An Adaptive Antenna Array for Single Carrier Modulation System with Cyclic Prefix

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Abstract— **In a single carrier modulation system with cyclic prefix (SC-CP), equalization is possible for the case where the delayed signals lie within the guard interval (GI). The reception of all the delayed signals within the GI can increase the received signal power, and hence may improve the system performance. However, it also increases the probability that the channel transfer function has a zero (or zeros) on the FFT grid, which results in serious noise enhancement (NE). Therefore, some appropriate reception methods, which can receive multiple delayed signals within the GI while avoiding the NE, are required.**

In this paper, in order to achieve such a reception, an adaptive antenna array is applied to the SC-CP system, and the optimum weights control method is proposed. Moreover, to verify the performance of the proposed method, some weights control methods, originally proposed for the Orthogonal frequency division multiplexing (OFDM) adaptive array, are applied to the SC-CP adaptive antenna array, and their performances are studied and compared with the proposed system via simulations.

I. INTRODUCTION

A single carrier modulation system with cyclic prefix (SC-CP) [1]-[3] has been drawing much attention as a candidate for the 4G (4th generation) mobile communications systems. The SC-CP scheme employs a block transmission with the insertion of a cyclic prefix (CP) as a guard interval (GI), which is also employed in the Orthogonal frequency division multiplexing (OFDM) scheme [4]. The SC-CP scheme is closely related to the OFDM scheme, and also posses the robustness to the frequency selective fading. Moreover, the SC-CP can relax the requirements to the transmitter front-end, since the transmitted signal is a single carrier modulated signal, and hence it does not suffer from high peak-to-average power ratio (PAPR) problem.

When delayed signals exist inside of the GI, the insertion of CP converts the effect of channel from a linear convolution to a circular convolution. Therefore, the received signal can be equalized by multiplying complex weights after FFT operation on the signal. Since the existence of delayed signals beyond the GI deteriorates the performance of the SC-CP system, such delayed signals have to be removed so that the channel effect becomes the circular convolution. However, as for the delayed signals within the GI, there is some room for discussion. The reception of all the delayed signals within the GI can increase the received signal power and it may improve the performance of the SC-CP system, but it also increases the probability that the channel transfer function has a zero (or zeros) on the FFT grid, which results in serious noise enhancement (NE). Therefore, some appropriate reception methods, which can receive multiple delayed signals within the GI while avoiding the NE, are required.

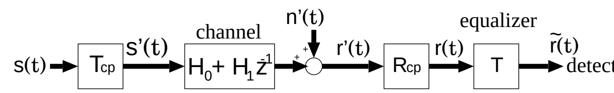

Fig. 1. Block description of CP manipulation

In this paper, in order to achieve such a reception, an adaptive antenna array is applied to the SC-CP system, and an optimum weights control method is proposed. Moreover, to verify the performance of the proposed method, some weights control methods, originally proposed for the OFDM system, are applied to the SC-CP adaptive antenna array, and their performances are also studied and compared with the proposed system via computer simulations.

II. SINGLE CARRIER MODULATION SYSTEM WITH CYCLIC PREFIX

The SC-CP scheme employs a block transmission with the insertion of a CP as the GI. The CP is a copy of the last part of the information bearing signals in the same block. The block transmission with the CP enables inter-block-interference (IBI)- and inter-symbol-interference (ISI)-free transmission by utilizing zero-forcing (ZF) based frequency domain equalizer (FDE) at the receiver.

Fig.1 shows the block description of the SC-CP system [5]. In the transmitter, the CP of length K is attached to the nth information block $\mathbf{s}(n)$ of length M. Therefore, the length of one transmission block $\mathbf{s}'(n)$ becomes $M+K$;

$$\mathbf{s}'(n) = \mathbf{T}_{cp}\mathbf{s}(n), \tag{1}$$

where $\mathbf{T}_{cp}$ denotes the $(M+K)\times M$ CP insertion matrix defined as

$$\mathbf{T}_{cp} = \begin{pmatrix} \mathbf{0}_{M\times K} & \mathbf{I}_{M\times M} \end{pmatrix}. \tag{2}$$

$\mathbf{0}_{M\times K}$ is the zero matrix of size $M\times K$, $\mathbf{I}_{M\times M}$ is an identity matrix of size $M\times M$. If we assume that there exists no delayed signal outside the GI, namely, the channel order L

satisfies $L < K$, the received signal block $\mathbf{r}'(n)$ is given by

$$\mathbf{r}'(n) = \mathbf{H}_0\mathbf{s}'(n) + \mathbf{H}_1\mathbf{s}'(n-1) + \mathbf{n}'(n), \tag{3}$$

where $\mathbf{H}_0$ and $\mathbf{H}_1$ denote $(M+K)\times(M+K)$ channel matrices defined as

$$\mathbf{H}_0 = \begin{pmatrix} h_1 & 0 & 0 & \dots & 0 \\ \vdots & h_1 & 0 & \dots & 0 \\ h_L & \dots & \ddots & \dots & \vdots \\ \vdots & \ddots & \dots & \ddots & 0 \\ 0 & \dots & h_L & \dots & h_1 \end{pmatrix}, \tag{4}$$

$$\mathbf{H}_1 = \begin{pmatrix} 0 & \dots & h_L & \dots & h_2 \\ \vdots & \ddots & 0 & \ddots & \vdots \\ 0 & \dots & \ddots & \dots & h_L \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & \dots & 0 & \dots & 0 \end{pmatrix}. \tag{5}$$

After discarding the CP, the received signal block $\mathbf{r}(n)$ can be written as

$$\mathbf{r}(n) = \mathbf{R}_{cp}\mathbf{r}'(n), \tag{6}$$

where $\mathbf{R}_{cp}$ denotes the $M\times(M+K)$ CP discarding matrix defined by

$$\mathbf{R}_{cp} = \begin{pmatrix} \mathbf{I}'_{K\times M} \\ \mathbf{I}_{M\times M} \end{pmatrix}, \quad \mathbf{I}'_{K\times M} = \begin{pmatrix} \mathbf{0}_{K\times(M-K)} & \mathbf{I}_{K\times K} \end{pmatrix}. \tag{7}$$

When $L < K$, $\mathbf{R}_{cp}\mathbf{H}_1 = \mathbf{0}_{M\times(M+K)}$, therefore, the nth received signal block $\mathbf{r}(n)$ can be written as

$$\mathbf{r}(n) = \mathbf{C}\mathbf{s}(n) + \mathbf{n}(n). \tag{8}$$

Here, the $M\times M$ circulant matrix $\mathbf{C}$ and the $M\times 1$ noise vector $\mathbf{n}(n)$ are defined as

$$\mathbf{C} = \mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}, \quad \mathbf{n}(n) = \mathbf{R}_{cp}\mathbf{n}'(n). \tag{9}$$

From (8), we can see that the IBI due to the $(n-1)$th information block is eliminated from the nth received signal block. Moreover, the effect of the channel is converted from the linear convolution to the circular convolution.

Then we define the channel impulse response vector $\mathbf{h} = [h_1, \dots, h_L, 0, \dots, 0]^T$. Here, $(\cdot)^T$ denotes the transpose. The matrix $\mathbf{C}$ is a circulant matrix whose elements are determined by the channel impulse response. Since a circulant matrix can be diagonalized by pre- and post- multiplication of the DFT matrix $\mathbf{D}$, we have

$$\mathbf{C} = \mathbf{D}^H\mathbf{\Lambda}\mathbf{D}, \quad \mathbf{\Lambda} = \text{diag}\left[\mathbf{D}\mathbf{h}\right], \tag{10}$$

where $\text{diag}\left[\mathbf{D}\mathbf{h}\right]$ denotes a diagonal matrix whose diagonal elements are $\mathbf{D}\mathbf{h}$, and $(\cdot)^H$ denotes the Hermitian transpose.

The ZF based channel equalization matrix $\mathbf{T}$ is given by

$$\mathbf{T} = \mathbf{C}^{-1} = \mathbf{D}^H\mathbf{\Lambda}^{-1}\mathbf{D}. \tag{11}$$

The received signal block after the equalization $\tilde{\mathbf{r}}(n)$ becomes

$$\tilde{\mathbf{r}}(n) = \mathbf{T}\mathbf{r}(n) = \mathbf{s}(n) + \mathbf{D}^H\mathbf{\Lambda}^{-1}\mathbf{D}\mathbf{n}(n). \tag{12}$$

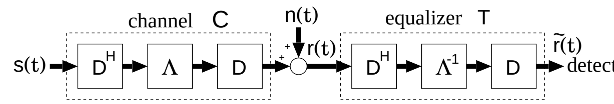

Fig. 2. Simplified block description of the SC-CP system

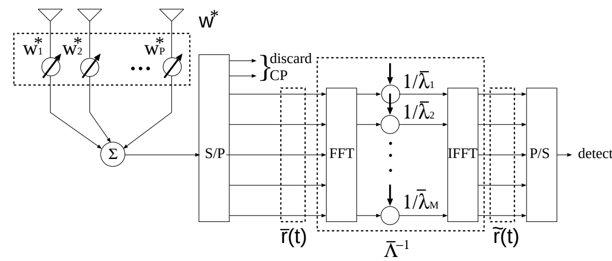

Fig. 3. Configuration of the proposed SC-CP adaptive array

Using the circulant matrix $\mathbf{C}$, the block description of the SC-CP system can be simplified as shown in Fig.2.

In the SC-CP system, the delayed signals within the GI can be equalized with the ZF based equalization. Therefore, the active reception of the delayed signals leads to the increase of the received signal power, and hence the improvement of the system performance. However, the active reception also increases the probability that the channel transfer function has a zero (or zeros) on the FFT grid, which results in serious NE.

III. Configuration of the Proposed System

Fig.3 shows the configuration of the proposed system, where an adaptive antenna array with P elements is employed. Let $\mathbf{h}_p = [h_{1,p}, h_{2,p}, \dots, h_{L,p}, 0, \dots, 0]^T$ denotes the channel impulse response vector whose element $h_{l,p}, (l = 0, \dots, L)$ is the channel impulse response at the pth antenna element, $\mathbf{n}_p(n)$ denotes $M\times 1$ noise vector, and w_p^* the weights of the adaptive antenna array. Here, $(\cdot)^*$ denotes the complex conjugate. We assume that the elements of $\mathbf{n}_p(n)$ are zero mean white noise with the variance of σ^2.

The channel impulse response vector at the array output $\bar{\mathbf{h}}$ and the noise vector $\bar{\mathbf{n}}(n)$ can be written as

$$\bar{\mathbf{h}} = \mathbf{H}\mathbf{w}^*, \quad \mathbf{H} = [\mathbf{h}_1, \dots, \mathbf{h}_P], \tag{13}$$

$$\bar{\mathbf{n}}(n) = \mathbf{N}(n)\mathbf{w}^*, \quad \mathbf{N}(n) = [\mathbf{n}_1(n), \dots, \mathbf{n}_P(n)], \tag{14}$$

where $\mathbf{w} = [w_1, w_2, \dots, w_P]^T$ denotes the weights vector of the adaptive antenna array.

After discarding the CP, the array output signal block $\bar{\mathbf{r}}(n)$ can be written as

$$\bar{\mathbf{r}}(n) = \bar{\mathbf{C}}\mathbf{s}(n) + \bar{\mathbf{n}}(n). \tag{15}$$

Here, $\bar{\mathbf{C}}$ is a circulant matrix formed by $\bar{\mathbf{h}}$, therefore, it can be diagonalized by the DFT matrix.

$$\bar{\mathbf{C}} = \mathbf{D}^H\bar{\mathbf{\Lambda}}\mathbf{D}, \quad \bar{\mathbf{\Lambda}} = \text{diag}\left[\mathbf{D}\bar{\mathbf{h}}\right] \tag{16}$$

The ZF based equalization matrix $\bar{\mathbf{T}}$ is given by

$$\bar{\mathbf{T}} = \bar{\mathbf{C}}^{-1} = \mathbf{D}^H\bar{\mathbf{\Lambda}}^{-1}\mathbf{D} \tag{17}$$

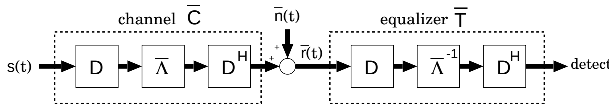

Fig. 4. Block description of the SC-CP adaptive antenna array

Fig.4 shows the simplified block description of the SC-CP system with the adaptive antenna array.

IV. Weight Control Methods

In this section, a weights control method to find the appropriate weights under the condition that delayed signals exist only within the GI is proposed. Moreover, in order to compare the performance of the proposed method with those of other methods, some weights control methods, originally proposed for the OFDM adaptive array, are applied to the SC-CP adaptive antenna array.

A. Maximization of SNR at equalizer output

Here, we propose a weights control method maximizing the average signal to noise power ratio (SNR) at the equalizer output. At the equalizer output, the noise vector $\tilde{\mathbf{n}}(n)$ can be written as

$$\tilde{\mathbf{n}}(n) = \mathbf{T}\bar{\mathbf{n}}(n) = \mathbf{D}^H \bar{\mathbf{\Lambda}}^{-1} \mathbf{D}\mathbf{N}(n)\mathbf{w}^*. \tag{18}$$

The noise power at the equalizer output P_n becomes

$$P_n = \frac{1}{M}\mathrm{Tr}\left[E\left[\tilde{\mathbf{n}}(n)\tilde{\mathbf{n}}^H(n)\right]\right] \tag{19}$$

$$= \frac{\sigma^2}{M}\sum_{m=1}^{M}\frac{\mathbf{w}^H\mathbf{w}}{|\bar{\lambda}_m|^2} = \frac{\sigma^2}{M}\sum_{m=1}^{M}\frac{\mathbf{w}^H\mathbf{w}}{\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w}}, \tag{20}$$

where $\bar{\lambda}_m$ denotes the mth diagonal element of $\bar{\Lambda}$, and $\tilde{\mathbf{R}}_m$ and $\tilde{\mathbf{h}}_m$ are defined as

$$\tilde{\mathbf{R}}_m = \tilde{\mathbf{h}}_m\tilde{\mathbf{h}}_m^H, \quad \left[\tilde{\mathbf{h}}_1, \ldots, \tilde{\mathbf{h}}_M\right] = (\mathbf{D}\mathbf{H})^T. \tag{21}$$

Since the ZF based equalizer is employed, the average SNR at the equalizer output SNR_e becomes $SNR_e = \frac{1}{P_n}$. Therefore, the weights that maximize the SNR at the equalizer output is equivalent to the solution of minimization of P_n subject to the constraint of $\mathbf{w}^H\mathbf{w} = 1$. Applying Lagrange multipliers method to this constrained minimization, the optimum weights satisfy the following equation:

$$\sum_{m=0}^{M-1}\frac{\tilde{\mathbf{R}}_m}{(\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w})^2}\mathbf{w} - \sum_{m=0}^{M-1}\frac{1}{\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w}}\mathbf{w} = 0 \tag{22}$$

Since the equation is nonlinear, it is difficult to find the analytical solutions. In this paper, the steepest decent method [6] is applied to find the weights that minimize P_n. The adaptation of the weights vector is given by

$$\mathbf{w}(t+1) = \mathbf{w}(n) - \mu\frac{\partial P_n}{\partial \mathbf{w}^*},$$

$$\frac{\partial P_n}{\partial \mathbf{w}^*} = \frac{\sigma^2}{M}\sum_{m=1}^{M}\frac{\mathbf{w}}{\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w}} - \mathbf{w}^H\mathbf{w}\frac{\tilde{\mathbf{R}}_m\mathbf{w}}{(\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w})^2}, \tag{23}$$

where μ denotes the step-size parameter.

B. Maximization of SNR at array output

A weights control method which maximizes the SNR at the array output is already proposed for the OFDM system [7] when the delayed signals exist only within the GI. Here, the method is applied to the SC-CP adaptive antenna array.

The signal power at the array output P_s' is given by

$$P_s' = \frac{1}{M}\mathrm{Tr}\left[E\left[\mathbf{s}(n)^H\bar{\mathbf{C}}^H\bar{\mathbf{C}}\mathbf{s}(n)\right]\right] = \frac{1}{M}\mathrm{Tr}\left[\mathbf{D}^H\bar{\mathbf{\Lambda}}^H\bar{\mathbf{\Lambda}}\mathbf{D}\right]$$

$$= \frac{1}{M}\sum_{m=1}^{M}|\bar{\lambda}_m|^2 = \frac{1}{M}\sum_{m=1}^{M}\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w}. \tag{24}$$

The noise power at the array output P_n' is given by

$$P_n' = \frac{1}{M}Tr\left[E\left[\bar{\mathbf{n}}(n)^H\bar{\mathbf{n}}(n)\right]\right] = \frac{\sigma^2}{M}\mathbf{w}^H\mathbf{w}. \tag{25}$$

Therefore, the average SNR at the array output SNR_a can be written as

$$SNR_a = \frac{P_s'}{P_n'} = \sum_{m=1}^{M}\frac{\mathbf{w}^H\mathbf{R}_m\mathbf{w}}{\sigma^2\mathbf{w}^H\mathbf{w}}. \tag{26}$$

Applying Lagrange multipliers method with the constraint of $\mathbf{w}^H\mathbf{w} = 1$, the optimum weights in term of maximization of the SNR at the array output, satisfy the following equation:

$$\left(\sum_{m=1}^{M}\tilde{\mathbf{R}}_m\right)\mathbf{w} = \left(\sum_{m=1}^{M}\mathbf{w}^H\tilde{\mathbf{R}}_m\mathbf{w}\right)\mathbf{w}. \tag{27}$$

Therefore, the eigenvector $\mathbf{w}$ corresponding to the maximum eigenvalue of $\sum_{m=1}^{M}\tilde{\mathbf{R}}_m$ maximizes the average SNR at the array output.

C. Utilization of filtered reference signals

A weights control method for receiving multiple delayed signals is proposed for the adaptive antenna array [8] and the method is applied to the OFDM system [9]. Originally, the method aims at eliminating the cochannel-interference (CCI) from received signals that consists of desired signals, delayed signals and other user's signals. The method assumes that the ISI caused by the received delayed signals can be equalized at the following equalizer.

The method employs a filtered reference signal $d(t)$ generated by a transversal filter consists of K tap weights c_k. Here, the jth tap is fixed to be 1. In this paper, the length of the transversal filter is set to be the same length as the GI.

The kth filtered reference signal $d(k)$ is generated by passing known pilot signal $s(k)$ through the transversal filter as

$$d(k) = \sum_{m=1}^{M}\mathbf{c}_m^* s(k+1-m) = \mathbf{c}^H\mathbf{s}(k), \tag{28}$$

where $\mathbf{s}(k)$ and $\mathbf{c}$ denote $M \times 1$ vectors defined as $\mathbf{s}(k) = [s(k), \ldots, s(k-M+1)]^T$ and $\mathbf{c} = [c_1, \ldots, c_K, 0\ldots, 0]^T$, respectively. The array output signal $y(k)$ can be written as

$$y(k) = \sum_{m=1}^{M}\bar{h}_m s(k+1-m) = \mathbf{w}^H\mathbf{H}^T\mathbf{s}(k). \tag{29}$$

TABLE I

SYSTEM PARAMETERS

Mod/Demod Scheme	QPSK
Symbols/block	M=64
Guard Interval	K=16
Antenna Array	8 sensors circular array
Sensor Spacing	Half of the carrier wave length

The error signal $e(k)$ is given by

$$e(k) = d(k) - y(k) = (\mathbf{c}^H - \mathbf{w}^H\mathbf{H}^T)\mathbf{s}(k). \qquad (30)$$

In this paper, we minimize the mean-squared error using LMS algorithm [6]. Using newly defined tap weight vector $\bar{\mathbf{c}} = [c_1, \ldots, c_{j-1}, c_{j+1}, \ldots, c_K]$, the weights update equations are given by

$$\mathbf{w}(k+1) = \mathbf{w}(k) + \mu_w \mathbf{H}^T \mathbf{s}(k) e(k)^*, \qquad (31)$$

$$\bar{\mathbf{c}}(k+1) = \bar{\mathbf{c}}(k) - \mu_c \mathbf{s}(k) e(k)^*, \qquad (32)$$

where $\mu_w and \mu_c$ denote the step size parameters.

V. COMPUTER SIMULATION

A. System parameters

The computer simulations are carried out for the following four weights control methods:

- Maximization of SNR at equalizer output (MSE)
- Maximization of SNR at array output (MSA)
- Utilization of filtered reference signals (FRS)
- Reception of only the primary wave (conventional)

System parameters used in the computer simulations are shown in Table I.

B. Shape of cost function

Fig.5 shows an example of the cost function of the MSE P'_n when the number of arrival waves is 2, and the number of the array weights is 2. In the figure, * shows the solution of (22) and the solid line shows the trajectory of adaptations with the steepest decent algorithm. Since the form of cost function does not implicate the existence of local solutions, the application of the steepest decent method may give approximate solutions of the optimum weights.

C. Antenna beam patterns

Fig.6 shows examples of antenna beam patterns of the weights control methods, where the number of arrival waves is 3 and the angles of arrival are fixed to be 0, 30, -60 [deg]. The conventional method gives gain to the primary wave and forms nulls toward the delayed waves, however, the other methods do not form nulls toward the delayed waves while giving gain to the primary wave. This means that that the MSE, MSA and FRS receive not only the primary wave, but also the delayed waves.

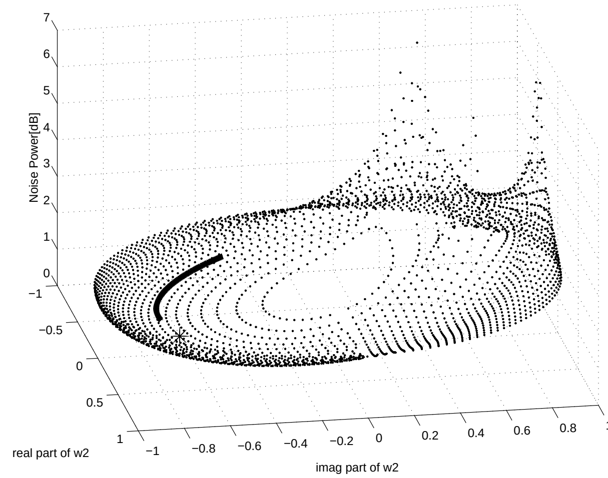

Fig. 5. Example of the shape of the cost function

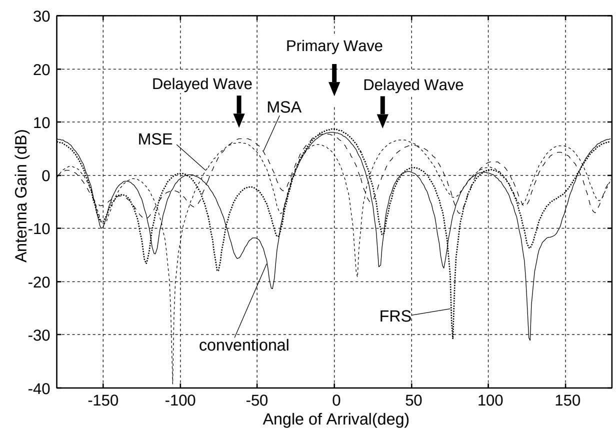

Fig. 6. Antenna beam patterns

D. SNR at the array output and the equalizer output

Fig.7 shows the average SNR at the array output vs. the average SNR at the array inputs for each method, where the number of incoming waves is 6. The arrival time of each path is randomly determined within the GI and the each path is affected by Rayleigh fading. The direction of arrival (DoA) of each path is also randomly determined. The channel impulse response is estimated at the receiver by an 8-stage maximum length shift register (M-) sequence and the estimated channel response is used for the equalization. The results show that the MSA maximizes the SNR at the array output, and the MSE and FRS methods also increase the SNR at the array output as compared with the conventional method.

Fig.8 shows the average SNR at the equalizer output vs. the average SNR at the array inputs for each method, where the number of incoming waves is 6. The results show that the MSE maximizes the average SNR at the equalizer output, however, the MSA and FRS methods have only small gain or decrease the SNR at the equalizer output though they increase the SNR at the array output.

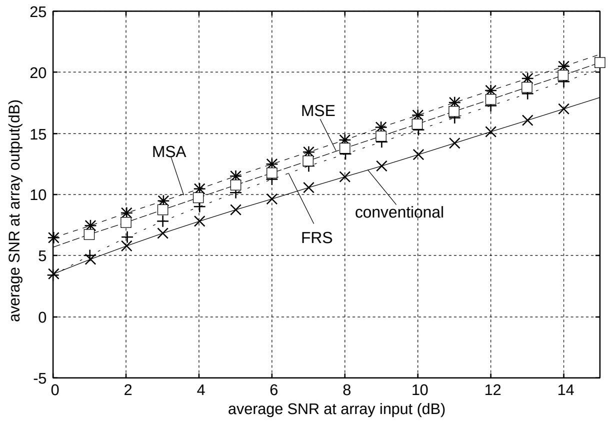

Fig. 7. Average SNR at the array output

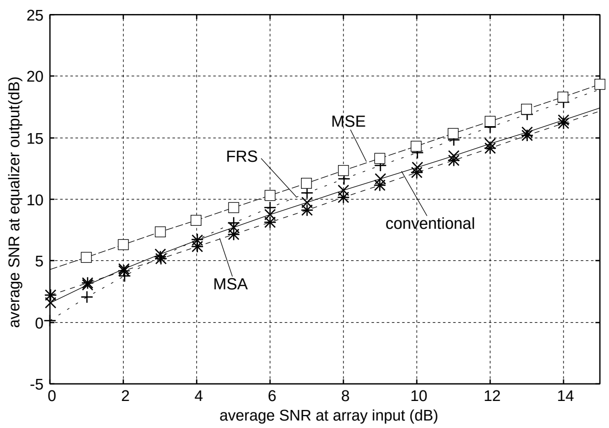

Fig. 8. Average SNR at the equalizer output

E. Bit error rate performance

Fig.9 shows the bit error rate (BER) vs. the ratio of the average energy per symbol to the noise power density E_s/N_0, where the number of incoming waves is 6. The MSE shows the best performance among the four methods. This is because, unlike the OFDM scheme, the BER of the SC-CP system is governed by the average SNR at the equalizer output [10]. Although the MSA maximize the average SNR at the array output, it shows the worst performance. The FRS achieves the reception of the delayed signals, however, it does not increase the SNR at the equalizer output so much, and results in almost the same performance as the conventional method.

From all the results, it can be concluded that the reception of delayed signals with consideration of the NE can improve the performance of the SC-CP adaptive antenna array, but the reception without the consideration of the NE does not necessarily guarantee the improvement of the performance.

VI. Conclusion

In order to improve the performance of the SC-CP system, an adaptive antenna array is employed for the SC-CP system and the optimum weights control method, maximizing SNR at

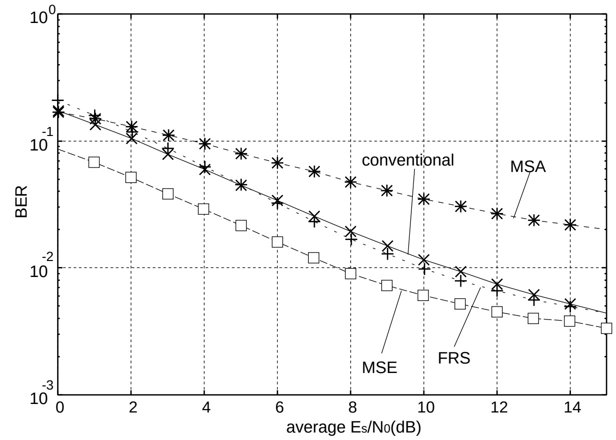

Fig. 9. Bit error rate performance

the equalizer outputs, is proposed. Additionally, some weights control methods, originally proposed for the OFDM adaptive array are also applied to the SC-CP adaptive antenna array. The performances of all the methods are studied in the computer simulations, and the results show that the proposed weights control methods can achieve the reception of the delayed signals while avoiding the NE and improve the performances of the SC-CP system.

In this paper, the discussions and simulations were carried out for the uncoded SC-CP adaptive antenna array under the condition that all the delayed signals are within the GI. Future investigation may include the study of the weight control method for the channel where the delayed signals exist beyond the GI, and of the introduction of error-correcting code to the proposed system.

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