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Abstract - **Coherent demodulation needs to maintain the symbol timing synchronization and to estimate the channel response in multicarrier code division multiple access systems. The use of inverse DFT (IDFT) is often considered to generate a pilot signal for this propose. However it requires a large implementation complexity, making it impractical for real applications. In this paper, we consider the design of a novel pilot signal for both symbol synchronization and channel estimation. By controlling the design parameters, the pilot signal can be generated to provide a desired signal-to-interference noise power ratio, while maintaining an acceptable peak-to-average power ratio. Simulation results show that the use of the proposed pilot signal can improve the detection performance compared to the use of conventional IDFT-based pilot signal, while significantly reducing the implementation complexity.**

I. INTRODUCTION

Multicarrier code division multiple access (MC-CDMA) schemes have been considered as one of major technologies for the next generation mobile radio communications, where the user signal is spread into multiple subcarriers using a spreading code [1-3]. The MC-CDMA receiver can exploit all the received signal scattered in the frequency domain using a coherent combining technique [1]. However, the use of coherent combining in the MC-CDMA system requires both accurate symbol timing synchronization and channel information.

Since the MC-CDMA scheme is very vulnerable to the timing offset of fast Fourier transform (FFT) window, it requires very accurate symbol synchronization. The symbol timing offset can destroy the orthogonality among the subcarriers and users' spreading codes, causing severe inter-symbol interference (ISI) and inter-carrier interference (ICI) [4]. The symbol timing can be acquired at the base station by correlating the received pilot signal of each user with a replica of the pilot signal. Since the base station suffers from multiple access interference (MAI) due to asynchronous multiple user signals, it is desirable for synchronization to use a pilot signal having good autocorrelation and crosscorrelation property [5, 6]. It is also desirable to use a pilot signal for channel estimation since we consider the use of coherent combining, such as equal gain combining (EGC) or maximum ration combining (MRC).

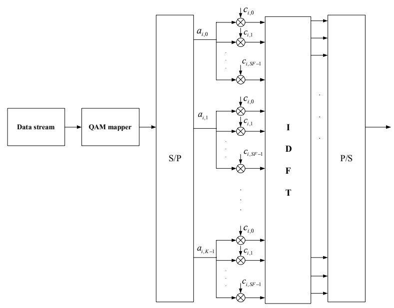
Significant amount of effort has been devoted for the design of pilot signals for synchronization and channel estimation in MC-CDMA or orthogonal frequency division multiplexing (OFDM) systems [7-9]. However, most of these works may not be applicable to the uplink system because they do not consider the MAI. For synchronization and channel estimation in the uplink, the pilot signal can be generated by using inverse discrete Fourier transform (IDFT) of an orthogonal Gold code [5, 6]. This pilot signal has good correlation characteristics for symbol timing acquisition. It can also be used for channel estimation by exploiting the code orthogonality. Since both synchronization and channel estimation can be achieved using one IDFT symbol, the bandwidth efficiency can be improved. However, the use of IDFT pilot signal requires a large computational complexity for the synchronization, while yielding the autocorrelation properties inferior to those of conventional Gold codes. Since the symbol timing synchronization must be achieved for all users in the cell, it may require a large computational complexity, making it impractical for real implementation.

To overcome this problem, we propose a novel pilot signal applicable to both synchronization and channel estimation. The proposed pilot signal can be designed using Gold codes. It can be used for symbol synchronization in the time domain and for channel estimation using the orthogonality property in the frequency domain. The use of the proposed pilot signal can improve the synchronization performance, while significantly reducing the computational complexity for the synchronization.

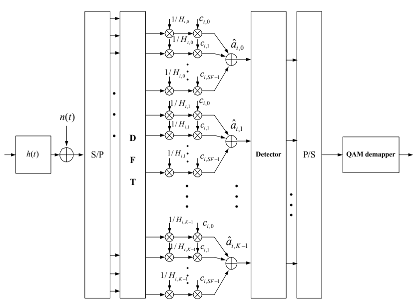
This paper is organized as follows. Section II describes a brief overview of symbol timing acquisition schemes in the MC-CDMA uplink system. A novel pilot signal design method is proposed in Section III. The timing detection performance and computational complexity of the proposed method are evaluated in Section IV. Finally, conclusions are given in Section V.

II. TIMING ACQUISITION IN MC-CDMA UPLINK

The MC-CDMA transmitter spreads the message symbol using a spreading code in the frequency domain. The basic transmitter structure of the MC-CDMA system is similar to that of OFDM. The main difference is that the MC-CDMA system transmits the same symbol in parallel over multiple subcarriers, whereas the OFDM system transmits one symbol per subcarrier [2]. Fig. 1. (a) depicts the MC-CDMA transmitter for the i-th user, $i = 0,1,\cdots I-1$. The input information sequence is first converted into K parallel QAM symbol sequences $(a_{i,0},\ a_{i,1},\ \cdots,\ a_{i,K-1})$ and then the output of the serial-to-parallel converter is multiplied by the i-th user's spreading code $(c_{i,0},\ c_{i,1},\ \cdots,\ c_{i,SF-1})$, where the subscript SF denotes the spreading factor. All the data in N ($= K \times SF$) subcarriers are modulated using the IDFT and then converted into a



(a) Transmitter



(b) Receiver

Fig. 1. The block diagram of MC-CDMA transceiver

serial data format. Fig. 1. (b) depicts the structure of coherent combing receiver of the i-th user. The subcarrier components corresponding to $a_{i,k}, 0 \le k \le K-1$, is first demodulated by the DFT and then equalized using an estimated channel coefficient $H_{i,k}$ of the i-th user and k-th frequency cluster. We assume that the channel is unchanged during the dispreading interval.

To demodulate the received signal without ISI or ICI, the symbol timing should be synchronized accurately. Since the distances of each user from the base station are different, the user signals are asynchronously received at the base station. To detect the beginning of a frame, the base station correlates the received pilot signal with a replica of the pilot signal and then finds the peak of the correlator output in the range of the timing detection window. The length N_D of the timing detection window can be determined by [6]

$$N_D = \left\lceil \frac{2R}{cT} \right\rceil \tag{1}$$

where $\lceil x \rceil$ denotes the largest integer less than x, R is the cell radius, c is the velocity of microwave and T is the sampling time interval.

Fig. 2 depicts a conventional acquisition scheme of the base station using this pilot signal. Assuming I is the number of active users in the cell, the timing offset of each user can be found using I correlators. The symbol timing of the i-th user can simultaneously be detected by finding the timing offset τ_i of the i-th user's pilot signal. The mobile station of the i-th user can compensate the timing offset using the received timing information.

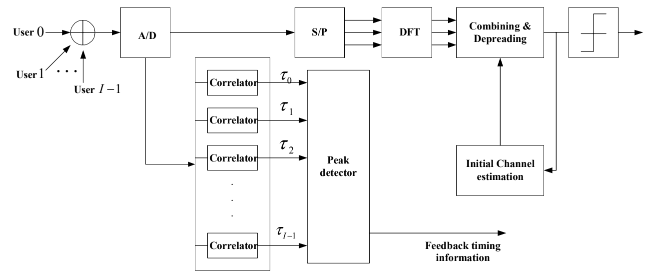


Fig. 2. Conventional acquisition scheme in the MC-CDMA uplink

III. The Proposed Pilot Signal

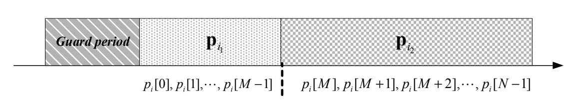
Each mobile station uses its own pilot symbol. It is desirable to use a pilot signal having good autocorrelation and crosscorrelation characteristics for timing recovery in the MC-CDMA uplink system [5, 6]. The use of binary Gold codes may be suitable for this propose [6]. But it requires an additional pilot symbol for channel estimation. To alleviate this problem, we design a novel pilot signal that can be used for both synchronization and channel estimation.

As illustrated in Fig. 3, a binary Gold code $\mathbf{p}_{i_1} (= p_i[0], p_i[1], \cdots, p_i[M-1])$ of length M is inserted in the time domain as the i-th user's pilot signal. The remaining $(N-M)$ signal $\mathbf{p}_{i_2}$ is designed such that the overall frequency response of the pilot signal becomes an orthogonal code. Assuming that K QAM symbols are transmitted in a single MC-CDMA symbol, the receiver needs the channel coefficient of these K clusters in the frequency domain [1]. The channel information can be estimated using K short orthogonal codes with a length of $L = (N-M)/K$.

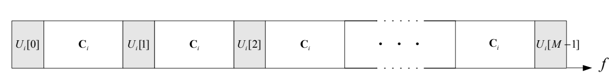
Assume that each user has its unique orthogonal code in the frequency domain represented as

$$\mathbf{C}_i = (C_i[0], C_i[1], \cdots, C_i[L-1]) \tag{2}$$

where the subscript i denotes the user index. We define the remaining M samples in the frequency domain as a redundant



(a) Time domain



(b) Frequency domain

Fig. 3. Structure of the proposed pilot signal

signal

$$\mathbf{U}_i = \left(U_i[0], U_i[1], \cdots, U_i[M-1]\right) \tag{3}$$

Since this signal is not used for channel estimation, it can arbitrarily be set to zero, to reduce the power of the redundant signal,

$$\mathbf{U}_i = (0,0,\ldots,0) = \mathbf{0} \tag{4}$$

Let $\mathbf{A}_i$ be the desired frequency response of the i-th user's pilot signal, comprising K short orthogonal codes and redundant signal, represented as

$$\mathbf{A}_i = \left(U[0], \mathbf{C}_i, U[1], \mathbf{C}_i, U[2], \cdots, \mathbf{C}_i, U[M-1]\right)^T \tag{5}$$

where the superscript T denotes the transpose of a matrix. Then $\mathbf{p}_{i_2}$ in Fig. 3 (a) can be determined by

$$\mathbf{D}\begin{pmatrix} \mathbf{p}_{i_1} \\ \mathbf{p}_{i_2} \end{pmatrix} = \mathbf{A}_i \tag{6}$$

where $\mathbf{D}$ is an ($N \times N$) DFT matrix whose (i,j)-th element is $d_{i,j} = e^{-j2\pi ij/N}$, $i,j = 0,1,\cdots,N-1$. Decomposing $\mathbf{D}$ into two matrices, $(N \times M)$ matrix $\mathbf{D}_1$ and $(N \times (N-M))$ matrix $\mathbf{D}_2$, we have

$$\left(\mathbf{D}_1 \mid \mathbf{D}_2\right)\begin{pmatrix} \mathbf{p}_{i_1} \\ \mathbf{p}_{i_2} \end{pmatrix} = \mathbf{D}_1\mathbf{p}_{i_1} + \mathbf{D}_2\mathbf{p}_{i_2} = \mathbf{A}_i \tag{7}$$

Thus $\mathbf{p}_{i_2}$ can be determined from

$$\mathbf{D}_2\mathbf{p}_{2_i} = \mathbf{A}_i - \mathbf{D}_1\mathbf{p}_{1_i} \tag{8}$$

Note that (8) is an overdetermined equation that can be solved using a weighted least-square (WLS) method [10]. We can control the amount of error in the "don't care" region (i.e., redundant signal) by adjusting a weight matrix $\mathbf{W}$. The weight matrix $\mathbf{W}$ is an ($N \times N$) diagonal matrix whose diagonal element w_j, $j = 0,1,\cdots,N-1$, has a value w_1 for the redundant signal $\mathbf{U}_i$ and w_2 for the short orthogonal code $\mathbf{C}_i$,

$$\mathbf{W} = \begin{pmatrix} w_1 & 0 & 0 & \cdots & 0 \\ 0 & w_2 & 0 & \cdots & 0 \\ 0 & 0 & w_2 & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \cdots & w_1 \end{pmatrix} \tag{9}$$

The power assigned to the "don't care" region can be controlled by adjusting w_1 and w_2. Let us define the weight ratio by

$$\gamma_w = \frac{w_1}{w_2} \tag{10}$$

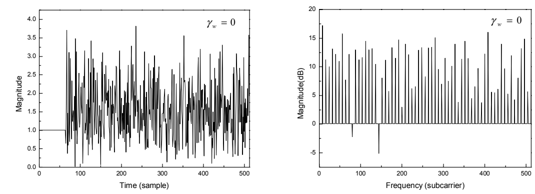
Form a modified equation for the WLS method represented as [10]

$$\mathbf{W}\mathbf{D}_2\mathbf{p}_{i_2} = \mathbf{W}(\mathbf{A}_i - \mathbf{D}_1\mathbf{p}_{i_1}) \tag{11}$$

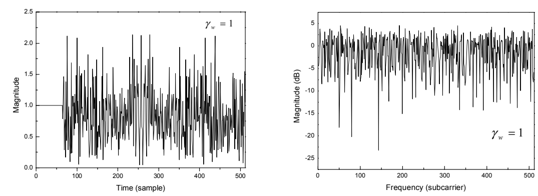
$\mathbf{p}_{i_2}$ can be determined by

$$\hat{\mathbf{p}}_{i_2} = \left(\mathbf{D}_2^{\mathrm{T}}\mathbf{W}^{\mathrm{T}}\mathbf{W}\mathbf{D}_2\right)^{-1}\mathbf{D}_2^{\mathrm{T}}\mathbf{W}^{\mathrm{T}}\mathbf{W}(\mathbf{A}_i - \mathbf{D}_1\mathbf{p}_{i_1}) \tag{12}$$

Fig. 4 depicts the time and frequency response of the proposed pilot signal according to γ_w when M=64, N=512 and L=64. In Fig. 4, first 64 samples in the time domain corresponds to the binary Gold code for timing synchronization. Assuming that total pilot signal power is constant, it can be seen that the redundant



(a) When $\gamma_w = 0$



(b) When $\gamma_w = 1$

Fig. 4. Time and frequency response of the proposed pilot signal

signal power decreases as γ_w increases. However, the interference power due to the corruption of orthogonality also increases as γ_w increases. Thus, it may be desirable to determine γ_w so as to maximize the signal to interference and noise power ratio (SINR) for channel estimation.

Let us define the frequency response of the pilot signal of the i-th user by

$$\begin{aligned} \mathbf{X}_\mathbf{i} &= \mathbf{D}\begin{pmatrix} \mathbf{p}_{i_1} \\ \hat{\mathbf{p}}_{i_2} \end{pmatrix} \\ &= \left(X_{u_i}[0], \mathbf{X}_{\mathbf{C}_i,0}, X_{u_i}[1], \mathbf{X}_{\mathbf{C}_i,1}, \cdots, \mathbf{X}_{\mathbf{C}_i,K-1}, X_{u_i}[M-1]\right) \end{aligned} \tag{13}$$

where $\mathbf{X}_{\mathbf{C}_i,k} = \left(X_{C_i,k}[0], X_{C_i,k}[1], \cdots, X_{C_i,k}[L-1]\right)$, $X_{u_i}[n] = U_i[n] + e_{u,i}[n]$, $n = 0,1,\cdots,M-1$, and $X_{c_i,k}[m] = C_i[m] + e_{c,i,k}[m]$, $m = 0,1,\cdots,L-1$. Here $\mathbf{X}_{\mathbf{C}_i,k}$ denotes the frequency response of the designed pilot signal at the k-th cluster, e_u is the difference between the designed pilot signal and desired one in the "don't care" region, and e_c is the difference between the designed pilot signal and desired one in the orthogonal code. For fair comparison, we normalize the power of the designed pilot signal, i.e.,

$$\frac{1}{N}\left|\alpha \mathbf{X}_i\right|^2 = 1 \tag{14}$$

where the normalization coefficient is determined as

$$\alpha = \sqrt{\frac{N}{\left|\mathbf{X}_i\right|^2}} \tag{15}$$

In the receiver, the channel coefficient $H_{i,k}$ of the i-th user can be obtained by despreading the received signal using the i-th user's orthogonal code $\mathbf{C}_i$. The received signal at the k-th cluster after FFT can be represented as

$$\mathbf{R}_k = \sum_{i=1}^{I-1} \alpha H_{i,k}\mathbf{X}_{C_i,k} + \mathbf{v} \tag{16}$$

where $\mathbf{v} = (v[0], v[1], \cdots, v[L-1])$ is zero mean additive Gaussian noise vector with variance σ_n^2 and $H_{i,k} = H_{i,k_I} + jH_{i,k_Q}$ is the channel coefficient of the i-th user at the k-th cluster. Note that H_{i,k_I} and H_{i,k_Q} are zero-mean Gaussian random variables with the same variance $\sigma^2/2$.

The despreaded signal can be represented as

$$\mathbf{R}_k\mathbf{C}_i = \alpha H_{i,k}\mathbf{X}_{C_i,k}\mathbf{C}_i + \sum_{j=1(j\neq i)}^{I-1}\left(\alpha H_{j,k}\mathbf{X}_{C_j,k}\mathbf{C}_i\right) + \mathbf{v}\mathbf{C}_i \qquad (17)$$

where $\alpha H_{i,k}\mathbf{X}_{C_i,k}\mathbf{C}_i$ is the channel information, $\sum_{j=1(j\neq i)}^{I-1}\left(\alpha H_{j,k}\mathbf{X}_{C_j,k}\mathbf{C}_i\right)$ is the sum of multi-user interference and $\mathbf{v}\mathbf{C}_i$ is the Gaussian noise term. Thus, the SINR is given by

$$\mathrm{SINR}_{\mathrm{CE}} = E\left\{\frac{\left|\alpha H_{i,k}\mathbf{X}_{C_i,k}\mathbf{C}_i\right|^2}{\left|\sum_{j=1(j\neq i)}^{I-1}\left(\alpha H_{j,k}\mathbf{X}_{C_j,k}\mathbf{C}_i\right) + \mathbf{v}\mathbf{C}_i\right|^2}\right\} \qquad (18)$$

Using the code orthogonality, it can be shown that

$$H_{j,k}\mathbf{X}_{C_j,k}\mathbf{C}_i = H_{j,k}\left(\sum_{l=0}^{L-1} C_i[l]e_{c,j,k}[l]\right) \qquad (19)$$

and

$$H_{i,k}\mathbf{X}_{C_i,k}\mathbf{C}_i = H_{i,k}\left(\sum_{l=0}^{L-1} C_i[l]^2 + \sum_{l=0}^{L-1} C_i[l]e_{c,i,k}[l]\right) \qquad (20)$$

Thus, (18) can be rewritten as

$$\mathrm{SINR}_{\mathrm{CE}} = E\left\{\frac{\left|H_{i,k}\left(\sum_{l=0}^{L-1} C_i[l]^2 + \sum_{l=0}^{L-1} C_i[l]e_{c,i,k}[l]\right)\right|^2}{\left|\sum_{\substack{j=0\\(j\neq i)}}^{I-1} H_{j,k}\left(\sum_{l=0}^{L-1} C_i[l]e_{c,j,k}[l]\right) + \frac{1}{\alpha}\sum_{l=0}^{L-1} C_i[l]v[l]\right|^2}\right\} \qquad (21)$$

Fig. 5 depicts $\mathrm{SINR}_{\mathrm{CE}}$ obtained using the Monte-Carlo method. It can be seen that $\mathrm{SINR}_{\mathrm{CE}}$ can be maximized by controlling γ_w for given number of active users, code length L, signal to noise power ratio (SNR) σ^2/σ_n^2. If γ_w is too small, it can cause SINR degradation due to large power of the redundant signal. On the other hand, if γ_w is too large, it can also cause SINR degradation due to the corruption of code orthogonality. It can be seen that the performance degradation becomes large as the number of active users increases. The increase of active users implies the increase of code interference power, reducing the SINR. Note that the peak-to-average power ratio (PAR) of the pilot signal varies depending upon γ_w. As an example, when M =64, N =512 and L =64, the PAR of the pilot signal decreases from 8.0 to 7.0 as γ_w increase for 0 to 1.0. Thus, it may be desirable to appropriately determine γ_w considering the PAR and SINR in a compromising manner.

Table 1. Simulation parameters

Data modulation	QPSK (MC-CDMA)
Bandwidth	64 MHz
Subcarrier number (N)	512
Time sample interval (T)	15.6 ns
Guard interval length	128 sample
Length of the frequency domain orthogonal Gold code (L)	64
Length of the time domain Gold code (M)	64
Weight ratio (γ_w)	0.2

Table 2. Comparison of the computational complexity

Operation	IDFT signal	Proposed signal
Multiplication	$4N_D M$	None
Addition	$N_D(4M-2)$	$2N_D(M-1)$

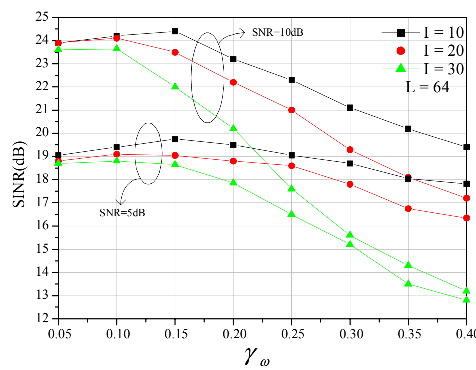
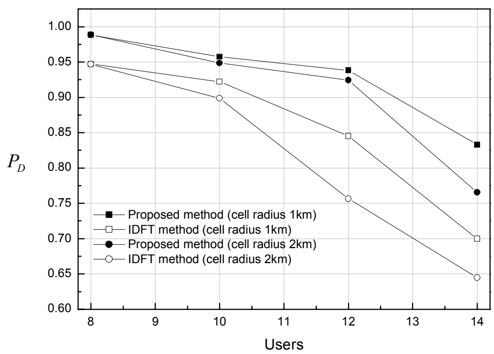


Fig. 5. $\mathrm{SINR}_{\mathrm{CE}}$ **according to** γ_w

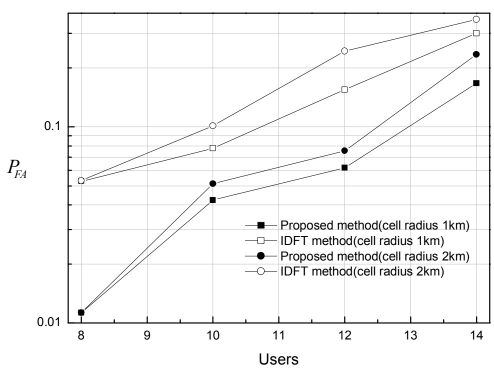
IV. PERFORMANCE EVALUATION

To verify the performance of the proposed pilot scheme, we evaluate the acquisition performance and computational complexity. For fair comparison, we use the same MC-CDMA parameters used in [11, 12]. Table 1 summarizes the simulation parameters.

Fig. 6 depicts the detection performance as a function of I in AWGN with Eb/No = 12dB, λ = -3dB (when we assume that the peak of code autocorrelation function is 0dB). The detection window length N_D is determined by (1). We assume that N_D =400 (for R = 1 km) and 800 (for R = 2km). It can be seen that the detection probability decreases and the false alarm probability increases as the number of active users increases. Since the partial autocorrelation property of the IDFT signal is not good as much as conventional Gold codes as shown in Fig. 7, the use of IDFT based pilot signal has detection performance inferior to the use of the



(a) Detection probability



(b) False detection probability

Fig. 6. Detection performance

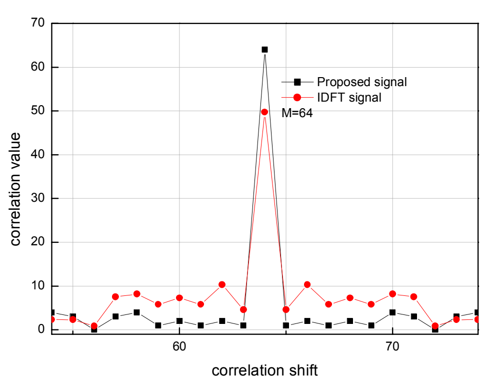


Fig. 7. Autocorrelation function of pilot signals

proposed one.

Since the IDFT pilot signal of an orthogonal code is a noise-like complex signal, it requires a large computational complexity for symbol timing detection. On the other hand, the proposed pilot signal uses real binary Gold codes, requiring no multiplication for the correlation. Table 2 compares the implementation complexity, where M is the length of the inserted time domain code. It can be seen that the detector complexity can be reduced significantly by using the proposed pilot signal.

V. CONCLUSION

In this paper, we have proposed a novel method for the design of pilot signal by using a WLS method. The designed pilot signal can be used for both synchronization and channel estimation. The use of the proposed pilot signal can provide good symbol timing acquisition performance compared to the IDFT signal based scheme, while reducing the implementation complexity.

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