Optimum Beamforming for Pre-FFT OFDM Adaptive Antenna Array

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***Abstract*—It is well known that orthogonal frequency-division multiplexing (OFDM) is robust to frequency-selective fading in wireless channels due to the exploitation of a guard interval that is inserted at the beginning of each OFDM symbol. However, once delayed signals beyond the guard interval are introduced in a channel with a large delay spread, intersymbol interference causes a severe degradation in the transmission performance. In this paper, we propose a novel pre-fast Fourier transform (FFT) OFDM adaptive antenna array, which requires only one FFT processor at a receiver, for suppressing such delayed signals. We derive the optimum weight set for beamformers based on the maximum signal-to-noise-and-interference power ratio (Max-SNIR) and the minimum mean square error (mmse) criteria, respectively. In addition, we propose a novel mmse-criterion-based *commutative* optimization scheme, which is more robust to the estimation error of the channel state information. Furthermore, we show the equivalence between the Max-SNIR-criterion-based scheme and the proposed commutative optimization scheme. Computer simulation results show its good performance even in channels where directions of arrival of arriving waves are randomly determined.**

***Index Terms*—Commutative scheme, delayed signal, pre-fast Fourier transform (pre-FFT) orthogonal frequency-division multiplexing (OFDM).**

I. Introduction

It IS well known that orthogonal frequency division multiplexing (OFDM) is an efficient technique for high-speed digital transmission over severe multipath fading channels [1], [2]. OFDM has been adopted in the physical layer of wireless local area network (LAN) systems [3] and has been recently considered to be a promising technique for next-generation mobile communications [4]. OFDM is robust to frequency-selective fading in wireless channels due to the exploitation of a guard interval, which is inserted at the beginning of each OFDM symbol [1], [2]. However, once delayed signals beyond the guard interval are introduced in a channel with a large delay spread, intersymbol interference (ISI) causes a severe degradation in the transmission performance. To cancel the ISI, time-domain signal-processing techniques have been proposed for wireless OFDM receivers, such as a decision feed-back equalizer (DFE) [5] and a maximum likelihood sequence estimator (MLSE) with a smoothed fast Fourier transform (FFT) window [6]. They have high interference cancellation capabilities; however, they have complicated structures because they need to estimate the channel impulse response, including the delayed signals beyond the guard interval.

On the other hand, to maintain high-speed reliable radio transmission, a multiple antenna array has been considered as an effective tool not only for gain enhancement and increased spectral efficiency [7], but also for interference suppression [8]–[12]. In terms of cancellation capability for cochannel interference including ISI, multiple antenna arrays are advantageous over time-domain signal-processing techniques, because they can cope with any interfering signals that are uncorrelated with desired signals by means of "nulling out." Unfortunately, the signal-processing techniques mentioned above can cancel only ISI by means of "equalization."

The structure of OFDM antenna array can be classified into two types, namely, *pre-* and *post*-FFT antenna array types. In [13], a post-FFT OFDM adaptive antenna array for cochannel interference suppression has been proposed. Although the proposed post-FFT subcarrier-by-subcarrier combining scheme is optimum in terms of maximizing signal-to-noise-and-interference power ratio (SNIR), it requires an increased number of FFT processors, large computations that increase with the number of antennas and subcarriers, and a quite long training signal. Furthermore, the proposed system requires that the received signals on different antenna elements are statistically independent. To establish such a diversity system, the antenna-element spacing has to be in the order of several carrier wavelengths. Therefore, depending on the number of antenna elements, the frequency band used, and the angular spread that is the angle over which the signal arrives at the receive antennas, the antenna size might become quite large. On the other hand, a pre-FFT combining diversity scheme proposed in [14], which requires only one FFT processor, can drastically reduce the computational complexity by tolerating a slight performance degradation, while achieving a space diversity gain. However, in [14], no interfering signal has been taken into account.

In this paper, we first propose a novel pre-FFT OFDM adaptive antenna array for suppressing *delayed signals beyond the guard interval* (or, in short, *delayed signals*) [11], [12]. We derive the optimum weight set for the beamformers based on the Maximum (Max-) SNIR and the minimum mean square error (mmse) criteria, respectively. Furthermore, we propose a novel mmse-criterion-based *commutative* optimization scheme that is more robust to the estimation error of the channel state information (CSI) than the Max-SNIR-based scheme. Computer simulation results show its good performance even

Manuscript received January 14, 2003; revised December 23, 2003 and March 4, 2004. This paper was presented in part at the IEEE Wireless Communications and Networking Conference (WCNC) 2003.

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Digital Object Identifier 10.1109/TVT.2004.830939

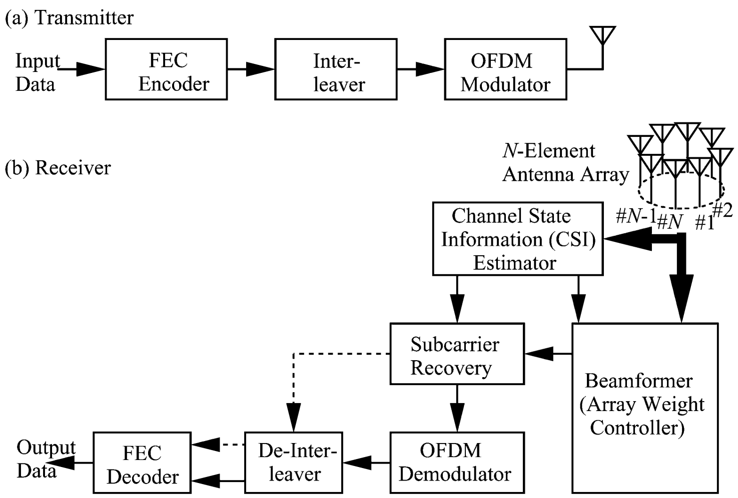

Fig. 1. Block diagrams of a transmitter and receiver.

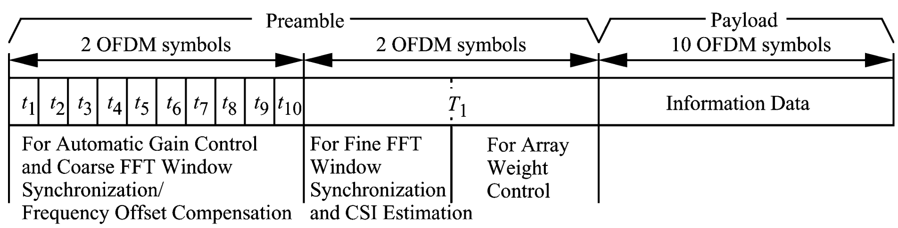

Fig. 2. OFDM signal burst format.

in channels where directions of arrival (DoAs) of arriving waves are randomly determined.

The rest of this paper is organized as follows. Section II describes the system model of the pre-FFT OFDM adaptive antenna array. Section III presents two optimum combining schemes based on the Max-SNIR and the mmse criteria, respectively. Section IV proposes the novel commutative maximization scheme. Section V summarizes the characteristics of the proposed schemes, Section VI demonstrates the computer simulation results and discusses the obtained results in detail. Finally, Section VII concludes this paper.

II. Pre-FFT OFDM System Description

Fig. 1(a) and (b) shows the block diagrams of a transmitter and a receiver, respectively. The data bits are applied to a forward error correction (FEC) encoder, followed by a bit-level interleaver and an OFDM modulator. The transmitted signal is disturbed by a frequency-selective fading and an additive white Gaussian noise (AWGN). At the receiver, CSI (namely, instantaneous channel impulse response) is first estimated with the aid of a training signal and then a set of array weights is determined. The received signal, after combining by an appropriate array weight set, is applied to an OFDM demodulator, followed by a deinterleaver and an FEC decoder.

Fig. 2 shows an OFDM signal burst format used in this paper, which is composed of a preamble and a payload [15]. One OFDM symbol corresponds to 4 μs and the payload is ten OFDM (data) symbols long. The preamble is composed of two parts, each of which is two OFDM symbols long. The first half is composed of ten repetitions of short training symbols, which

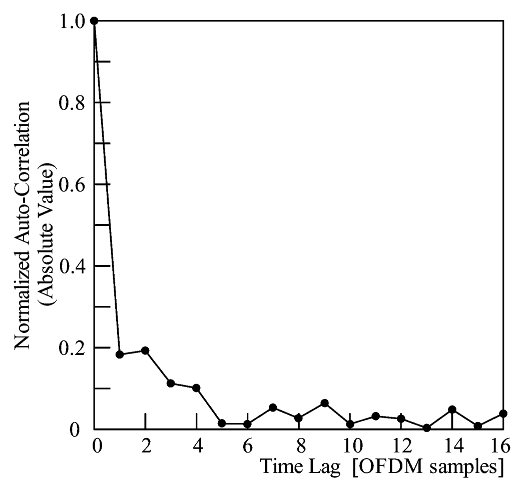

Fig. 3. Autocorrelation property of the first half of the long training symbol.

is usually used for automatic gain control (AGC) and coarse FFT timing synchronization/frequency offset compensation, while the second half is composed of a long training symbol for fine FFT timing synchronization/frequency offset compensation and subcarrier recovery. Our proposed array system uses the first half of the long training symbol for fine FFT timing synchronization and CSI estimation and the second half for array weight control. Here, subcarrier recovery to carry out coherent demodulation is done in the time domain by utilizing the estimated CSI [16]. Fig. 3 shows the autocorrelation property of the first half of the long training symbol.

Fig. 4 shows the CSIs for desired signals and delayed signals. Defining the lengths of the guard interval and the useful symbol interval as L_g and L_o in OFDM samples, respectively, assume that the maximum delay time of the delayed signals is within L_o from the end of the guard interval. In this case, the CSI matrices of the desired signals $(\mathbf{H}_x)$ and of the delayed signals $(\mathbf{H}_y)$ can be defined as

$$\mathbf{H}_x = [\mathbf{h}_{x1}, \ldots, \mathbf{h}_{xn}, \ldots, \mathbf{h}_{xN}]^T \tag{1}$$
$$\mathbf{H}_y = [\mathbf{h}_{y1}, \ldots, \mathbf{h}_{yn}, \ldots, \mathbf{h}_{yN}]^T \tag{2}$$

respectively, where N denotes the number of antenna array elements and superscript T stands for transposition. The $(L_g \times 1)$ vector $\mathbf{h}_{xn}$ and $(L_o \times 1)$ vector $\mathbf{h}_{yn}$ are the CSI vectors of the desired signals and delayed signals at the nth antenna element, respectively. The dimensions of the $\mathbf{H}_x$ and $\mathbf{H}_y$ are $(N \times L_g)$ and $(N \times L_o)$, respectively.

For data processing at the lth time, we define the $(L_g \times 1)$ desired signal vector and the $(L_o \times 1)$ delayed signal vector as

$$\mathbf{S}_x = [s_l, s_{l-1}, \ldots, s_{l-(L_g-1)}]^T \tag{3}$$
$$\mathbf{S}_y = [s_{l-L_g}, s_{l-(L_g+1)}, \ldots, s_{l-(L_g+L_o-1)}]^T \tag{4}$$

respectively, where s_{l-i} denotes the transmitted signal at the $(l-i)$th time. In this paper, we assume a slowly fading channel; in other words, we consider a case where the channel fading rate is so slow that the CSI does not significantly change over one signal burst (56 μs), i.e., quasi-static fading.

The $(N \times 1)$ received signal vector can be written as

$$\begin{aligned}\mathbf{r} &= [\mathbf{r}(1), \mathbf{r}(2), \ldots, \mathbf{r}(N)]^T \\ &= \mathbf{H}_x\mathbf{S}_x + \mathbf{H}_y\mathbf{S}_y + \mathbf{n}\end{aligned} \tag{5}$$

where the general notation $\mathbf{x}(i)$ denotes the ith element of vector $\mathbf{x}$ and $\mathbf{n}$ denotes the $(N \times 1)$ complex Gaussian noise vector. Fig. 5 shows the structure of the pre-FFT OFDM adaptive antenna array. The received signal combined by an $(N \times 1)$ array weight vector $(\mathbf{w}_a)$ is written as

$$\begin{aligned}z &= \mathbf{w}_a^H\mathbf{r} \\ &= \mathbf{w}_a^H\mathbf{H}_x\mathbf{S}_x + \mathbf{w}_a^H\mathbf{H}_y\mathbf{S}_y + \mathbf{w}_a^H\mathbf{n} \\ &= z_x + z_y + z_n\end{aligned} \tag{6}$$

where z_x, z_y, and z_n are the desired signal and delayed signal and noise components, respectively, and superscript H stands for Hermitian transposition.

III. Optimum Array Weight Set

In this section, we review the formulations of array weight set for the optimum beamformers based on the Max-SNIR and the mmse criteria.

A. Max-SNIR Criterion

As shown in Appendix A, the autocorrelation property of OFDM signal is not perfectly white when virtual subcarriers (VSs) exist. Here, the VS is a subcarrier that is not used for data transmission; in other words, zero information is transmitted instead of useful data information. The purpose of VS is to confine the sidelobe spectrum of the OFDM signal and to avoid a

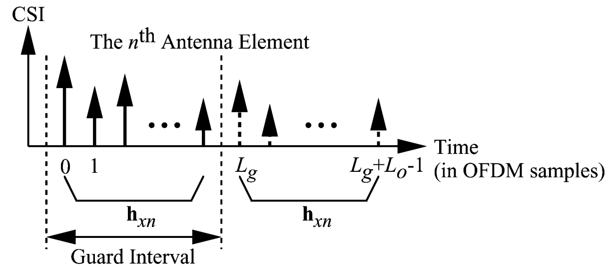

Fig. 4. CSIs of desired signals (h_{xn}) and of delayed signals (h_{xn}) at the nth antenna element.

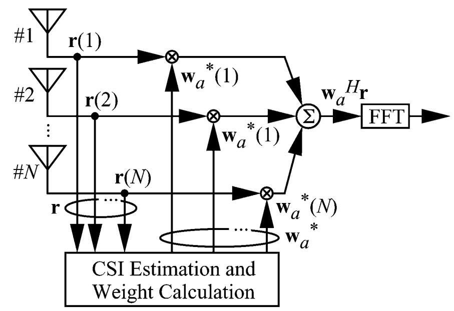

Fig. 5. Structure of the pre-FFT OFDM adaptive antenna array.

direct current (dc) offset in the demodulation process. The utilization of VS appears in most of wireless LAN standards, such as IEEE 802.11a [15], Mobile Multimedia Access Communication (MMAC) [17], and High Performance Local Area Network Type 2 (HiperLAN/2) systems [18]. Therefore, the discussions of the Max-SNIR criterion are divided into two cases, namely, with and without-VS cases.

1) Without Virtual Subcarrier: When all the subcarriers are fully occupied, the OFDM signal satisfies the following important property:

$$E\left[\mathbf{S}_x\mathbf{S}_x^H\right] = \sigma_s^2\mathbf{I}_{L_g \times L_g} \tag{7}$$
$$E\left[\mathbf{S}_y\mathbf{S}_y^H\right] = \sigma_s^2\mathbf{I}_{L_o \times L_o} \tag{8}$$
$$E\left[\mathbf{S}_x\mathbf{S}_y^H\right] = E\left[\mathbf{S}_y\mathbf{S}_x^H\right] = \mathbf{0} \tag{9}$$

where $E(\cdot)$ stands for ensemble average and $\mathbf{I}_{k\times k}$, $\mathbf{0}$, and σ_s^2 denote the $(k \times k)$ identity matrix, the zero matrix, and the average transmitted signal power, respectively. Furthermore, with the assumption that the AWGN is uncorrelated with the OFDM signal; in other words, the background noise field is *spatially incoherent*, the $(N \times N)$ received signal correlation matrix can be written as

$$\begin{aligned}\mathbf{R}_r &= E[\mathbf{r}\mathbf{r}^H] \\ &= \mathbf{H}_x E\left[\mathbf{S}_x\mathbf{S}_x^H\right]\mathbf{H}_x^H + \mathbf{H}_y E\left[\mathbf{S}_y\mathbf{S}_y^H\right]\mathbf{H}_y^H + E\left[\mathbf{n}\mathbf{n}^H\right] \\ &= \sigma_s^2\mathbf{R}_x + \sigma_s^2\mathbf{R}_y + \sigma_n^2\mathbf{I}\end{aligned} \tag{10}$$

where σ_n^2 denotes the average noise power and the $(N \times N)$ channel matrices of the desired signals and of the delayed signals are respectively given by

$$\mathbf{R}_x = \mathbf{H}_x\mathbf{H}_x^H \tag{11}$$
$$\mathbf{R}_y = \mathbf{H}_y\mathbf{H}_y^H. \tag{12}$$

From (6), the SNIR is written as

$$\begin{aligned}\mathrm{SNIR} &= \frac{E\left[z_x z_x^H\right]}{E\left[z_y z_y^H\right]+E\left[z_n z_n^H\right]} \\ &= \frac{\sigma_s^2 \mathbf{w}_a^H \mathbf{R}_x \mathbf{w}_a}{\sigma_s^2 \mathbf{w}_a^H \mathbf{R}_y \mathbf{w}_a + \sigma_n^2 \mathbf{w}_a^H \mathbf{w}_a} \\ &= \frac{\sigma_s^2 \mathbf{w}_a^H \mathbf{R}_x \mathbf{w}_a}{\mathbf{w}_a^H\left(\mathbf{R}_r - \sigma_s^2 \mathbf{R}_x\right)\mathbf{w}_a}. \end{aligned} \tag{13}$$

The optimum array weight vector $\mathbf{w}_{a,\mathrm{opt}}$ to maximize the SNIR of (13) is given by the solution of the maximization problem

$$\begin{aligned}&\text{maximize} \quad \frac{\sigma_s^2 \mathbf{w}_a^H \mathbf{R}_x \mathbf{w}_a}{\mathbf{w}_a^H\left(\mathbf{R}_r - \sigma_s^2 \mathbf{R}_x\right)\mathbf{w}_a} \\ &\text{subject to} \quad \mathbf{w}_a^H \mathbf{w}_a = 1. \end{aligned} \tag{14}$$

Using the method of Lagrange multipliers [19], $\mathbf{w}_{a,\mathrm{opt}}$ is obtained by taking the derivative of SNIR $+\ \beta\left(\mathbf{w}_a^H \mathbf{w}_a - 1\right)$ with respect to $\mathbf{w}_a^*$, where β is a real Lagrange multiplier. The problem in (14) then becomes the *generalized* eigenvalue problem

$$\begin{aligned}\mathbf{R}_x \mathbf{w}_{a,\mathrm{opt}} &= \frac{\mathbf{w}_{a,\mathrm{opt}}^H \mathbf{R}_x \mathbf{w}_{a,\mathrm{opt}}}{\mathbf{w}_{a,\mathrm{opt}}^H \mathbf{R}_r \mathbf{w}_{a,\mathrm{opt}}} \mathbf{R}_r \mathbf{w}_{a,\mathrm{opt}} \\ &= \lambda_{1,max} \mathbf{R}_r \mathbf{w}_{a,\mathrm{opt}}. \end{aligned} \tag{15}$$

It is worth noting that the result in [14] for the diversity antenna array can be considered as a special case of (15); that is, if there are no delayed signals, i.e., $\mathbf{R}_y = \mathbf{0}$, $\mathbf{w}_{a,\mathrm{opt}}$ becomes equivalent to the eigenvector of $\mathbf{R}_x$, which corresponds to the maximum eigenvalue.

2) With Virtual Subcarrier: The derivation of the optimum array weight set here is similar to that in the without-VS case; therefore, we give it in short. Defining the correlation matrices of the desired signals and of the delayed signals as

$$\mathbf{R}_x^c = \mathbf{H}_x E\left[\mathbf{S}_x \mathbf{S}_x^H\right] \mathbf{H}_x^H \tag{16}$$

$$\mathbf{R}_y^c = \mathbf{H}_y E\left[\mathbf{S}_y \mathbf{S}_y^H\right] \mathbf{H}_y^H \tag{17}$$

$$\mathbf{R}_{xy}^c = \mathbf{H}_x E\left[\mathbf{S}_x \mathbf{S}_y^H\right] \mathbf{H}_y^H \tag{18}$$

$$\mathbf{R}_{yx}^c = \mathbf{H}_y E\left[\mathbf{S}_y \mathbf{S}_x^H\right] \mathbf{H}_x^H = \left(\mathbf{R}_{xy}^c\right)^H \tag{19}$$

the optimum weight vector can be found from

$$\mathbf{R}_x^c \mathbf{w}_{a,\mathrm{opt}} = \lambda_{2,max}\left(\mathbf{R}_y^c + \sigma_n^2 \mathbf{I}\right) \mathbf{w}_{a,\mathrm{opt}}. \tag{20}$$

Since the receiver knows no information on delayed signals (unless we intentionally try to estimate them), (20) cannot be calculated directly. However, if the time difference in arrival time between the desired and delayed signals is not so small, from Appendix A we can reasonably assume that their cross-correlation is zero; that is, $\mathbf{R}_{xy}^c$ and $\mathbf{R}_{yx}^c$ are $\mathbf{0}$. Therefore, we can get a practically realizable equation

$$\mathbf{R}_x^c \mathbf{w}_{a,\mathrm{opt}} = \lambda_{3,max} \mathbf{R}_r \mathbf{w}_{a,\mathrm{opt}}. \tag{21}$$

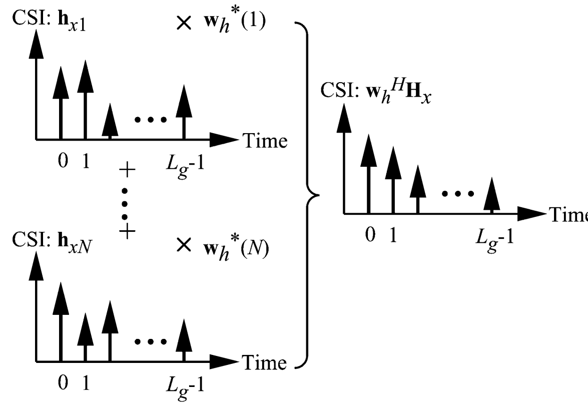

Fig. 6. Combined CSI.

which can be obtained by substituting $\mathbf{R}_x$ of (15) with $\mathbf{R}_x^c$.

B. mmse Criterion

In this approach, a reference signal is first generated by using a training signal with or without the aid of the estimated CSI. The array weight set is then determined based on the mmse criterion by employing an adaptive filtering such as least-mean square (LMS) or recursive least square (RLS) filtering [19] so as to make the array output signal closer to the reference signal. Depending on how to deal with a reference signal and how to optimize the array weight set, here we describe two schemes, as follows.

1) Suboptimal Scheme: Fig. 6 shows how to combine the CSI weight set in the suboptimal scheme. Let $p_{\mathrm{ref}} = \mathbf{w}_h^H \mathbf{H}_x \mathbf{S}_x$ be a reference signal, where $\mathbf{w}_h$ is an $(N \times 1)$ CSI weight vector for combining the CSI at each antenna element together. The cost function (J_1) to be minimized can be written as

$$\begin{aligned} J_1(\mathbf{w}_a) &= E\left[\left|p_{\mathrm{ref}} - \mathbf{w}_a^H \mathbf{r}\right|^2\right] \\ &= E\left[\left|\mathbf{w}_h^H \mathbf{H}_x \mathbf{S}_x - \mathbf{w}_a^H \mathbf{r}\right|^2\right] \end{aligned} \tag{22}$$

where $\mathbf{S}_x$ is replaced by the second half of the long training symbol during the adaptation stage. The remaining task is to determine the suitable CSI weight vector $\mathbf{w}_h$. Without *a priori* information on the delayed signals, the suitable weight set is one that yields the maximum ratio combining (MRC) for the desired signals. This is found to be the eigenvector of $\mathbf{R}_x$ (when VSs do not exist) or $\mathbf{R}_x^c$ (when VSs exist), which corresponds to the maximum eigenvalue and is equal to the optimum array weight set for the case where the delayed signals do not exist [11], [14].

The reference signal is determined without *a priori* information on the delayed signals; therefore, it may be no longer optimum, especially when DoAs of the desired signals and delayed signals are close together (closer than the beamwidth of the antenna array). This is why we call this scheme "suboptimal."

2) Weight Combining Scheme: Another looser setting of a reference signal is given by

$$p_{\mathrm{ref}} = \mathbf{w}_o^H \mathbf{S}_x \tag{23}$$

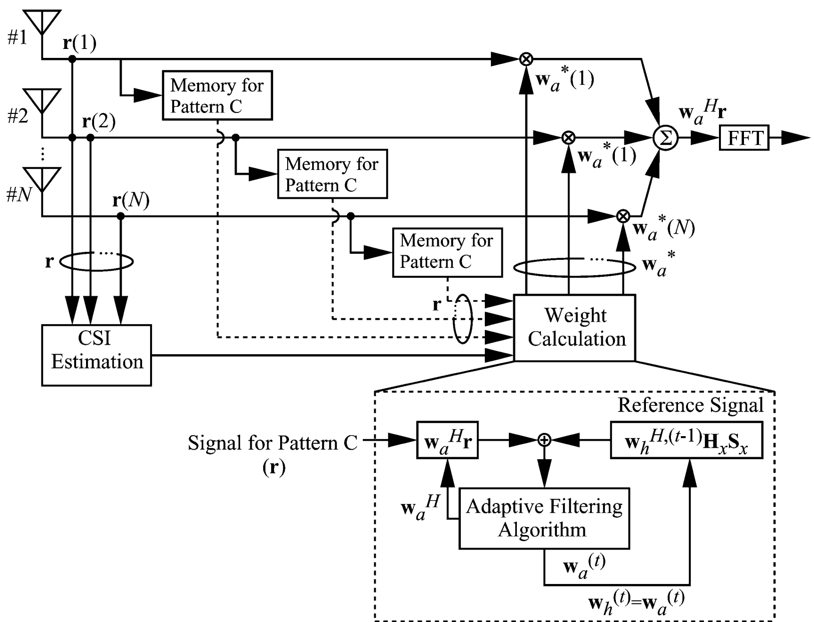

Fig. 7. Structure of the OFDM adaptive antenna array with the commutative optimization scheme.

where $(L_g \times 1)$ $\mathbf{w}_o$ is an arbitrary weight vector for gathering the desired signal components from the desired signal vector $\mathbf{S}_x$. The cost function (J_2) to be minimized can be written as

$$J_2(\mathbf{w}_o, \mathbf{w}_a) = E\left[\left|\mathbf{w}_o^H \mathbf{S}_x - \mathbf{w}_a^H \mathbf{r}\right|^2\right]. \tag{24}$$

Unlike the suboptimal scheme, which has a fixed reference signal, we now need to consider the problem of how to jointly determine both $\mathbf{w}_o$ and $\mathbf{w}_a$. Even though this problem has two *independent* variables to be jointly optimized, the cost function is a second-order function of the weight vectors; therefore, the error performance surface has a unique minimum (global minimum). Without any constraints on $\mathbf{w}_o$, it is clear that the minimization problem has a trivial solution of $\mathbf{w}_o = \mathbf{w}_a = \mathbf{0}$, where the bottom of the error performance bowl is at $J_2 = 0$.

To avoid the trivial solution, we impose one element of $\mathbf{w}_o$ (assumed to be the jth element, $\mathbf{w}_o(j)$) to be a constant and deal with the other remaining elements of $\mathbf{w}_o$ and $\mathbf{w}_a$ together as a new vector, $\mathbf{w}_n$. We call this scheme a "*weight-combining* scheme." After rewriting (24), the cost function to be optimized becomes

$$\begin{aligned} J_2(\mathbf{w}_n) &= E\left[\left|\mathbf{w}_o^*(j)\mathbf{S}_x(j) - \left(-\hat{\mathbf{w}}_o^H \hat{\mathbf{S}}_x + \mathbf{w}_a^H \mathbf{r}\right)\right|^2\right] \\ &= E\left[\left|\mathbf{w}_o^*(j)\mathbf{S}_x(j) - [-\hat{\mathbf{w}}_o; \mathbf{w}_a]^H \cdot [\hat{\mathbf{S}}_x; \mathbf{r}]\right|^2\right] \\ &= E\left[\left|\mathbf{w}_o^*(j)\mathbf{S}_x(j) - \mathbf{w}_n^H \cdot [\hat{\mathbf{S}}_x; \mathbf{r}]\right|^2\right] \end{aligned} \tag{25}$$

where $\hat{\mathbf{w}}_o$ and $\hat{\mathbf{S}}_x$ are the jth-element-reduced vectors of the weight and desired signal, respectively, and $\mathbf{S}_x(j)$ is the jth element of the vector $\mathbf{S}_x$. Here, the operation $[\mathbf{v}_1; \mathbf{v}_2]$ of a $(L_1 \times 1)$ vector $\mathbf{v}_1$ and a $(L_2 \times 1)$ vector $\mathbf{v}_2$ is defined as $[\mathbf{v}_1(1), \mathbf{v}_1(2), \ldots, \mathbf{v}_1(L_1), \mathbf{v}_2(1), \mathbf{v}_2(2), \ldots, \mathbf{v}_2(L_2)]^T$. The new combined vector $\mathbf{w}_n$ is given as

$$\mathbf{w}_n = [-\hat{\mathbf{w}}_o; \mathbf{w}_a]. \tag{26}$$

TABLE I
SYSTEM PARAMETER

Antenna Array	
Number of elements	8 (circular)
Element spacing	Half wavelength
OFDM	
Modulation/Detection	QPSK/Coherent detection
Number of subcarriers (N_c)	52 (including 4 pilot subcarriers)
FFT/Guard interval length	64/16 [samples]
Over sampling factor	4
FEC	
Coding/Decoding	Convolutional coding/ Soft Viterbi decoding
Code rate (R_c)	$1/2$
Constraint length (K_c)	7
Interleaver type/depth	Block interleaver/12×8

Since the delay times of the received signals are known from the estimated CSI, the dimension of $\mathbf{w}_o$ can be reduced from L_g to the number of arrival paths. Note that this constraint trick is similar to that employed in [20] for solving the problem of insufficient degrees of freedom of adaptive array and equalizer in a single carrier modulation (SCM) scenario. However, for this case, since most degrees of freedom of the reference signal is still freely adjusted, one could not guarantee that the mmse-criterion-based weight-combining scheme leads to the Max-SNIR criterion as desired.

IV. mmse-Criterion-Based Commutative Optimization Scheme

The MRC-based reference signal in the suboptimal scheme is not certainly optimum for every channel condition. In this section, we propose a novel scheme that is still based on the mmse criterion but can commutatively optimize $\mathbf{w}_h$ and $\mathbf{w}_a$.

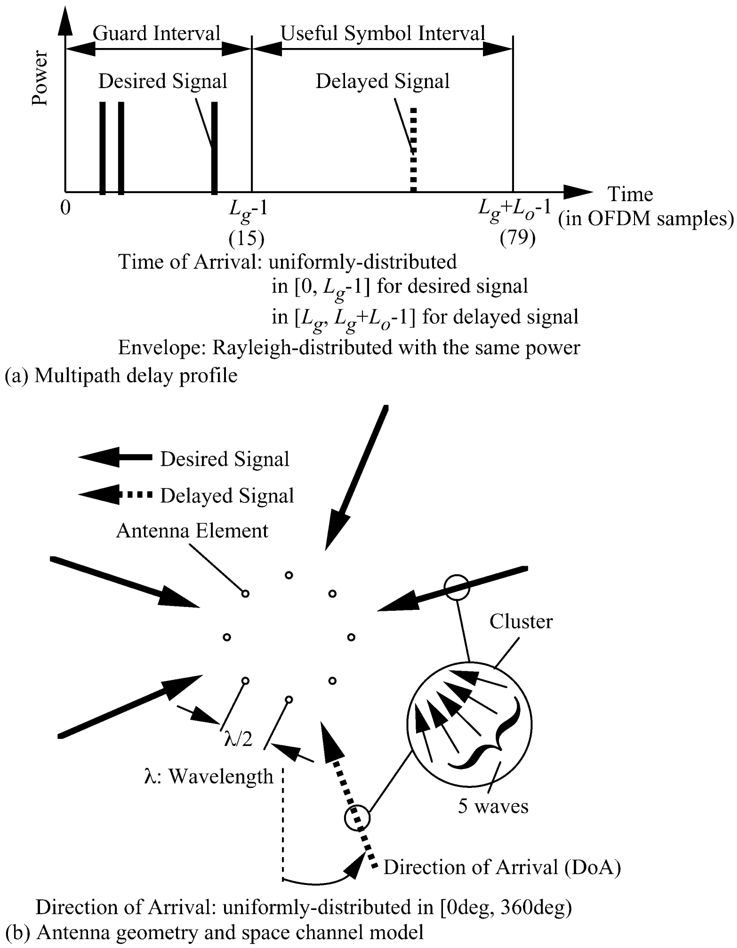

Fig. 8. Space–time channel model.

With the assumption that the desired signals and delayed signals are uncorrelated, the cost function J_1 in (22) can be rewritten as

$$\begin{aligned} J_1(\mathbf{w}_h, \mathbf{w}_a) &= E\left[\left(\mathbf{w}_h^H \mathbf{H}_x \mathbf{S}_x - \mathbf{w}_a^H \mathbf{r}\right)^2\right] \\ &= \mathbf{w}_h^H \mathbf{R}_x^c \mathbf{w}_h - \mathbf{w}_h^H \mathbf{R}_x^c \mathbf{w}_a \\ &\quad - \mathbf{w}_a^H \mathbf{R}_x^c \mathbf{w}_h + \mathbf{w}_a^H \mathbf{R}_r \mathbf{w}_a. \end{aligned} \tag{27}$$

To determine $\mathbf{w}_h$ and $\mathbf{w}_a$ commutatively, we first fix $\mathbf{w}_h$ and take the derivative of conditional mean square error (mse) $J_1(\mathbf{w}_a|\mathbf{w}_h)$ with respect to $\mathbf{w}_a^*$. The conditional optimum array weight vector $\tilde{\mathbf{w}}_{a,\mathrm{opt}}$ when $\mathbf{w}_h$ is given is found to be

$$\tilde{\mathbf{w}}_{a,\mathrm{opt}} = \mathbf{R}_r^{-1} \mathbf{R}_x^c \mathbf{w}_h. \tag{28}$$

Next, we fix $\mathbf{w}_a$ and take the derivative of conditional mse $J_1(\mathbf{w}_h|\mathbf{w}_a)$ with respect to $\mathbf{w}_h^*$. The conditional optimum CSI weight vector $\tilde{\mathbf{w}}_{h,\mathrm{opt}}$ when $\mathbf{w}_a$ is given is found to be

$$\tilde{\mathbf{w}}_{h,\mathrm{opt}} = \mathbf{w}_a. \tag{29}$$

Fig. 7 shows the structure of the OFDM adaptive antenna array with the proposed commutative optimization scheme. The algorithm is summarized as follows. We start from calculating $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(1)}$ of (28) with an initial CSI weight vector $(\mathbf{w}_h^{(0)})$ by using an adaptive filtering algorithm so as to make the array output signal be as close as to the reference signal. The initial weight vector $\mathbf{w}_h^{(0)}$ should be (but not necessary) the eigenvector that corresponds to the maximum eigenvalue of $\mathbf{R}_x^c$ (when VSs exist) or $\mathbf{R}_x$ (when VSs do not exist). From (29), the new CSI weight vector $\tilde{\mathbf{w}}_{h,\mathrm{opt}}^{(1)}$ is directly given by $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(1)}$ and then $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(2)}$ of the next round of commutation is calculated. One can heuristically see that the steady state could be reached at a time t when $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(t)}$ does not essentially change in direction compared with $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(t-1)}$. In other words, $\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(t)} \approx K\tilde{\mathbf{w}}_{a,\mathrm{opt}}^{(t-1)}$, where K is a complex constant. Note that if the number of total iterations of an adaptive filtering for all rounds of commutation exceeds the number of samples obtained from the second half of the long training symbol, the memory for the training symbol is needed.

This scheme is a *descent method* because as each new weight vector is generated, the difference between the corresponding array weight vector and the optimum array weight vector decreases in value (i.e., the scheme possesses a descent property) [21]. The descent property of this scheme is clearly proved in Appendix B. In the following, we briefly show that the commutative optimization scheme could be equivalent to the Max-SNIR scheme. With the fact that the weight set could converge to the correct solution, the weight vector $\mathbf{w}_{h,\mathrm{opt}}$ is obtained by the solution of the constraint minimization problem

$$\begin{aligned} \text{minimize} \quad & J_1(\mathbf{w}_h, \mathbf{R}_r^{-1}\mathbf{R}_x^c \mathbf{w}_h) \\ \text{subject to} \quad & \mathbf{w}_h^H \mathbf{R}_x^c \mathbf{w}_h = 1. \end{aligned} \tag{30}$$

The constraint in (30) is adopted for fixing the power of a reference signal to be constant or not be zero, which is always true for this scheme. Therefore, the trivial solution could be avoided. It is not difficult to further show that the solution of the above problem is equivalent to (21) of the Max-SNIR criterion.

V. Characteristics of the Proposed Array

The equations given by (15) and (21) for the Max-SNIR scheme and the cost functions given by (22), (25), and (27) for the mmse, the weight combining, and the commutative schemes contain only the parameters on the desired signals and received signals, so all the proposed schemes require no *a priori* knowledge that cochannel interference is delayed OFDM signals. Only the condition for the interference to be suppressed is its uncorrelation with the desired signals given by (9). Therefore, the proposed schemes need to estimate the CSI only within the guard interval and can suppress any cochannel interference that satisfies this condition. Temporal domain signal processing techniques [5], [6] cancel delayed signals using *a priori* information on them, so they need to estimate CSI estimation including the delayed signals. Therefore, they can cope only with delayed signals as cochannel interference; however, some techniques [6] can effectively utilize their power for data demodulation.

On the other hand, in the signal burst shown in Fig. 3, the pilot signals available for array weight control are not imbedded in the payload [15]; therefore, the array weight adaptation needs to be terminated at the end of the preamble and the obtained beam pattern is kept until the end of the burst. This implies that, if the channel fading variation is not negligibly small over one signal burst, then the performance degradation is expected, with more bit errors around at the end of the burst.

VI. Computer Simulation Results and Discussions

In this section, we demonstrate the performance of the proposed schemes by computer simulation. Table I summarizes the system parameters for the computer simulation. To avoid symmetric beam patterns obtained, a circular array is employed. In addition, taking into consideration its application for a base station in a cellular environment and for an access point and a note-size personal computer in a 5-GHz band wireless local area network (WLAN) environment, a *half*-wavelength adjacent spacing is employed with eight antenna elements. Here again, we assume that the rate of the channel variation is negligibly small over one signal burst. In addition, we assume a coherent quaternary phase-shift keying (QPSK) with a half-rate convolutional encoding/Viterbi decoding with a constraint length (K_c) of 7 and a (12×8) block interleaver. The subcarrier arrangement in this paper is all the same as in the high-rate wireless local area network (LAN) family [15]. One OFDM symbol is composed of 80 samples, where the guard interval length is 16 samples and the useful symbol length is 64 samples. Furthermore, the OFDM symbol is generated with the 64-point inverse FFT (IFFT), where only 48 subcarriers convey information, four subcarriers are known pilot signals, and other the 12 subcarriers are VSs.

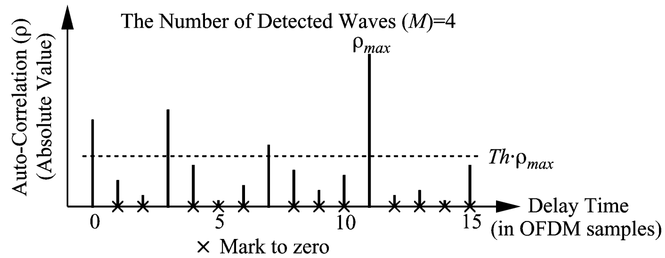

Fig. 9. Channel estimation method.

Fig. 8(a) and (b) show a multipath delay profile and an antenna geometry/space channel model, respectively, where there are four desired signals arriving within the guard interval and one delayed signal arriving beyond the guard interval. The envelope of each signal is Rayleigh distributed with the same power. The arrival time of each desired signal is uniformly distributed within the guard interval and that of the delayed signal is uniformly distributed within the useful symbol interval. Each signal arrives at the antenna array forming a cluster that is composed of five waves and the DoA of each signal is uniformly distributed in $[0^\circ, 360^\circ)$. Spatial fading among antenna elements is correlated with the exact correlation depending on the antenna acing and the angular spread observed [22]. Therefore, assuming that the microdistribution of the individual cluster is a uniform distribution, we discuss the performance of the proposed schemes for mainly two cases, namely, uniform distributions over 5° and 30° width (smaller angular spread) and a uniform distribution over 360° width (larger angular spread). The channel model with smaller angular spread is suited for a case where the place of the array antenna is sufficiently high so that few local scattering occurs [23], while that with larger angular spread is suited for a 5-GHz-band indoor WLAN system that is rich in local scatterers [24].

A. Channel-Estimation Method

The autocorrelation (ρ) at each antenna array is first calculated by employing the first half of the long training symbol. Fig. 9 shows a typical autocorrelation (in absolute value). We deem that a desired signal with arrival delay time i exists if the correlation value at time i, ρ_i exceeds the product of a threshold Th and a maximum correlation value $\rho_{\max}$, $(Th \cdot \rho_{\max})$. The number of detected waves is assumed to be M and, for example, in Fig. 2, four incoming desired signals are detected, i.e., $M = 4$. Based on this estimated correlation function and path decision, the CSI matrix $\mathbf{H}_x$ is then determined.

B. BER Performance

Fig. 10 shows the BER versus the threshold Th of the proposed pre-FFT OFDM adaptive antenna array system with the estimated CSI when the ratios of the average received signals (desired and delayed signals) energy per bit to the white noise power spectral density $(\overline{E_b/N_0s})$ per antenna are 4 and 8 dB, respectively. Here, loss due to the guard interval is not taken into the account of $\overline{E_b/N_0}$ calculation. The four schemes considered here are the Max-SNIR given by (21), suboptimal scheme, weight combining, and commutative optimization, respectively.

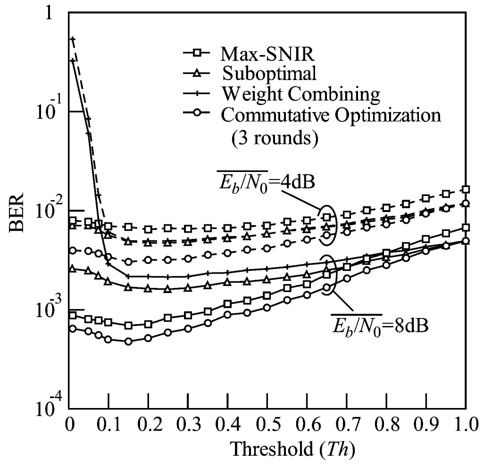

Fig. 10. BER versus the path-selection threshold.

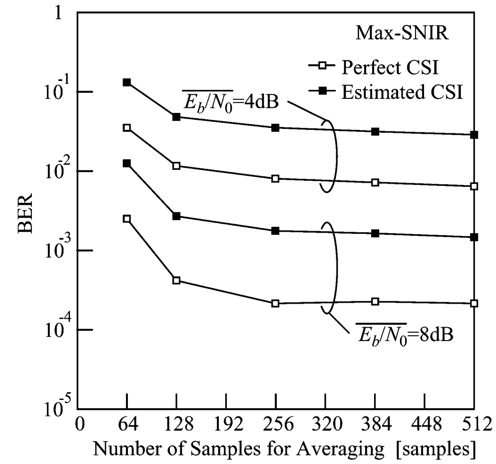

Fig. 11. BER of the Max-SNIR scheme versus the number of samples for estimating $\mathbf{R}_r$.

The received preamble is four times oversampled to yield 320 ($= 80 \times 4$) samples per OFDM symbol. For the Max-SNIR scheme, 256 received signal samples corresponding to the second half of the long training symbol are employed for estimating the $\mathbf{R}_r$ of (21). For the mmse-based schemes, the RLS filtering is employed. The number of RLS iterations (N_{cal}) is 320 for the suboptimal and weight-combining schemes. For the commutative optimization scheme with three-round commutation, the numbers of RLS iterations for the first, second, and third rounds are 150, 200, and 300, respectively. Here, the total number of RLS iterations for the commutative optimization scheme with three-round commutation is around two times of the suboptimal and weight combining schemes'.

Setting a smaller Th increases the probability that paths caused by noise are wrongly *selected*, while setting a lager Th increases the probability that real paths are wrongly *not selected*. Therefore, for a given $\overline{E_b/N_0}$, there is an optimum value in the path selection threshold to minimize the BER. From this figure, in the following simulation, we set the threshold to be 0.15, 0.20, 0.25, and 0.15 for the Max-SNIR, suboptimal, weight combining, and commutative optimization schemes, respectively.

Fig. 11 shows the effect of the number of samples for estimating $\mathbf{R}_r(N_{\mathrm{avg}})$ on the BER performance of the Max-SNIR scheme when $\overline{E_b/N_0}s$ are 4 and 8 dB, respectively. The $\mathbf{R}_r$ estimation is carried out employing the samples corresponding to the long training symbol. It can be seen that the BER performance is almost constant when $N_{\mathrm{avg}} > 256$, so in the following simulation we set $N_{\mathrm{avg}} = 256$.

Figs. 12 and 13 show the BER versus the $\overline{E_b/N_0}$ of the proposed four schemes with the perfect and the estimated CSIs, respectively. Here, the angular spread width is 5°. The BER performance when (20) with perfectly known parameters is applied is also plotted in Figs. 12 and 13, which is considered as the BER lower bound for all the schemes. We can see that the effect of the delayed signal on the BER performance is sufficiently suppressed. Among the four schemes, the BER performance of the suboptimal and weight-combining schemes are much inferior to that of the Max-SNIR and commutative optimization schemes in higher $\overline{E_b/N_0}$ region. We can also see that even though the BER performance of the commutative optimization scheme with three-round commutation is worse than that of the Max-SNIR scheme when the perfect CSI is assumed, it becomes superior when the estimated CSI is used.

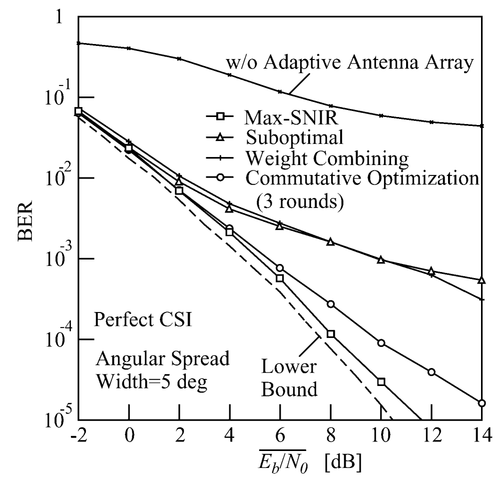

Fig. 12. BER performance of the four schemes with perfect CSI for angular spread width $= 5^\circ$.

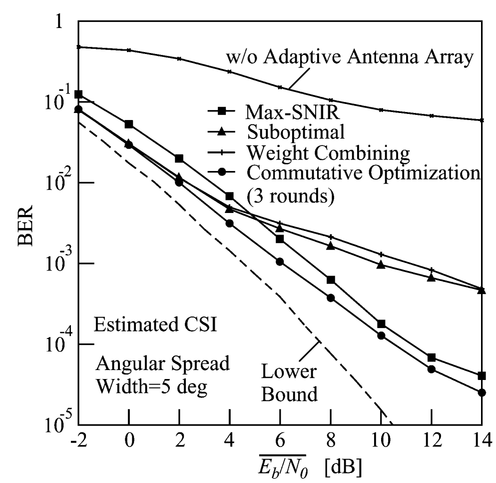

Fig. 13. BER performance of the four schemes with estimated CSI for angular spread width $= 5^\circ$.

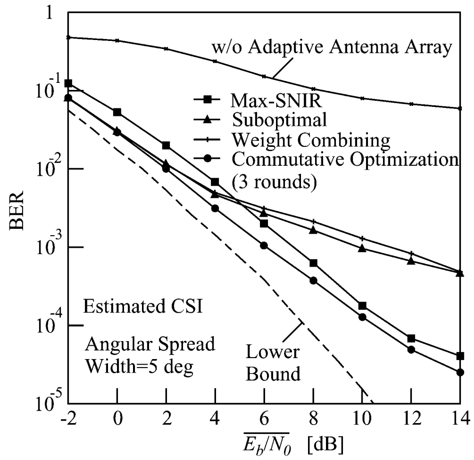

Fig. 14. BER performance of the four schemes with estimated CSI for angular spread width $= 5°$.

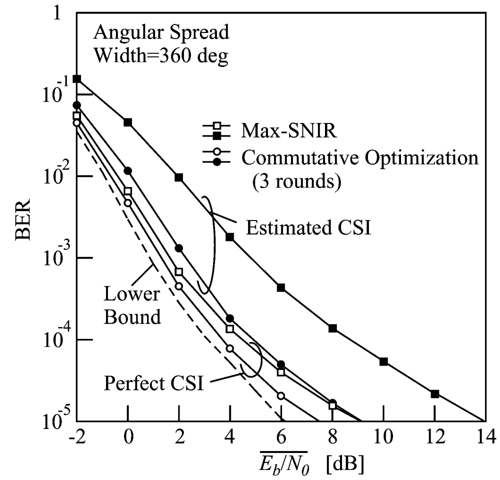

Fig. 15. BER performance of the four schemes with estimated CSI for angular spread width $= 5°$.

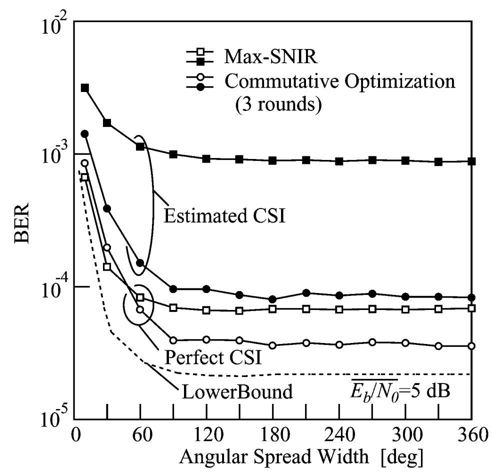

Fig. 16. BER versus the angular spread width.

It is well known that in SCM adaptive antenna array system when the DoAs of desired and interfering signals are very close, i.e., smaller than the beamwidth of the antenna array, the output SNIR could be drastically degraded due to high side lobe level and spatial whiteness of noise; in other words, *noise enhancement*. However, in the OFDM adaptive antenna array system with the Max-SNIR or commutative optimization schemes, multiple arrival paths within the guard interval can be considered as desired signals and be *appropriately* combined together with one array weight set. Therefore, the noise-enhancement degradation could be alleviated and the good BER performance is obtained even in a channel where DoAs of incoming waves are randomly determined. On the other hand, since the MRC-based reference signal in the suboptimal scheme is fixed, the above problem as in the SCM case could occur when the DoA of a delayed signal is close to that of a desired signal. The problem is serious especially when the delayed signal has a large channel gain while the desired signal has a low channel gain. In this case, suppression of the desired signal together with the delayed signal can much improve the array output SNIR, because the obtained antenna beam pattern can decrease the delayed signal power much more than the desired signal power, but it requires a change of the reference signal in the weight-adaptation process. With the commutative optimization scheme, the new reference signal for the next commutation round is automatically adjusted so as to make the output SNIR be even larger at the sacrifice of the desired signal power. Therefore, this problem can be alleviated.

Figs. 14 and 15 show the BER versus the $\overline{E_b/N_0}$ of the Max-SNIR and commutative optimization schemes with the estimated CSI for the angular spread widths of 30° and 360°, respectively. In addition, Fig. 16 shows the BER versus the angular spread width for $\overline{E_b/N_0} = 5$ dB. Comparing the performance of the same scheme in among Figs. 13–15 and also from Fig. 16, we can see that the BER performance is improved as the angular spread width increases. This is because the array output SNIR is improved only by means of antenna gain when the angular spread is small, while it is improved not only by means of antenna gain, but also by means of diversity gain when the angular spread becomes larger. However, the performance improvement of the Max-SNIR scheme as the increase of the angular spread is much less than that of the commutative optimization scheme. This is because the Max-SNIR scheme in principle has a difficulty in exact calculation of the autocorrelation of the received signals and the increase of the angular spread width magnifies the difficulty. This is clearly observed in Fig. 16, where the Max-SNIR scheme is inferior to the commutative optimization scheme even for the perfect CSI case when the angular spread width is larger than around 50°. For the estimated CSI case, the commutative optimization scheme always performs better than the Max-SNIR scheme; therefore, we can conclude that it is more robust to the CSI estimation error.

Finally, we would like to note here that the assumption of uncorrelated sample pairs, which is no longer valid when the estimate is extracted on the basis of a single realization of a correlated process (OFDM signal with VSs), i.e., a single data record, unavoidably causes some degradation in BER performance of the Max-SNIR scheme [25].

VII. Conclusion

In this paper, we proposed a novel pre-FFT OFDM adaptive antenna array for suppressing delayed signal beyond the guard interval. Based on the Max-SNIR and mmse criteria, respectively, the optimum array weights were derived. Furthermore, we proposed the mmse-criterion-based commutative optimization scheme that was more robust to the estimation error of the CSI than the Max-SNIR-criterion-based scheme and could give the best BER performance among the four schemes, with the proof on the equivalence between the two schemes. Computer simulation results showed its good performance even in channels where DoAs of incoming waves were randomly determined.

Appendix A

A. Correlation Property of OFDM Signal

The OFDM signal in discrete equivalent low-pass representation can be written as

$$x(n) = \sum_{k=0}^{N_c-1} X(k) e^{j2\pi kn/N} \tag{31}$$

where $X(k)$ and N_c are a symbol data to be placed on the kth subcarrier and the number of FFT operation points, respectively. The autocorrelation function, which is defined as ensemble average, is written as

$$\begin{aligned} R(n_1, n_2) &= R(n_1 - n_2) = E\left[x(n_1) x^*(n_2)\right] \\ &= E\left[\sum_{k_1=0}^{N_c-1} X(k_1) e^{j2\pi k_1 n_1/N_c} \cdot \sum_{k_2=0}^{N_c-1} X^*(k_2) e^{-j2\pi k_2 n_2/N_c}\right] \\ &= E\left[\sum_{k_1=k_2=k=0}^{N_c-1} |X(k)|^2 e^{j2\pi k(n_1-n_2)/N_c} + \sum_{k_1 \neq} \sum_{k_2} X(k_1) X^*(k_2) e^{j2\pi(k_1 n_1 - k_2 n_2)/N_c}\right]. \end{aligned}$$

It is reasonable to assume that data symbols are uncorrelated, i.e., for $k_1 \neq k_2$, $E[X(k_1)X^*(k_2)] = 0$. We further assume a case of phase-shift keying (PSK) modulation where $E[|X(k)|^2] = \sigma^2$ for all k. Thus, $R(n_1, n_2)$ can be written as

$$R(n_1, n_2) = \sigma^2 \sum_{k=0}^{N_c-1} e^{j2\pi k(n_1-n_2)/N_c} = \sigma^2 \Delta. \tag{32}$$

It can be seen that if there are no VSs, the summation term Δ is zero when $n_1 \neq n_2$ and is N_c when $n_1 = n_2$; that is, the correlation function is given by

$$\begin{aligned} R(n_1, n_2) &= 0 \quad n_1 \neq n_2 \\ &= N_c \sigma^2 \quad n_1 = n_2. \end{aligned} \tag{33}$$

This has shown the *white* property of the OFDM signal. On the other hand, if there are δ subcarriers of VSs and, for an example, let $|n_1 - n_2| = 1$, it can be easily shown that the correlation function becomes

$$\begin{aligned} R(n_1, n_2) &\leq \delta\sigma^2 \quad |n_1 - n_2| = 1 \\ &= (N_c - \delta)\sigma^2 \quad n_1 = n_2. \end{aligned} \tag{34}$$

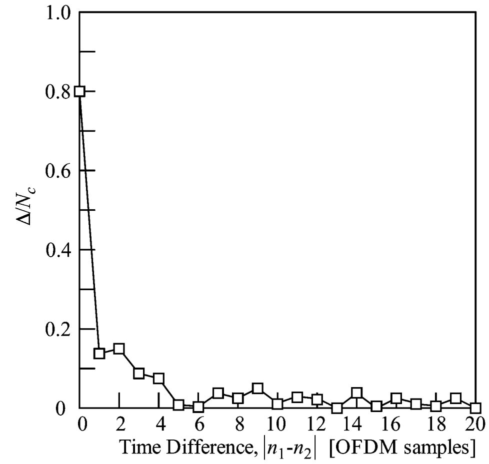

Fig. 17. Δ/N_c versus $|n_1 - n_2|$.

This has shown that the OFDM signal is *colored* when there exist VSs. In another point of view, the VSs of OFDM signal show the discontinuous frequency spectrum; hence, its time-domain signal is not white. Fig. 17 shows the Δ/N_c versus $|n_1 - n_2|$ according to the subcarrier arrangement in this paper.

Appendix B

A. Descent Property of the Commutative Optimization Scheme

Let $A = \mathbf{R}_r^{-1}\mathbf{R}_x^c$. Refer from (28) and (29) that we can write the array weight vector at the tth-round commutation as

$$\tilde{\mathbf{w}}_{a,\text{opt}}^{(t)} = A^t \mathbf{w}_h^{(0)}. \tag{35}$$

By unitary similarity transformation, we can write $A = Q\Lambda Q^H$, where $\Lambda = diag(\lambda_1, \lambda_1, \ldots, \lambda_N)$ is the diagonal matrix whose diagonal elements consist of the eigenvalues λ_i of the matrix A, and $Q = [\mathbf{v}_1, \mathbf{v}_2, \ldots, \mathbf{v}_N]$ is the unitary matrix whose columns are the eigenvectors $\mathbf{v}_i$ corresponding to λ_i. The $\tilde{\mathbf{w}}_{a,\text{opt}}^{(t)}$ can be expressed as

$$\begin{aligned} \tilde{\mathbf{w}}_{a,\text{opt}}^{(t)} &= Q\Lambda^t Q^H \mathbf{w}_h^{(0)} \\ &= \sum_{i=1}^{N} \lambda_i^t \left(\mathbf{v}_i^H \mathbf{w}_h^{(0)}\right) \mathbf{v}_i. \end{aligned} \tag{36}$$

Without loss of generality, we assume that the trace of A is normalized to be unity so that $\forall\ \lambda_i < 1$. We can see that as t approaches infinity, $\tilde{\mathbf{w}}_{a,\text{opt}}^{(t)}$ always approaches $K\mathbf{v}_1$ or $\mathbf{w}_{a,\text{opt}}$, where K is a constant, if λ_1 is the largest eigenvalues. This is because as t increases, λ_i^t $(i \neq 1)$ decreases very quickly, compared with λ_1^t. It can be seen from the above equation that the convergence rate is in exponential order of λ_i and depends on the difference between the maximum eigenvalue and other eigenvalues. The fast convergence rate is also observed by the computer simulation.

ACKNOWLEDGMENT

The authors would like to thank Dr. H. Kazunori, Assistant Professor, Kyoto University, Kyoto, Japan, and Mr. K. Oketani of NEC Electric Corporation,Kanagawa, Japan, for their helpful suggestions.

REFERENCES

[1] J. A. C. Bingham, "Multicarrier modulation for data transmission: An idea whose time has come," *IEEE Commun. Mag.*, vol. 28, pp. 5–14, May 1990.

[2] S. Hara, M. Mouri, M. Okada, and N. Morinaga, "Transmission performance analysis of multi-carrier modulation in frequency selective fast Rayleigh fading channel," *Wireless Pers. Commun.*, vol. 2, no. 4, pp. 335–356, 1995/1996.

[3] R. van Nee, M. Morikura, H. Takanashi, and K. Halford, "New high rate wireless LAN standards," *IEEE Commun. Mag.*, vol. 37, pp. 82–88, Dec. 1999.

[4] J. Chuang and N. Sollenberger, "Beyond 3G: Wideband wireless data access based on OFDM and dynamic packet assignment," *IEEE Commun. Mag.*, vol. 38, pp. 78–87, July 2000.

[5] Y. Sun and L. Tong, "Channel equalization for wireless OFDM systems with ICI and ISI," in *Proc. Int. Conf. Communications (ICC'99)*, June 1999, pp. 182–186.

[6] S. Suyama, Y. Hara, H. Suzuki, Y. Kamio, and K. Fukawa, "A maximum likelihood OFDM receiver with smoothed FFT-window for large multipath delay difference over the guard interval," in *Proc. Vehicular Technology Conf. (VTC'02)*, May 2002, pp. 1247–12 521.

[7] J. H. Winters, J. Salz, and R. D. Gitlin, "The impact of antenna diversity on the capacity of wireless communication systems," *IEEE Trans. Commun.*, vol. 42, pp. 1740–1751, Apr. 1994.

[8] R. A. Monzingo and T. W. Miller, *Introduction to Adaptive Arrays*. New York: Wiley, 1980.

[9] R. T. Compton, *Adaptive Antennas*. Englewood Cliffs, NJ: Prentice-Hall, 1988.

[10] J. H. Winters, "Optimum combining in digital mobile radio with cochannel interference," *IEEE Trans. Veh. Technol.*, vol. VT-33, pp. 144–155, Aug. 1984.

[11] S. Hara, A. Nishikawa, and Y. Hara, "A novel OFDM adaptive antenna array for delayed signal and Doppler-shifted signal suppression," in *Proc. IEEE Int. Conf. Communications (ICC'01)*, June 2001, pp. 2302–2306.

[12] M. Budsabathon, Y. Hara, and S. Hara, "A pre-FFT OFDM adaptive antenna array for delayed signal suppression," in *Proc. 7th Int. OFDM Workshop*, Sept. 2002, pp. 92–96.

[13] Y. (G.) Li and N. R. Sollenberger, "Adaptive antenna arrays for OFDM system with cochannel interference," *IEEE Trans. Commun.*, vol. 47, pp. 217–229, Feb. 1999.

[14] M. Okada and S. Komaki, "Pre-DFT combining space diversity assisted COFDM," *IEEE Trans. Veh. Technol.*, vol. 50, pp. 487–496, Mar. 2001.

[15] *Wireless Medium Access Control (MAC) and Physical Layer (PHY) Specifications: High-Speed Physical Layer Extension in the 5 GHz Band*, IEEE Standard 802.11a, 1999.

[16] S. Hara and M. Budsabathon, "Spread spectrum based subcarrier recovery method for multicarrier code division multiplexing system," *Eur. Trans. Telecommun.*, vol. 10, no. 4, pp. 369–376, 1999.

[17] *Lower Power Data Communication Systems Broadband Mobile Access Communication System (CSMA)*, ARIB STD-T70, Dec. 2000.

[18] "Broadband Radio Access Networks (BRAN); HIPERLAN Type2; Physical (PHY) Layer," ETSI BRAN, ETSI TR 101 475, 2000.

[19] S. Haykin, *Adaptive Filter Theory*, 3rd ed. Englewood Cliffs, NJ: Prentice-Hall, 1996.

[20] M. L. Leou, C. C. Yeh, and H. J. Li, "A novel hybrid of adaptive array and equalizer for mobile communications," *IEEE Trans. Veh. Technol.*, vol. 49, pp. 1–10, Jan. 2000.

[21] E. K. P. Chong and S. H. Zak, *An Introduction to Optimization*, 2nd ed. New York: Wiley, 2001.

[22] R. B. Ertel, P. Cardieri, K. W. Sowerby, T. S. Rappaport, and J. H. Reed, "Overview of spatial channel models for antenna array communication systems," *IEEE Pers. Commun.*, vol. 5, pp. 10–22, Feb. 1998.

[23] H. Bölsckei, D. Gesbert, and A. J. Paulraj, "On the capacity of OFDM-based spatial multiplexing systems," *IEEE Trans. Commun.*, vol. 50, pp. 225–234, Feb. 2002.

[24] C.-C. Chong, C.-M. Tan, D. I. Laurenson, S. McLaughlin, M. A. Beach, and A. R. Nix, "A new statistical wideband spatio-temporal channel model for 5-GHz band WLAN systems," *IEEE J. Select. Areas Commun.*, vol. 21, pp. 139–150, Feb. 2003.

[25] G. Jacovitti and R. Cusani, "Performances of the hybrid-sign correlation coefficient estimators for Gaussian stationary processes," *IEEE Trans. Acoustics, Speech, Signal Processing*, vol. ASSP-33, pp. 731–733, June 1985.

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