{t,1}+\ell_1-\ell_{r,1},0})\, \delta_{\ell_{t,0}+\ell_0-\ell_{r,0},\ell_{t,1}+\ell_1-\ell_{r,1}} \\ &\times H(\ell_0)\boldsymbol{w}_t(\ell_{t,0})\boldsymbol{w}_t^H(\ell_{t,1})H^H(\ell_1) \end{aligned} \tag{7}$$

Similarly, the Hermitian matrix $R_{i,r}$ with the size of N_tL_t-by-N_tL_t has $(\ell_{r,0}, \ell_{r,1})$-th block shown by the N_t-by-N_t matrix as follows

$$\begin{aligned} [R_{i,r}]_{\ell_{t,0},\ell_{t,1}} = &\sum_{\ell_{r,0},\ell_{r,1}=0}^{L_r-1} \sum_{\ell_0,\ell_1=0}^{L-1} (1-\delta_{\ell_{t,0}+\ell_0-\ell_{r,0},0}) \\ &\times (1-\delta_{\ell_{t,1}+\ell_1-\ell_{r,1},0})\, \delta_{\ell_{t,0}+\ell_0-\ell_{r,0},\ell_{t,1}+\ell_1-\ell_{r,1}} \\ &\times H^H(\ell_0)\boldsymbol{w}_r(\ell_{r,0})\boldsymbol{w}_r^H(\ell_{r,1})H(\ell_1) \end{aligned} \tag{8}$$

In Eqs. (7) and (8), $\delta_{m,n}$ denotes the Kronecker delta. Those equations show that two equivalent quadratic forms $\boldsymbol{w}_t^H R_{i,t} \boldsymbol{w}_t$ and $\boldsymbol{w}_r^H R_{i,r} \boldsymbol{w}_r$ in the denominators of Eqs. (4) and (5) are got by expressing the sum of energies

$$\boldsymbol{w}_r^H(\ell_{r,0})H(\ell_0)\boldsymbol{w}_t(\ell_{t,0})\boldsymbol{w}_t^H(\ell_{t,1})H^H(\ell_1)\boldsymbol{w}_r(\ell_{t,0})$$

concerning all interference signals, about $\boldsymbol{w}_t$ and $\boldsymbol{w}_r$, respectively.

The closed form expression of the solutions of the above equations is not yet derive since the cost functions (4) and (5) have a complicated form (to the best of authors' knowledge, there is no reference describing the answer of this type of problem), hence we consider solving them by iterative procedure. The whole process is divided into two stages for the alternate determination of $\boldsymbol{w}_t$ and $\boldsymbol{w}_r$: The fundamental strategy is to fix one side (of Tx or Rx) while the optimization of other side, and repeat this operation until convergence. The actual steps for the weight derivation are as follows (they are shown in Fig. 4):

(i) Initial conditions:

Initialize transmitter and receiver weights by giving certain nonzero values. In this paper, we adopt

$$\boldsymbol{w}_t = 1/\sqrt{N_t L_t}[1, \cdots, 1]^T$$
$$\boldsymbol{w}_r = 1/\sqrt{N_r L_r}[1, \cdots, 1]^T,$$

though any nonzero values could be used. The choice of initial conditions to optimize the convergence speed is under investigation.

(ii) Update of $\boldsymbol{w}_t$:
For a fixed receiver weight vector $\boldsymbol{w}_r$, the updated version of $\boldsymbol{w}_t$ which maximize the SINR in Eq. (4) is calculated. Since SINR becomes higher as the norm of $\boldsymbol{w}_t$ becomes larger, we give a constraint $\|\boldsymbol{w}_t\| = 1$ to keep the transmission power constant. By using the method of Lagrange multiplier, the optimal weight vector $\boldsymbol{w}_{t.o}$ is derived as follows (the notation I_M denotes the identity matrix of M-th order):

$$\boldsymbol{w}_{t.o} = \frac{\boldsymbol{w}_t^H R_{i,r} \boldsymbol{w}_t P_S + \|\boldsymbol{w}_r\|^2 P_N}{\boldsymbol{w}_t^H H^H \boldsymbol{w}_r P_S} \times \left(R_{i,r} + \|\boldsymbol{w}_r\|^2 I_{N_t} P_S\right)^{-1} H_s^H \boldsymbol{w}_r \tag{9}$$

The above equation still contains $\boldsymbol{w}_t$ in the right-hand side, but the value of $\boldsymbol{w}_t$ does not affect the direction of $\boldsymbol{w}_{t,o}$. Therefore, we can get correct solution of the above constrained problem after normalizing $\boldsymbol{w}_{t,o}$.

(iii) Update of $\boldsymbol{w}_r$:
For a fixed transmitter weight vector $\boldsymbol{w}_t$, the updated version of $\boldsymbol{w}_r$ which maximize the SINR in equation (5) is calculated by using following equation:

$$\boldsymbol{w}_{r.o} = \frac{\boldsymbol{w}_r^H R_{i,t} \boldsymbol{w}_r P_S + \|\boldsymbol{w}_r\|^2 P_N}{\boldsymbol{w}_r^H H \boldsymbol{w}_t P_S} \times \left(R_{i,t} + I_{N_r L_r} P_N\right)^{-1} H_s \boldsymbol{w}_t \tag{10}$$

Though the norm of $\boldsymbol{w}_r$ has no influence on SINR, this vector is also normalized to keep the receiver power constant.

(iv) Computation of SINR:
Compute SINR using weights derived in (ii) and (iii), then compare it with that of the previous iteration. If the change $\Delta\text{SINR} = |\text{new SINR} - \text{old SINR}|/|\text{old SINR}|$ is within a previously given threshold ε (this paper: 10^{-3}), then the algorithm is terminated. If not, go back to (ii) and continue the weight update process.

The convergence of above steps is not theoretically assured, but by our experience, it successfully reaches to a good solution if an adequate unit length vector is given as an initial value. Even if it has a few possibility of divergence, the fatal breakdown of communication could be avoided by employing the best solution during the convergence. Of course, the best solution to this problem is the derivation of the closed form expressions of the optimal weights, and this is left as a future work.

The matrix inversion in (ii) and (iii) could be replaced by a suitable adaptive algorithm which reduces the computational cost, but we don't adopt them to avoid the influence of the choice of parameters and to know bare performance of the system.

4. Performance Analysis

In this section, the performance of TDL-MIMO systems designed by using the proposed method is investigated through computer simulations.

The simulation conditions are show in Table 1. The elements of $H(\ell)$ are assumed to be i.i.d., and channel matrix sequences are generated by truncating the impulse response of a Rayleigh fading model with the exponential power delay profile [23] by L. Namely,

$$P_H(\tau) = E[|H_{n_r,n_t}(\ell)|^2] = \frac{1}{\sigma_S} \sum_{\ell=0}^{L-1} \exp\left(-\frac{\ell}{\sigma_S}\right) \delta(\tau - \ell)$$

where delay spread is $\sigma_S = 3$ (the truncated impulse response is an unrealistic assumption since actual length of delay profiles is infinite, but it is suitable to make the influence of channel duration clear). The quality of the output signal $\hat{s}(t)$ is evaluated using output SINR defined by following equations:

$$\rho = \frac{E[\hat{s}(t)s^*(t)]}{\sqrt{E[|\hat{s}(t)|^2]E[|s(t)|^2]}} \tag{11}$$

$$\text{SINR} = \frac{|\rho|^2}{1 - |\rho|^2} \tag{12}$$

and the input SNR ($= P_S/P_N$) is hold to 20 dB during simulations. The estimation error of CSI is not considered here. In actual situation, particularly in case of low SNR, the correct estimation of $\{H(\ell)\}$ using enough number of samples and/or adoption of adequate coding is required.

In this study following four types of system are considered: Type 1: Systems without delay units [13] Type 2: Systems with TDL structure in the receiver side (Rx-TDL) Type 3: Systems with TDL structure in the transmitter side (Tx-TDL) Type 4: Systems with TDL structure in both transmitter and receiver sides (Tx-Rx-TDL). The details are summarized in Table 2 for some experiments.

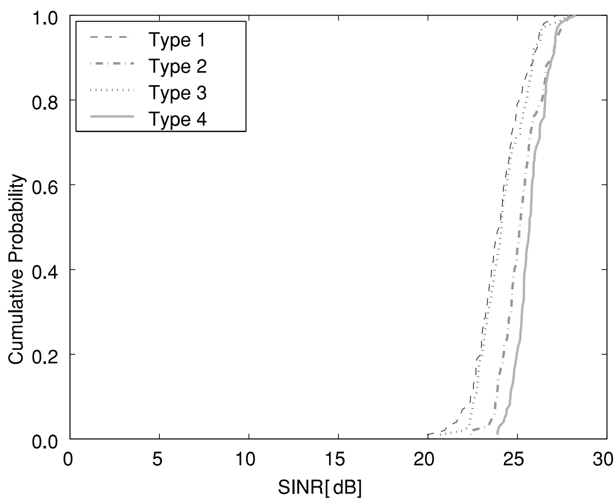
In Fig. 5, we have plotted distribution functions of

Table 1 Simulation conditions.

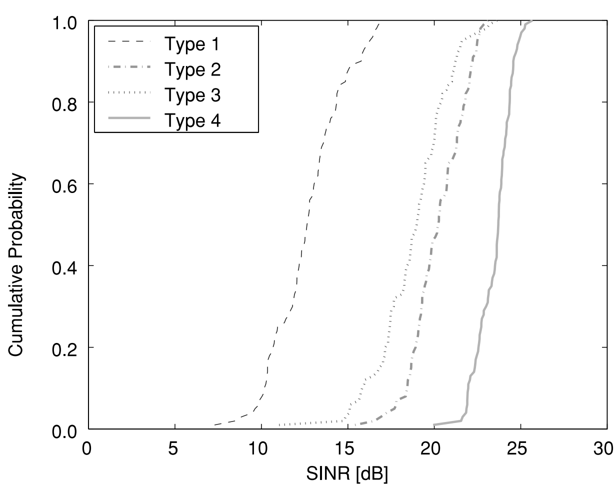
Modulation Scheme	BPSK
input SNR	20dB
Delay Profile	Exponetial
Delay Spread (default)	$\sigma_S = 3$
Snup Shots	4096
Channel Samples	100
Threshold	$\varepsilon = 10^{-3}$

Table 2 Four types of MIMO systems considered in this study. Array size is fixed to $(N_r, N_r) = (4, 4)$.

	Structure	(L_t, L_r) Fig. 5, Fig. 6 Fig. 7, Fig. 10	Fig. 8, Fig. 12
Type 1	w/o delay units [13]	$(1, 1)$	$(1, 1)$
Type 2	Rx-TDL	$(1, 3)$	$(1, 5)$
Type 3	Tx-TDL	$(3, 1)$	$(5, 1)$
Type 4	Tx-Rx-TDL	$(3, 3)$	$(3, 3)$



(a) $L = 2$



(b) $L = 14$

Fig. 5 Distribution functions of output SINR for various MIMO schemes.

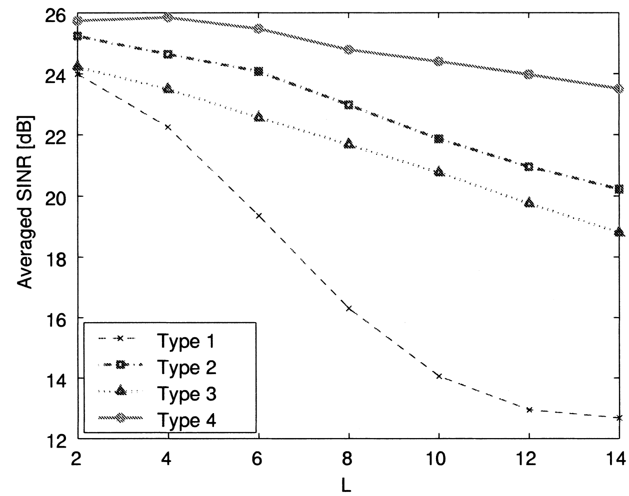


Fig. 6 Duration of FSF channel L versus averaged SINR.

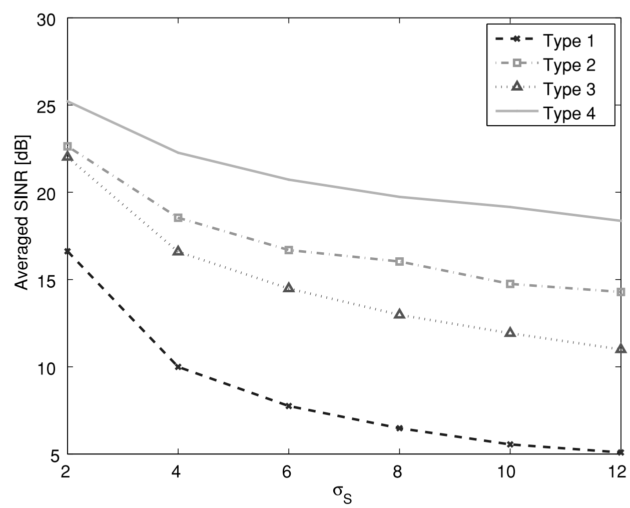


Fig. 7 Delay spread of FSF channel σ_S versus averaged SINR.

SINR for above four systems, where each of them is drawn after 10^2 independent trials changing $\{H(\ell)\}$. From this figure, it is observed that the systems equipped with TDL (Type 2~Type 4) has higher ability than the delay-less structure (Type 1). Among three TDL-MIMO systems, Type 4 gives the best performance. The ability of the system with TDL structure in receiver side (Type 2) overcomes that of Type 3 if the number antennas and TDL length are same, and a possible reason is that the adoption of TDL structure in the receiver has a spatio-temporal averaging effect of the thermal noise in addition to interference cancellation ability. The difference of four schemes are magnified as the length L of delay profile increases, and it shows the advantage of TDL structure in actual FSF environment with long channel duration.

This fact is more clearly seen from Fig. 6, which shows how the (output) averaged SINR (in all the examples in this section except Fig. 12, it is calculated over 10^2 samples of channel matrices) degrades as the duration L becomes longer. While the averaged SINR of Type 1 system significantly decrease after L exceeds the degrees of freedom of the system ($L - 1 > N_t + N_r - 2 = 6$) [13], the reduction curves of other structures are quite moderate.

Figure 7 plots the relation between averaged SINR and delay spread σ_S instead of channel duration L. The channel duration is fixed to $L = 14$, but σ_S changes from 2 to 14. Though the degradation of Type 1 system is not significant as the case of Fig. 6, TDL-MIMO systems have sufficient advantage against Type 1 structure, and keep averaged SINR more than 10dB for large value of σ_S.

Another comparison between Type 2~4 could be made

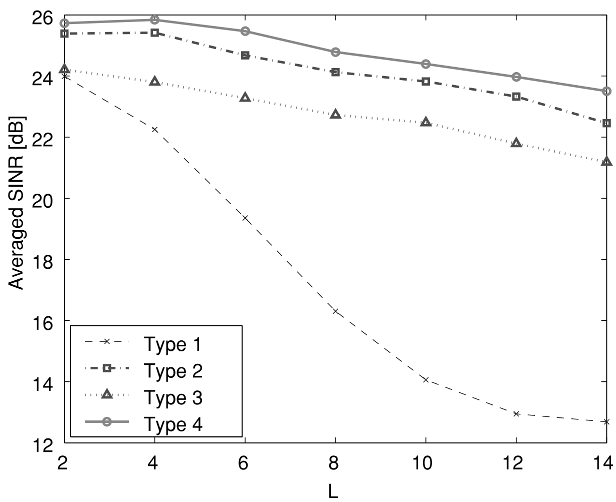


Fig. 8 Duration of FSF channel L versus averaged SINR.

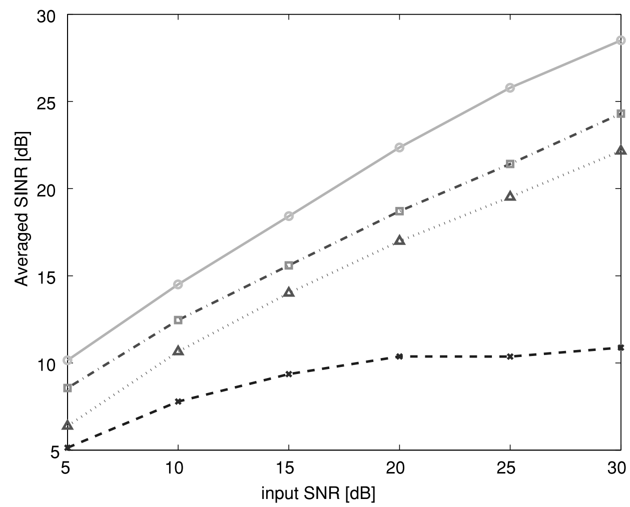
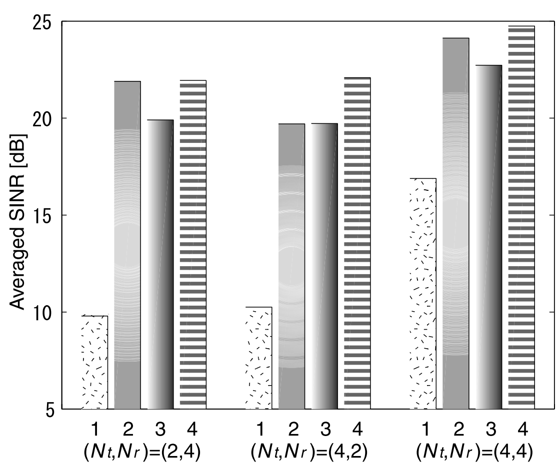


Fig. 10 Input SNR versus output averaged SINR.

(N_t, N_r)	(L_t, L_r)			
	Type 1	Type 2	Type 3	Type 4
(2, 4)	(1,1)	(1,4)	(7,1)	(3,3)
(4, 2)	(1,1)	(1,7)	(4,1)	(3,3)
(4, 4)	(1,1)	(1,5)	(5,1)	(3,3)

(a) Structures of MIMO systems.



(b) Simulation results.

Fig. 9 Comparison of various MIMO schemes.

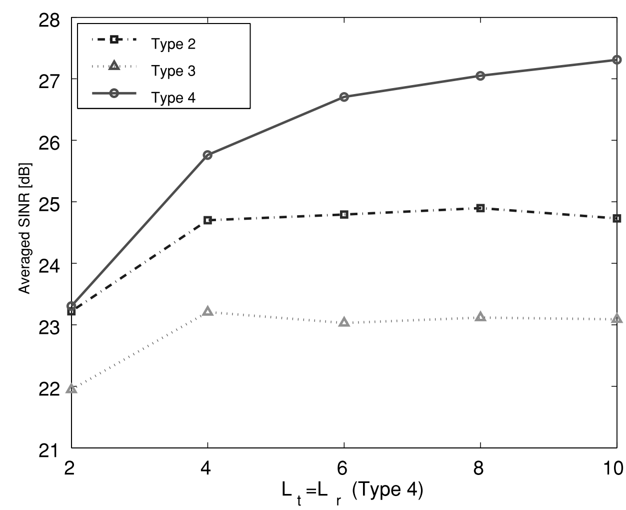


Fig. 11 The length of delay units $L_t = L_r$ in Type 4 versus averaged SINR.

by setting the total number of weights N_w and delay units N_d to be constant, namely, (L_t, L_r) is $(1, 5)$ for Type 2, $(5, 1)$ for Type 3, and $(3, 3)$ for Type 4 ($N_w = 24, N_d = 16$). This experiment gives an answer to the question "If we can use same number of resources (weights and delay units), how to allocate them to the transmitter and receiver sides?" As seen from the result of Fig. 8, the answer is that the equable distribution of them to both sides has an advantage. This is a suitable result from the point of computational load (It is defined by $Q = N_{itr} \cdot N_{mlp}$, where N_{itr} is the average number of iteration for convergence of SINR and $N_{mlp} = O(N_t^3 L_t^3 + N_r^3 L_r^3)$ denotes the number of complex multiplications for one iteration. Actual number of N_{mlp} depends on the algorithm to be adopted. For example, in the left column of Table 2 ($L_t = 3$ and/or $L_r = 3$), $Q = N_{itr} \cdot N_{mlp}$ is Type 1: $Q = 14 \cdot 128 = 1792$, Type 2: $Q = 36 \cdot 1792 = 64512$, Type 3: $Q = 42 \cdot 1792 = 75264$ and Type 4: $Q = 50 \cdot 3456 = 172800$, assuming $N_{mlp} = N_t^3 L_t^3 + N_r^3 L_r^3$.), since the one-sided concentration of weights accelerate the increase of operation number for the matrix inversion. It is also shown that the cooperative operation of transmitter and receiver results in a better performance than one-side resource allocation, and this fact demonstrates the effectiveness of the proposed MIMO system equipped with TDL structure in both ends.

The situation becomes more complex, however, when the case of $N_t \neq N_r$. Figure 9 shows the comparison of MIMO schemes under the condition of constant number of weights and delay units, in which the number n (=1∼4) in (b) under each bar denotes type of MIMO system given in (a). This figure shows it is better to give more resources to the side with larger number of array elements (But in this figure, Type 4 remains as the best choice). To introduce the

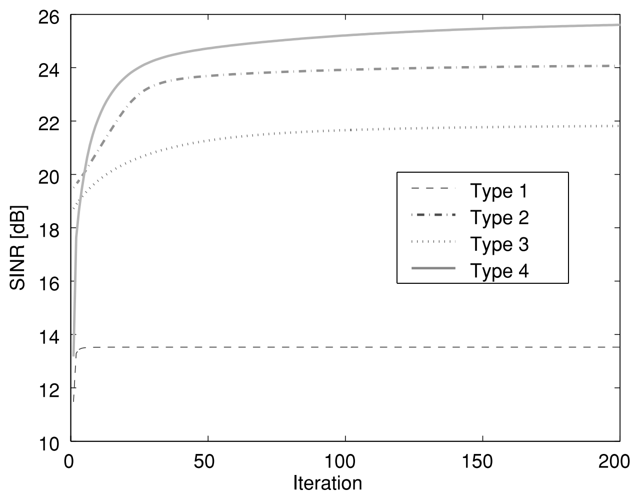


Fig. 12 Leaning curves of design algorithms for MIMO systems with and without TDL structure.

general rule of the resource distribution, further investigations are required.

Figure 10 plots the curves of input SNR versus output averaged SINR for $L = 14$ and $\sigma_S = 4$, where conventional spatial filtering method without TDL strucre (Type 1) suffers from floor, while other spatio-temporal methods (Type 2 $\sim$ 4) continue to improve as increase of input SNR.

The relation of the maximum length of delay units and averaged SINR is shown in Fig. 11 for $L = 8$. The abscissa of this figure denotes $L_t = L_r$ in Type 4 system, and the same number of total weights and delay units are used in Type 2 and 3 systems (The maximum delay length is $L_t + L_r - 1$ for the receiver in Type 2 and for the transmitter in Type 3). Array size is fixed to $(N_t, N_r) = (4, 4)$. While averaged SINR of Type 2 and 3 systems has almost no change after $L_t = L_r = 4$, Type 4 still continues to improve, which reinforces the superiority of this scheme.

Figure 12 plots an example of the learning curves of the proposed design algorithms for MIMO systems. The SINR of TDL-MIMO systems converge slowly compared to that of Type 1 (remark that the vertial axes is 'instantaneous' SINR). Particularly, Type 4 system which requires large scale matrix inversion in both transmitter and receiver sides is, in spite of good performance, quite inefficient in its computational cost. Hence a fast algorithm is strongly required for the proposed method to be used in the actual mobile communications.

5. Conclusions

In this paper, we have presented a mathematically simple method for the maximum SINR design of TDL-MIMO communication systems and analyzed their performance under FSF. For the design of MIMO systems, a pair of operations — the optimization of transmitter weights fixing the receiver side, and the calculation of optimum receiver weights keeping the transmitter weights constant — is repeated until SINR converge. The computer simulations show that the TDL structure is effective, and their abilities become clear as the duration of the delay profile becomes longer. The good choice of system structure is the equi-distribution of the weights and delay units to both arrays, and the preferential allocation of them to receiver side follows it. The disadvantage of the proposed scheme is in its heavy computational load (particularly when the both arrays have TDL structure), but conversely, it is obvious that they possess high potential to mitigate the influence of FSF.

The future work is the improvement of the convergence speed and the reduction of computational cost of the proposed design algorithm. Other problems such as closed form expression of optimum weights, design of orthogonal weights for the utilization of substreams, are also important topics which should be investigated. The extension of the proposed method to multiuser MIMO system working under FSF is also required from the practical point.

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