

Pre- and Post-FFT SIMO Array Techniques in Hiperlan/2 Environments

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Abstract—The performance of very high data rate Wireless Local Area Networks (WLANs) is degraded by several problems such as long delay spread in outdoor environments or interference that may arise both from other WLANs or from other systems sharing the same band. The current standards allow the use of multiple antennas at the receiver in a Single Input Multiple Output (SIMO) configuration. Traditionally, algorithms have been applied in the frequency domain, but this paper presents a comparison of three approaches to be applied also in the time domain. It is shown that a significant uncoded Bit Error Rate (BER) performance improvement can be achieved by working before the FFT at the receiver side, especially for channels with large delay spreads.

I. INTRODUCTION

The three major standards for WLAN, i.e. ETSI BRAN HL/2 [1] in Europe, IEEE 802.11a in the USA and ARIB MMAC in Japan, have chosen Orthogonal Frequency Division Multiplexing (OFDM) as their transmission mode. In their typical deployment scenarios, the use of an adequate cyclic prefix length may mitigate the effects of multipath propagation. The use of multiple antennas at the receiver in a standard-compliant SIMO configuration improves significantly the performance of these kind of systems.

Traditionally, algorithms have been applied in the frequency domain [2], [3]. This approach is well suited in scenarios where the cyclic prefix (CP) is able to cope with the multipath propagation. However, if the channel length exceeds the duration of the CP, OFDM suffers from Inter Symbol Interference (ISI) and Inter Carrier Interference (ICI), degrading the ultimate link quality. Therefore, beamforming in the time domain [4] improves the performance of the system by reducing the equivalent channel response duration.

In this paper, the performance of the traditional Sample Matrix Inversion (SMI) algorithm is tested for OFDM systems [5] compared to the Matched Desired Impulse Response (MDIR) [6] and the Maximum Ratio Combining (MRC) scheme. The first two algorithms are tested both in the frequency domain and in the time domain, whereas the latter is only computed in the frequency domain. These algorithms use *only* the known preamble to compute the beamformer, thus subcarrier grouping seems to be a convenient option as in [7]. In the time domain the optimum length of the filters at each of the antennas is found.

An adaptive scheme may also be considered by updating the weights on a frame by frame basis. The pilot position in HL/2 do not allow an accurate channel tracking. If one wants to deal with very bursty interference appearing in the middle of a user burst, block algorithms dealing with the whole set of OFDM symbols should be considered as the SMI-CP approach in [8].

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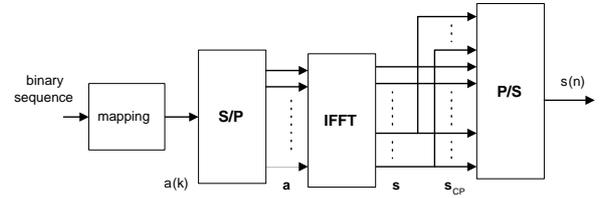


Fig. 1. Baseband transmitter (no channel coding).

This paper is organized as follows. In Section II, the signal model is firstly presented. Then, the two approaches at the receiver are presented, i.e. the *natural* frequency domain combining and the time diversity view. After that, the studied algorithms are presented, i.e. the MRC, the SMI and the MDIR, Section III. Finally, in IV results from simulations are presented and discussed, giving also some practical hints immediately before the final conclusions.

II. SYSTEM MODELING

In the following, boldface capital letters refer to matrices and lowercase boldface letters refer to vectors. The operator $(\cdot)^*$ denotes conjugation, $(\cdot)^T$ transposition, and $(\cdot)^H = ((\cdot)^*)^T$. The $N \times N$ unitary Fourier matrix is denoted by \mathbf{F} and \mathbf{F}^H refers to the IFFT_N operation.

A. Signal Model

At the transmitter (Figure 1), the mapped symbols feed the Serial to Parallel (S/P) block to perform the IFFT_N , i.e. $\mathbf{s}^{\text{CP}}(\mathbf{m}) = \mathbf{F}^H \mathbf{a}(\mathbf{m})$, where the vector $\mathbf{a}(\mathbf{m})$ has been filled with the mapped symbols with the corresponding padding zeros at the unused subcarriers. Then, the transmitted OFDM symbol m in the time domain with a cyclic prefix of L samples is $\mathbf{s}(\mathbf{m}) = [s(m, 0) s(m, 1) \dots s(m, P-1)]^T$, where $s(m, n) = s(m, n+N-1)$, $0 \leq n \leq L-1$ and $s(m, L) = s^{\text{CP}}(m, 0)$. The total number of samples is $P = N+L$.

Let $\mathbf{h}_q(\mathbf{m}) = [h_q(m, 0) h_q(m, 1) \dots h_q(m, K-1)]^T$ be the channel impulse response during OFDM symbol m of a given subchannel q , i.e. the channel from the single transmitter antenna to the receiver antenna q . The received OFDM symbol m in time domain can be expressed as

$$\mathbf{r}_q(\mathbf{m}) = \mathbf{H}_q(\mathbf{m})\mathbf{s}(\mathbf{m}) + \mathbf{v}_q(\mathbf{m}), \quad (1)$$

where $\mathbf{v}_q(\mathbf{m})$ contains the contribution from the AWGN, the interference sources and also the Inter Symbol Interference (ISI) from the previous symbol, while $\mathbf{H}_q(\mathbf{m})$ is the $P \times P$ Toeplitz filtering matrix with first row $[h_q(m, 0) 0 \dots 0]$ and first column $[\mathbf{h}_q(\mathbf{m})^T 0 \dots 0]^T$. The channel is considered time-invariant within a OFDM symbol.

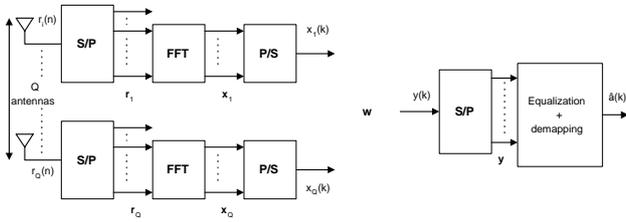


Fig. 2. Post-FFT combining in HL/2

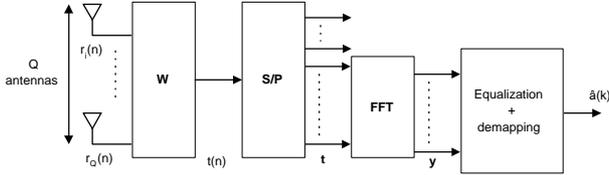


Fig. 3. Time diversity receiver for HL/2

III. ALGORITHMS

A. Frequency Domain

B. Time Domain

IV. SIMULATIONS

A. Practical Issues

V. CONCLUSIONS

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B. Post-FFT Combining

C. Time diversity Receiver

The time diversity receiver (Figure 3) performs the beamforming in the time domain as a filter bank, where each FIR filter for OFDM symbol m has M coefficients, i.e. $\mathbf{w}_q(\mathbf{m}) = [w_q(m, 0) \ w_q(m, 1) \ \dots \ w_q(m, M-1)]^T$. The filter vector $\mathbf{w}(\mathbf{m}) = [\mathbf{w}_1(\mathbf{m})^T \ \mathbf{w}_2(\mathbf{m})^T \ \dots \ \mathbf{w}_Q(\mathbf{m})^T]^T$ and $\mathbf{r}(\mathbf{m}, \mathbf{n}) = [\tilde{\mathbf{r}}_1(m, n)^T \ \tilde{\mathbf{r}}_2(m, n)^T \ \dots \ \tilde{\mathbf{r}}_Q(m, n)^T]^T$ are used to compute output sample n as $t(m, n) = \mathbf{w}(\mathbf{m})^H \mathbf{r}(\mathbf{m}, \mathbf{n})$. The received data vector containing the necessary samples to perform the filtering for antenna q th is $\tilde{\mathbf{r}}_q(m, n) = [r_q(m, n) \ r_q(m, n+1) \ \dots \ r_q(m, n+M-1)]^T$. In the following, $\mathbf{r}^i(\mathbf{m}, \mathbf{n})$ denotes the $MQ \times 1$ vector containing the $M-1$ samples of the Q antennas that follow sample n of the i th part in the received OFDM symbol m . An index of a sample exceeding $L-1$ refers to samples from next signal block, i.e. if $R > L-1$, $r_q^i(m, R) = r_q^{i+1}(m, R-L)$.

The output samples can be gathered up in a common vector $\mathbf{t}(\mathbf{m}) = [t(m, 0) \ t(m, 1) \ \dots \ t(m, P-1)]^T$. The cyclic prefix is disregarded by the matrix \mathbf{G} , therefore $\mathbf{y}(\mathbf{m}) = \mathbf{F}\mathbf{G}\mathbf{t}(\mathbf{m})$. In the sequel, $\mathbf{y}_u(\mathbf{m})$ denotes the $N_u \times 1$ vector containing only the useful subcarriers. If the global response of the channel and the spatio-temporal filter is shorter than the cyclic prefix, the mapped symbols $a(m, k)$ can be recovered by simple equalization in the frequency domain, i.e. $\hat{a}(m, k) = y_u(m, k) / \hat{c}(k)$, where $\hat{c}(k)$ denotes the estimated total response coefficient for subcarrier k .