PAPER

Non-iterative Algorithm of MIMO Adaptive Array Based on Correlation Matrix Including Parasitic Antennas

Naoki HONMA[†a)], ***Member***, **Kentaro NISHIMORI**[††], ***Senior Member***, **Takefumi HIRAGURI**[†††], ***and*** **Yoshitaka TSUNEKAWA**[†], ***Members***

SUMMARY Parasitic antenna elements with tunable terminations can be used for interference suppression in multi-antenna systems without using the degrees of freedom. The authors have proposed a fast non-iterative algorithm for optimizing the termination conditions. However, this method cannot be used for suppressing the interference from unknown systems since it requires the channel state information. In this paper, a fast non-iterative algorithm based on the correlation matrix, which can be obtained even from unknown interference sources, is proposed for the multi-antenna system with parasitic antenna elements. The correlation matrix including both receiving and parasitic antennas can be estimated from a few observations of the signals even without receiving signals at the parasitic antenna. By using this correlation matrix, the power of the interference with the arbitrary termination conditions can be easily estimated. Therefore, the termination condition, which minimizes the interference power, can be calculated without knowledge of the channel state information or additional estimations. The results of a numerical analysis indicate that proposed method works well in suppressing the interference without the perfect channel state information.

key words: *adaptive array, tunable reactance device, MIMO, correlation matrix*

1. Introduction

It is well known that antenna-systems exploiting tunable parasitic antennas can enhance a wireless communication quality, such as the reception power improvement, interference cancellation, and so on [1]–[4]. These antennas can be recognized as a kind of analog adaptive antenna, but the arbitrary excitation weight at the antenna element cannot be given since the current at the antenna element strongly depends on the mutual coupling. This deteriorates the performance in interference suppression compared to that of the digital adaptive array antenna. Nevertheless, they have relatively simple hardware compared with the digital adaptive array antennas. This feature allows to use many more parasitic antennas, and the effective degree of freedom can be greatly increased [5], [6].

However, it is difficult to predict the channel behavior against the termination conditions at the parasitic antennas, since the relationship between the channel response and the termination impedance are known to be non-linear. The steepest gradient algorithm is commonly used for optimizing the termination condition of the parasitic antennas [1], [2]. In this algorithm, the gradient of the evaluation function, e.g. signal-to-interference ratio (SIR), is estimated by varying the termination conditions. This scheme requires significant numbers of iterative measurements to obtain optimum radiation patterns or desired channels. The work [7] presented a fast control method of the parasitic antennas by measuring the impedance matrix of the array antenna including the termination ports at the parasitic elements. Though the fast control of the radiation pattern can be achieved, this method cannot be applied to the interference suppression since the response of unknown interference cannot be estimated.

Authors have proposed a deterministic solution for controlling analog adaptive antenna, which requires only a few channel observations [8], [9]. In this method, the idea of a 'parasitic channel', which is observed at the parasitic antenna, is newly introduced. The parasitic channel is an invisible channel but can be estimated by using this method. However, this method requires the channel state information (CSI) between the receiving antenna and interference source for suppressing the interference. Since the CSI estimation requires the known training signal whose information is shared between the transmitter and receiver in advance, it is difficult to obtain the CSI from unknown signals that can arrive from other wireless systems.

In this paper, a fast non-iterative algorithm based on correlation matrix for multiple-input multiple-output (MIMO) adaptive array is proposed. This method does not require the CSI but the correlation matrix, which is obtained from the received signals. The correlation matrix including both receiving antennas and parasitic antennas can be estimated from a few observations of the signals even without receiving signals at the parasitic antenna. It is shown how this method works in suppressing the interference, and the simulations are conducted to validate this method.

Section 2 describes the system model of the multi-antenna, which is considered in this paper. Also, the proposed algorithm is described in detail. In Sect. 3, the simulation results are indicated, and show this method is effective in enhancing the channel data rate even without any iterative channel measurements.

Manuscript received October 1, 2012.
Manuscript revised March 11, 2013.

[†]The authors are with Iwate University, Morioka-shi, 020-8551 Japan.
[††]The author is with Niigata University, Niigata-shi, 950-2181 Japan.
[†††]The author is with Nippon Institute of Technology, Saitama-ken, 345-8501 Japan.
a) E-mail: honma@iwate-u.ac.jp

DOI: 10.1587/transcom.E96.B.1945

2. System Model and Proposed Algorithm

The method shown in [9] is the easiest and most secure way to predict the channel response with arbitrary termination conditions in the parasitically steerable arrays. However, this method cannot be applicable to the interference from other wireless systems since the CSI between the interference source and the receiving antennas is not available. Although the CSI is not available, the stochastic characteristics of the interference can be observed.

In the proposed method, the correlation matrix calculated from the received signals at the receiving antennas is observed and exploited instead of using the CSI. The correlation matrix including both receiving antennas and parasitic antennas can be estimated from several training rounds. If this adaptive antenna is used for suppressing the interference, the correlation matrix must be observed during the absence of the desired signal. Also, the same interference signals must arrive repeatedly. This precondition can be obtained in the various wireless systems. For example, a guard time is set in the actual systems in order to avoid the collision between the desired users. The interference cancellation method with adaptive array by using such guard is proposed [10]. In the proposed method, guard time can be exploited to obtain the correlation matrix of the interference signals.

2.1 System Model

Figure 1 shows the configuration of the multi-antenna system considered in this paper. The number of the receiving and parasitic antennas are M_r and M_p, respectively. The antennas shown in this figure are dipoles, but the following discussion can be applied to arbitrary configurations. Each parasitic antenna is terminated by a tunable reactance, z_i. Here, i is the index number of the parasitic antenna. These arrays are placed close together since the mutual coupling between the receiving and parasitic elements is needed to form the radiation pattern by controlling the termination condition at each parasitic element. In this model, the intereference signals arriving at the receiving antenna system are considered. Figure 2 shows the circuit model used in this discussion. This model considers the receiving and parasitic elements. Term $\boldsymbol{S}_R$ is the S-parameter matrix, including the receiving and parasitic elements. $\boldsymbol{S}_R$ can be defined as,

$$\boldsymbol{S}_R = \begin{pmatrix} \boldsymbol{S}_{RR} & \boldsymbol{S}_{RP} \\ \boldsymbol{S}_{PR} & \boldsymbol{S}_{PP} \end{pmatrix}, \tag{1}$$

where R and P represents the receiving element and parasitic element ports, respectively. $\boldsymbol{S}_R$ is assumed to be known and stable. In an actual operation, this can be measured in advance of the following process. $\boldsymbol{b}_0$ denotes the complex vector that represents the interference signalsat the antenna elements including the parasitic elements, and it can be observed only when all ports are terminated by the reference impedance, z_0. $\boldsymbol{b}_0$ can be divided into vectors $\boldsymbol{b}_{R0}$ and $\boldsymbol{b}_{P0}$

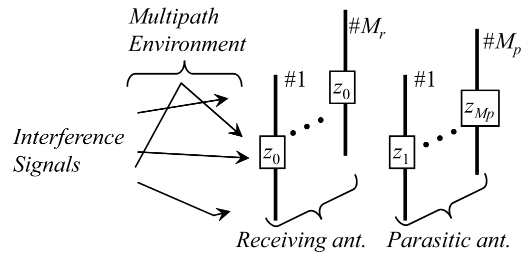


Fig. 1 Antenna system configuration.

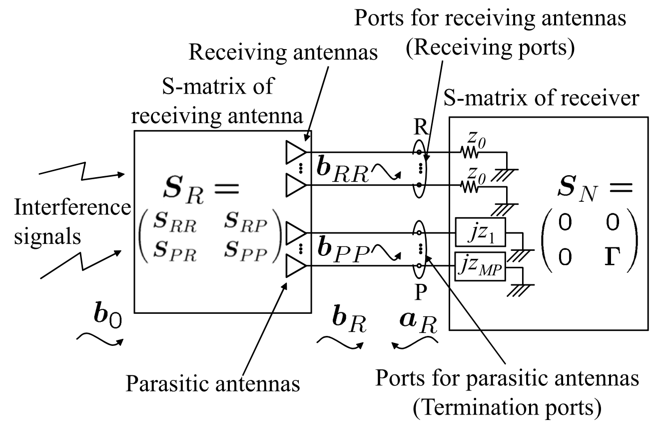


Fig. 2 Equivalent circuit model of the multi-antenna system.

at the receiving and parasitic elements, respectively. $\boldsymbol{b}_R$ represents the complex output vector observed at the antenna ports when the antennas are terminated and connected to the receivers. $\boldsymbol{b}_R$ can also be divided into vectors $\boldsymbol{b}_{RR}$ and $\boldsymbol{b}_{PP}$ observed at the receiver ports and parasitic element ports, respectively. $\boldsymbol{a}_R$ denotes signals that are reflected from the receivers and terminations to the antenna elements, and can be divided into $\boldsymbol{a}_{RR}$ and $\boldsymbol{a}_{PP}$. These matrices and vectors must satisfy

$$\begin{pmatrix} \boldsymbol{b}_{RR} \\ \boldsymbol{b}_{PP} \end{pmatrix} = \begin{pmatrix} \boldsymbol{S}_{RR} & \boldsymbol{S}_{RP} \\ \boldsymbol{S}_{PR} & \boldsymbol{S}_{PP} \end{pmatrix} \begin{pmatrix} \boldsymbol{a}_{RR} \\ \boldsymbol{a}_{PP} \end{pmatrix} + \begin{pmatrix} \boldsymbol{b}_{R0} \\ \boldsymbol{b}_{P0} \end{pmatrix} \tag{2}$$

$$\boldsymbol{a}_{PP} = \boldsymbol{\Gamma}\boldsymbol{b}_{PP}. \tag{3}$$

$\boldsymbol{\Gamma}$ is the termination condition and defined as,

$$\boldsymbol{\Gamma} = \begin{pmatrix} \Gamma_1 & & 0 \\ & \ddots & \\ 0 & & \Gamma_{Mp} \end{pmatrix}, \tag{4}$$

where Γ_i is the reflection coefficient of the i-th reactive termination; it is defined as $\Gamma_i = (z_i - z_0)/(z_i + z_0)$, (z_0: reference impedance). When the internal impedance of the receivers is ideal, i.e. equal to the reference impedance, z_0, $\boldsymbol{a}_{RR}$ can be assumed to be $\boldsymbol{0}$. From (2) and (3), the observed vector at the receiving ports can be expressed as,

$$\begin{aligned} \boldsymbol{b}_{RR} &= \boldsymbol{b}_{R0} + \boldsymbol{S}_{RP}\boldsymbol{\Gamma}(\boldsymbol{I}_{Mp} - \boldsymbol{S}_{PP}\boldsymbol{\Gamma})^{-1}\boldsymbol{b}_{P0} \\ &= \boldsymbol{A}\boldsymbol{b}_0, \end{aligned} \tag{5}$$

where,

$$E\left[\begin{pmatrix}\boldsymbol{b}_{RR,1}\\ \vdots \\ \boldsymbol{b}_{RR,K}\end{pmatrix}(\boldsymbol{b}_{RR,1}^H,\ldots,\boldsymbol{b}_{RR,K}^H)\right] = E\left[\begin{pmatrix}\boldsymbol{A}_1\boldsymbol{b}_0\\ \vdots \\ \boldsymbol{A}_K\boldsymbol{b}_0\end{pmatrix}(\boldsymbol{b}_0^H\boldsymbol{A}_1^H,\ldots,\boldsymbol{b}_0^H\boldsymbol{A}_K^H)\right] = \begin{pmatrix}\boldsymbol{A}_1\\ \vdots \\ \boldsymbol{A}_K\end{pmatrix}E[\boldsymbol{b}_0\boldsymbol{b}_0^H](\boldsymbol{A}_1^H,\ldots,\boldsymbol{A}_K^H) \tag{8}$$

$$\boldsymbol{R}_0 = E[\boldsymbol{b}_0\boldsymbol{b}_0^H] = \begin{pmatrix}\boldsymbol{A}_1\\ \vdots \\ \boldsymbol{A}_K\end{pmatrix}^+ E\left[\begin{pmatrix}\boldsymbol{b}_{RR,1}\\ \vdots \\ \boldsymbol{b}_{RR,K}\end{pmatrix}(\boldsymbol{b}_{RR,1}^H,\ldots,\boldsymbol{b}_{RR,K}^H)\right](\boldsymbol{A}_1^H,\ldots,\boldsymbol{A}_K^H)^+ \tag{9}$$

$$\boldsymbol{A} = \left[\boldsymbol{I}_{Mr}\ \ \boldsymbol{S}_{RP}\boldsymbol{\Gamma}(\boldsymbol{I}_{Mp} - \boldsymbol{S}_{PP}\boldsymbol{\Gamma})^{-1}\right]. \tag{6}$$

Here, $\boldsymbol{I}_{Mr}$ and $\boldsymbol{I}_{Mp}$ are $M_r \times M_r$ and $M_p \times M_p$ identity matrices, respectively. $\boldsymbol{A}$ transfers the signal arriving at the antennas to the observed signal at the receivers, and depends on the termination condition, $\boldsymbol{\Gamma}$. The size of $\boldsymbol{A}$ is $M_r \times (M_r + M_p)$. The correlation matrix observed at the receiver is,

$$\begin{aligned} E[\boldsymbol{b}_{RR}\boldsymbol{b}_{RR}^H] &= \boldsymbol{A}E[\boldsymbol{b}_0\boldsymbol{b}_0^H]\boldsymbol{A}^H \\ &= \boldsymbol{A}\boldsymbol{R}_0\boldsymbol{A}^H, \end{aligned} \tag{7}$$

where, $(\cdot)^H$ and $E[\cdot]$ are Hermitian transposition and ensemble averaging, respectively. $\boldsymbol{R}_0$ is the correlation matrix including the components of both receiving and parasitic antennas. It can be seen that the correlation matrices with arbitrary termination conditions can be obtained from (7) if $\boldsymbol{R}_0$ is known. Since the sum of the diagonal component of $E[\boldsymbol{b}_{RR}\boldsymbol{b}_{RR}^H]$ corresponds to the received power, the termination condition that minimizes the received power, is needed to suppress the interference when the received signal contains only the interference components.

2.2 Estimation Method of Correlation Matrix

In this subsection, technique for estimating the correlation matrix, $\boldsymbol{R}_0$, is described. Let us assume that $\boldsymbol{S}_R$ is known. First step is giving the various termination conditions and observing the signals at the receiver. K sets of non-identical termination conditions, defined as $\boldsymbol{A}_1,\ldots,\boldsymbol{A}_K$, are given by (6) and they are assumed to be known. The received signal vectors corresponding to the termination conditions are defined as $\boldsymbol{b}_{RR,1},\ldots,\boldsymbol{b}_{RR,K}$. Second step is estimation of the correlation matrix. The correlation matrix can be written as (8). From this equation, the correlation matrix, $\boldsymbol{R}_0$ can be estimated from the received signal vectors by using (9), where $(\cdot)^+$ represents a pseudo inverse operation.

In (9), the size of the matrices, i.e. $(\boldsymbol{A}_1^H,\ldots,\boldsymbol{A}_K^H)$ and its Hermitian transpose, needs to be considered to obtain the desired $\boldsymbol{R}_0$ in terms of the validity of the pseudo inverse operations. Since the size of $(\boldsymbol{A}_1^H,\ldots,\boldsymbol{A}_K^H)$ is $(M_r + M_p) \times KM_r$, KM_r needs to be equal to or larger than $(M_r + M_p)$ to obtain the correct solution. Therefore, the condition,

$$K \geq \frac{M_p}{M_r} + 1, \tag{10}$$

must be fulfilled. This means K sets of measurements with non-identical termination conditions are required. If the transmitted signal is interference, the termination condition that minimizes the interference power can be calculated by using the typical optimization methods, like [1], [2].

As described above, the interference power can be suppressed without any knowledge of the CSI. However, the same interference signal must arrive repeatedly since (9) assumes $E[\boldsymbol{b}_0\boldsymbol{b}_0^H]$ is always constant thruough K times observations. It is natural that the same signals are transmitted many times since many wireless systems employ frame formats that use identical preambles, and this feature can be used for obtaining (9).

2.3 Effect of Noise on Estimation Accuracy of Correlation Matrix

Here, it is also assumed that the noise caused by the receiver is much greater than the external noise arriving at the antenna elements [11]. When this noise is taken into account, the received signal observed at the receiver can be described as,

$$\boldsymbol{b}_{RR} = \boldsymbol{A}\boldsymbol{b}_0 + \boldsymbol{n}, \tag{11}$$

where, $\boldsymbol{n}$ represents the noise vector observed at the receivers. When the noise is modeled as zero-mean Gaussian white, the correlation matrix derived in (9) can be modified as,

$$\boldsymbol{R}_{0e} = \boldsymbol{R}_0 + \boldsymbol{R}_n, \tag{12}$$

where, $\boldsymbol{R}_{0e}$ is the estimated correlation matrix with errors. $\boldsymbol{R}_n$ is the component caused by the noise, and is described as (13), where $\boldsymbol{n}_1,\ldots,\boldsymbol{n}_K$ corresponds to the observed noise vector when K sets of termination conditions are applied. When the number of received symbols for each training round is large enough, (13) can be rewritten as,

$$\boldsymbol{R}_n = \sigma^2 \begin{pmatrix}\boldsymbol{A}_1\\ \vdots \\ \boldsymbol{A}_K\end{pmatrix}^+ (\boldsymbol{A}_1^H,\ldots,\boldsymbol{A}_K^H)^+, \tag{14}$$

where σ^2 is the expected noise power. Since the interference signals and noise are uncorrelated, the first two terms in $E[\cdot]$ of (13) become zero. In the following consideration, the estimation error is defined as,

$$\boldsymbol{J} = \|\boldsymbol{R}_n\|_F, \tag{15}$$

where $\|\cdot\|_F$ represents a Frobenius norm.

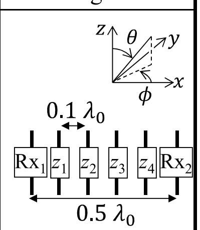
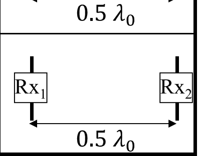
$$\boldsymbol{R}_n = \begin{pmatrix} \boldsymbol{A}_1 \\ \vdots \\ \boldsymbol{A}_K \end{pmatrix}^{+} E\left[\begin{pmatrix} \boldsymbol{b}_{RR,1} \\ \vdots \\ \boldsymbol{b}_{RR,K} \end{pmatrix} (\boldsymbol{n}_1^H, \ldots, \boldsymbol{n}_K^H) + \begin{pmatrix} \boldsymbol{n}_1 \\ \vdots \\ \boldsymbol{n}_K \end{pmatrix} (\boldsymbol{b}_{RR,1}^H, \ldots, \boldsymbol{b}_{RR,K}^H) + \begin{pmatrix} \boldsymbol{n}_1 \\ \vdots \\ \boldsymbol{n}_K \end{pmatrix} (\boldsymbol{n}_1^H, \ldots, \boldsymbol{n}_K^H) \right] (\boldsymbol{A}_1^H, \ldots, \boldsymbol{A}_K^H)^{+} \tag{13}$$

3. Simulation

3.1 Simulation Model

To verify the fundamental performance of the proposed estimation method, a simple antenna model was studied. Table 1 shows the antenna configurations considered in this numerical analysis. Here, 'Rx_1', 'Rx_2' and 'z_i' ($i = 1 \sim 4$) mean receiving antennas, #1, #2, and parasitic antennas, respectively. In the model (A) named as MIMO adaptive, all antennas are half-wavelength dipole, and, and this linear array antenna comprises two receiving and four parasitic antennas ($M_r = 2, M_p = 4$). The antenna distance is $0.1\lambda_0$, where λ_0 is the wavelength in a vacuum. The receiving antennas are placed at both sides of the array, and the parasitic antennas are placed between the receiving antennas. The model (B) is array without the parasitic antenna, and other dimensions are set to identical to model (A). The distance between receiving antennas is set to $0.5\lambda_0$ to offer the same aperture size.

Table 1 Antenna configuration for numerical analysis.

	Arrangement
(A) MIMO adaptive	
(B) Two dipoles	

For antenna (B), two schemes are adopted to compare the performances to antenna (A). One is no interference suppression, and the other is interference suppression using MMSE (Minimum Mean Squared Error) algorithms. In the MMSE algorithm, the weight matrix, $\boldsymbol{W}_{MMSE}$, is determined to minimize the error in the desired signals. The number of the training symbols for MMSE is set to 12, and is identical to that for observing $\boldsymbol{b}_{RR}$ in antenna (A).

Figure 3 shows the channel model and sketch of the transmitting and interference antennas. As a transmitter, two dipoles with spacing of one wavelength were used, and the interference signal was transmitted from a single dipole. The geometry-based statistical model was used for this calculation [12]–[14]. Around all of the antennas, the scatterers are arranged in the horizontal direction; this assumption comes from measurements in an actual environment [15]. To consider the statistical characteristics of the channel, 500 Monte Carlo simulations were performed in each condition. The S-parameters and radiation patterns of the receiving array are calculated by moment method, and they were taken into account in the following simulation. The operation frequency was set to 2.4 GHz.

3.2 Sheme of Proposed Algorithm

Figure 4 explains the flowchart of the actual procedures in this scheme. This scheme is applied only for antenna (A). Before observing the interference signals, we need to know S-parameter of the antenna. After evaluation of the S-parameter, $\boldsymbol{A}_1, \ldots, \boldsymbol{A}_K$ must be determined. They correspond to the termination conditions for K observations

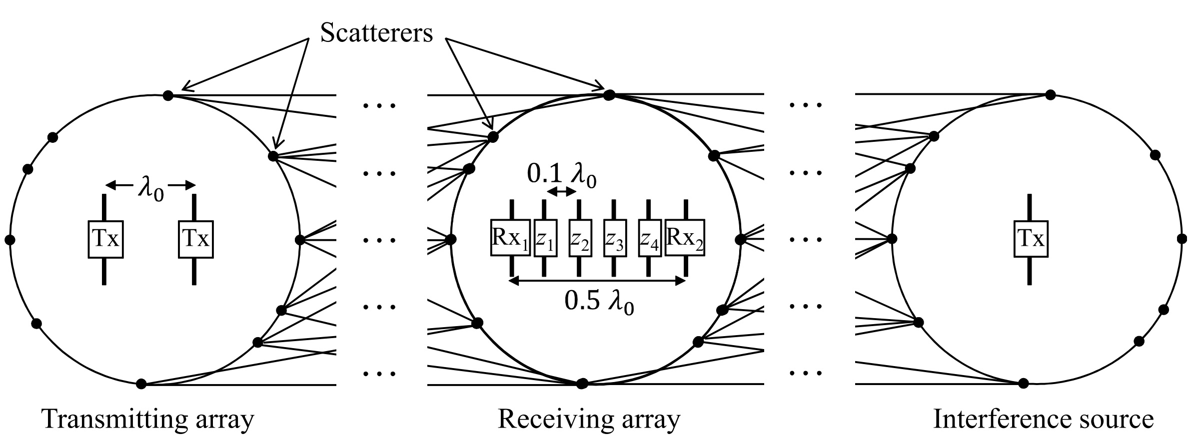


Fig. 3 Channel model and antenna arrangement.

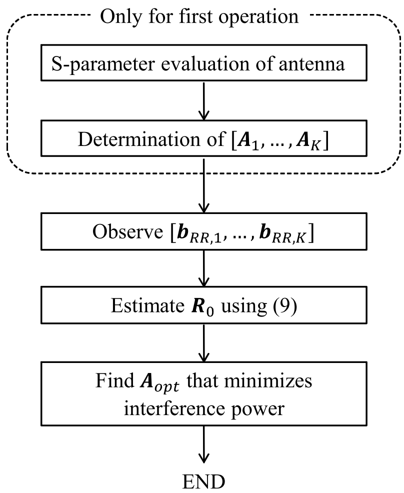


Fig. 4 Flowchart of proposed algorithm.

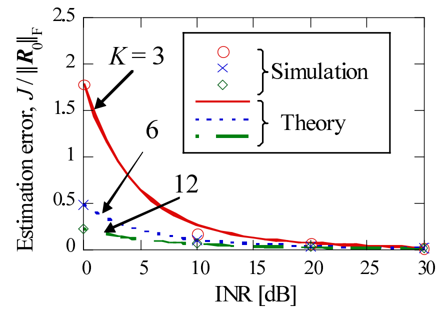


Fig. 5 Estimation error, J, versus INR.

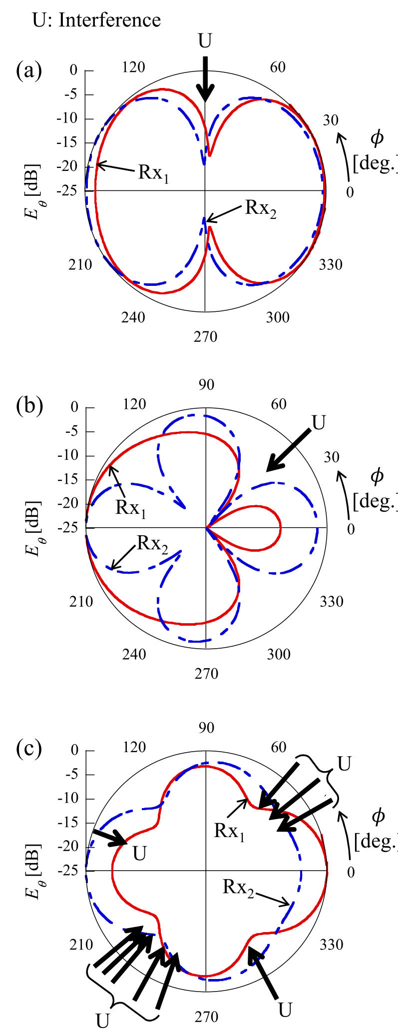


Fig. 6 Radiation patterns of MIMO adaptive antenna, (A), in xy-plane under interference condition: (a) single interference wave (U: $\phi = 90°$), (b) single interference wave (U: $\phi = 45°$), (c) multiple interference waves (U: $\phi = 30°, 40°, 50°, 160°, 220°, 225°, 230°, 240°, 250°$, and $300°$).

and must be non-identical to each other. They are chosen so as to minimize (15). In this simulation, we chose the best one out of randomly generated 100 sets of the termination conditions under the constraint of the reactance range, $-100j \leq z_i \leq 100j$, where the insertion resistance was neglected. It should be noted that these first two procedures are required only for the first operation since S-parameter and $\boldsymbol{A}_1, \ldots, \boldsymbol{A}_K$ are independent of the propagation characteristics.

The antenna observes the interference signals, $\boldsymbol{b}_1, \ldots, \boldsymbol{b}_K$, corresponding to the predetermined termination conditions, $\boldsymbol{A}_1, \ldots, \boldsymbol{A}_K$. Then, the correlation matrix, $\boldsymbol{R}_0$, is estimated by using (9). After this observation process, the reactance values of the terminations were optimized to minimize the interference power, and finally the desired termination condition, $\boldsymbol{A}_{opt}$, is obtained. In the simulation for antenna (A), the steepest gradient algorithm [2] is used as an optimization algorithm.

This scheme can work independently to the number of the interference sources since $\boldsymbol{R}_0$ can be obtained in the same manner as in (9). The analog adaptive antenna is effective in suppressing the multiple interferences [6], and this advantage is also obtained even when proposed scheme is used for such environment. However, the single interference case is studied in this paper for simplicity.

3.3 Result

Figure 5 indicates the estimation error of the correlation matrix shown in (15) versus INR. Here, it is assumed that only the interference signal is observed for this calculation, and the estimation error is normalized by Frobenius norm of the ideal correlation matrix, $\boldsymbol{R}_0$. The number of the trials is 500, and Rician factor is set to $-\infty$ dB. It can be seen that the estimation error decreases when the INR and the training rounds, K, are increased. From this result, the minimal number of training rounds in (10) is not sufficient for estimating the correlation matrix accurately. Therefore, the number of the training rounds is set to $K = 12$ in the following discussion.

Figure 6 shows the radiation patterns of antenna, (A), in xy-plane under the various interference conditions. Here, the letter, 'U' represents interference wave, and the INR is 30 dB, K is set to 12. Figures 6(a) and (b) indicate

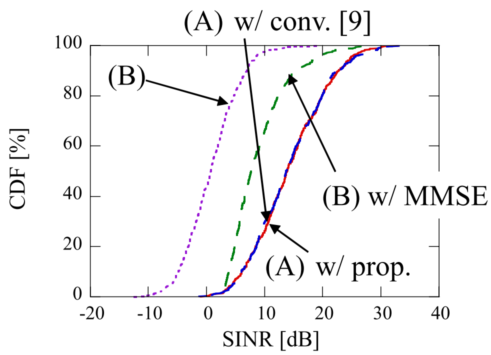


Fig. 7 CDF of SINR.

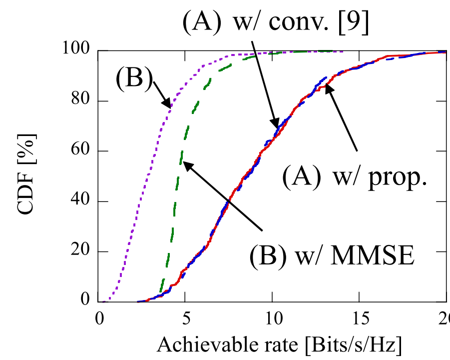


Fig. 8 CDF of achievable rate.

the patterns when a single interference wave arrives at the antenna from angles, 90°, and 45°, respectively. It can be clearly seen that nulls are formed at the angle of the interference wave for both Rx_1 and Rx_2. The reception power of the interference signal is lowered by at least 10 dB in these cases. Figure 6(c) shows the pattern when 10 interference waves arrive at the antenna from various angles, i.e. 30°, 40°, 50°, 160°, 220°, 225°, 230°, 240°, 250°, and 300°. Note that these 10 waves are same signals with different directions of arrivals. It can be seen that the radiation patterns for Rx_1 and Rx_2 do not have deep nulls. Nevertheless the observed interference power suppression by the proposed algorithm was about 10 dB in this case. This means the radiation pattern can be formed so as to suppress the synthesized interference signals.

Figure 7 shows the cumulative distribution function (CDF) of signal-to-interference-plus-noise ratio (SINR). The median signal-to-noise ratio (SNR) of the desired signal is determined to be 30 dB. The median value of the interference power is equal to that of the desired signal power. The path distribution is uniform in the horizontal plane. Other parameters are given identically to that in Fig. 5. Two types of the algorithms, i.e. the proposed and previous [9] ones, are tested and compared for the antenna (A). The number of the training rounds, K, is set to 12 for both two algorithms. For each training round, the termination set is given to be different to that for other training rounds. It can be seen two algorithms for (A) yield almost identical results. This means that the proposed method can offer the deterministic antenna control without any knowledge of the channel of the interference signal. Also, it is found that 50% values of the SINR in (B) with and without MMSE are lower by 13 dB and 6 dB, respectively, than that in (A).

Figure 8 shows the cumulative distribution function (CDF) of achievable rate. The achievable rate is approximately calculated by,

$$R = \log_2 \left| \boldsymbol{I} + \boldsymbol{H}\boldsymbol{H}^H \frac{\sigma_s^2}{(\sigma_n^2 + \sigma_i^2)M_t} \right| \tag{16}$$

where, σ_s^2, σ_i^2, and σ_n^2 represent the received powers of the desired and interference signals, and noise, respectively. M_t represents the number of the transmitting antennas. Though the capacity under the interference can be analyzed more in detail [16], the asymptotic capacity can be easily approximated if the interference can be recognized as Gaussian and the propagation environment is random [9]. For MMSE algorithm in antenna (B), the rate is also calculated by,

$$R_{MMSE} = \log_2 \left| \boldsymbol{I} + \boldsymbol{W}_{MMSE}^H \boldsymbol{H}\boldsymbol{H}^H \boldsymbol{W}_{MMSE} \frac{\sigma_s^2}{(\sigma'^2_n + \sigma'^2_i)M_t} \right|, \tag{17}$$

where, σ'^2_n and σ'^2_i are the received noise and interference powers after the MMSE filtering.

By comparing the results of (A) with proposed algorithm and [9], it is found that the two algorithms yield identical performances. Also, the MIMO adaptive antenna, (A), achieves three times higher rate than (B) without MMSE. Though (B) with MMSE can suppress the interference, significant improvement in the achievable rate is not observed. Since (B) has only two degrees of freedom, the rate is seriously degraded even when the interference is suppressed by the digital adaptive array.

Figure 9 shows 50% bitrate versus raw SIR. Where, Figs. 9(a) and (b) represent the results with Rayleigh environment (Rician factor = $-\infty$ dB) and Rician environment (Rician factor = 10 dB), respectively. The raw SIR represents the SIR for (B) without MMSE. The SNR is always set to 30 dB, and only the interference power is varied. It can be seen that the proposed algorithm can provide same performance to the previous work [9] over wide range of the SIR. In the Rayleigh environment, antenna (A) with the proposed algorithm yields higher bitrate than (B) does even when MMSE algorithm is applied to (B). On the other hand, the antenna, (B), with MMSE yields higher rate than antenna (A) especially when SIR is low in Rician environment. From these results, it can be seen that existence of the direct path affects the performance of the MIMO adaptive array. Also, it can be seen that the overall rate in Fig. 9(b) is lower than that in Fig. 9(a). This is because this channel is highly correlated and the rate enhancement effect of MIMO is not sufficiently obtained. That is, saving the degree of freedom is not effective in enhancing the rate for Rician environment. Nevertheless, the antenna, (A), outperforms antenna, (B), when SIR is higher than 0 dB. From these results, it can be seen that the proposed algorithm can offer satisfactory per-

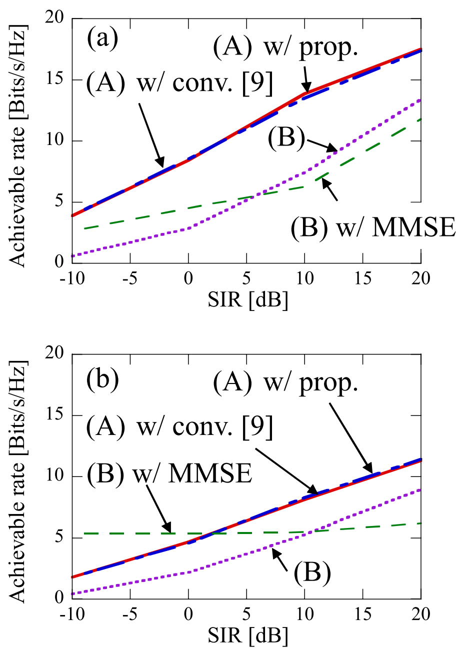


Fig. 9 50% rate versus SIR: (a) Rician factor = $-\infty$ dB, (b) Rician factor = 10 dB.

formance even without knowledge of the CSI.

4. Conclusion

In this paper, a fast non-iterative algorithm based on correlation matrix has been proposed and studied. The proposed algorithm does not need any knowledge of the CSI and avoids iterative training. The correlation matrix of the received interference signal at the antenna elements including the parasitic elements can be estimated by several training rounds with non-identical termination conditions. The performance of the proposed algorithm has been studied based on the simulation. It is shown the MIMO adaptive antenna with the proposed method can suppress the interference power by at least 10 dB in this simulation model, and enhances the bitrate greatly with the deterministic control scheme. Also, the MIMO adaptive antenna with the proposed algorithm offers satisfactory performance identical to that of the previous work [9]. From these results, it is found that the proposed algorithm is effective in reducing the time to control the MIMO adaptive antennas.

Acknowledgements

This research was partially supported by JSPS KAKENHI (22760261) and Murata Science Foundation.

References

[1] R.J. Dinger, "Reactively steered adaptive array using microstrip patch elements at 4 GHz," IEEE Trans. Antennas Propag., vol.34, no.3, pp.423–431, March 1986.
[2] T. Ohira and K. Gyoda, "Electronically steerable passive array radiator antennas for low-cost analog adaptive beamforming," 2000 IEEE International Conference on Phased Array Systems and Tech., pp.101–104, May 2000.
[3] M.D. Migliore, D. Pinchera, and F. Schettino, "Improving channel capacity using adaptive MIMO antennas," IEEE Trans. Antennas Propag., vol.54, no.11, pp.3481–3489, Nov. 2006.
[4] S. Smith and D.V. Thiele, Switched parasitic antennas for cellular communications, Artech House, 2001.
[5] T. Ohira, "Analog smart antennas: an overview," IEEE Int. Symp. Personal, Indoor Mobile Radio Commun. (PIMRC), vol.4, pp.1502–1506, Sept. 2002.
[6] E. Kataoka, M. Morishita, H. Yamada, M. Taromaru, and T. Ohira, "Interference cancellation performance of adaptive array of ESPAR antennas," Int. Symp. Antennas Propagat (ISAP), pp.760–763, Aug. 2007.
[7] Q. Han, B. Hanna, K. Inagaki, and T. Ohira, "Mutual impedance extraction and varactor calibration technique for ESPAR antenna characterization," IEEE Trans. Antennas Propag., vol.54, no.12, pp.3713–3720, Dec. 2006.
[8] N. Honma, K. Nishimori, R. Kudo, Y. Takatori, and K. Okada, "Using MIMO approach to get out of many iterations in analog adaptive beamforming," 2008 IEEE Antennas and Propagation Society International Symposium, no.304.3, July 2008.
[9] N. Honma, K. Nishimori, R. Kudo, Y. Takatori, and M. Mizoguchi, "Fast control method of parasitic antennas using non-iterative algorithm in multi-antenna system," IEEE Trans. Antennas Propag., vol.60, no.4, pp.2044–2051, April 2012.
[10] K. Nishimori, K. Cho, Y. Takatori, and T. Hori, "Adaptive array for reducing high-power CCI on asynchronous TDD systems," IEICE Trans. Commun., vol.E84-B, no.7, pp.1727–1734, July 2001.
[11] J.W. Wallace and M.A. Jensen, "Mutual coupling in MIMO wireless systems: A rigorous network theory analysis," IEEE Trans. Wireless Commun., vol.3, no.4, pp.1317–1325, July 2004.
[12] W.C. Jakes, Microwave Mobile Communications, IEEE, New York, 1993.
[13] B.K. Lau, J.B. Andersen, G. Kristensson, and A.F. Molisch, "Impact of matching network on bandwidth of compact antenna arrays," IEEE Trans. Antennas Propag., vol.54, no.11, pp.3225–3238, Nov. 2006.
[14] A.F. Molisch, "A generic model for MIMO wireless propagation channels in macro- and microcells," IEEE Trans. Signal Process., vol.52, no.1, pp.61–71, Jan. 2004.
[15] K. Nishimori, N. Tachikawa, Y. Takatori, R. Kudo, and K. Tsunekawa, "Frequency correlation characteristics due to antenna configurations in broadband MIMO transmission," IEICE Trans. Commun., vol.E88-B, no.6, pp.2438–2445, June 2005.
[16] R.S. Blum, "MIMO capacity with interference," IEEE J. Sel. Areas Commun., vol.21, no.5, pp.793–801, June 2003.

Naoki Honma received the B.E., M.E., and Ph.D. degrees in electrical engineering from Tohoku University, Sendai, Japan in 1996, 1998, and 2005, respectively. In 1998, he joined the NTT Radio Communication Systems Laboratories, Nippon Telegraph and Telephone Corporation (NTT), in Japan. He is now working for Iwate University. He received the Young Engineers Award from the IEICE of Japan in 2003, the APMC Best Paper Award in 2003, and the Best Paper Award of IEICE Communication Society in 2006, respectively. His current research interest is planar antennas for high-speed wireless communication systems. He is a member of IEEE.

Kentaro Nishimori received the B.E., M.E., and Dr.Eng. degrees in electrical and computer engineering from Nagoya Institute of Technology, Aichi, in 1994, 1996, and 2002, respectively. He joined NTT Wireless Systems Laboratories in 1996. He was a visiting researcher at the Center for TeleInFrastruktur, Aalborg University, Denmark in 2006. He received the Young Engineers Award from the Institute of Electronics, Information and Communication Engineers (IEICE) of Japan in 2001 and Young Engineer Award from IEEE AP-S Japan Chapter in 2001. His current research interests are multi-user MIMO systems and cognitive radio systems. He is a member of IEEE.

Takefumi Hiraguri received the M.E. and Ph.D. degree from the University of Tsukuba, Ibaraki, Japan, in 1999 and 2008, respectively. In 1999, he joined the NTT Access Network Service Systems Laboratories, Nippon Telegraph and Telephone Corporation (NTT) in Japan. He has been engaged in research and development of MAC protocol for the high speed and the high communication quality in wireless LANs systems. He is now associate professor in Nippon Institute of Technology (NIT). He is a member of IEEE.

Yoshitaka Tsunekawa received the B.E. degree from Iwate University, Morioka, Japan, in 1980 and the M.E. and the Doctor of Engineering degrees from Tohoku University, Sendai, Japan, in 1983 and 1993, respectively. Since 1983, he has been with the Faculty of Engineering, Iwate University, where he is now a professor at the Department of Electrical Engineering and Computer Science. His current research interests include digital signal processing and digital control. He is a member of the IEEE.