Adaptive Noncoherent Receivers for MC-CDMA Communications

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Abstract—In this paper, we propose differential phase-shift keying (DPSK) noncoherent receivers for multicarrier code division multiple access (MC-CDMA) systems in multipath channels. The noncoherent receivers are composed of a linear equalizer and a decision-feedback differential detector to detect DPSK signals. The performances of the proposed noncoherent receivers can be improved by increasing the number of feedback symbols. For an infinite number of feedback symbols, the performances of the proposed noncoherent receivers approaches that of the conventional coherent receiver. Some simulation examples are given to show the system performances of the four proposed receivers.

Keywords—MC-CDMA, DF-DD, noncoherent receiver

1. INTRODUCTION

Recently, multicarrier multiple access schemes based on the combination of CDMA schemes and orthogonal frequency division multiplexing (OFDM) signalling, which are referred to as MC-CDMA systems, have attracted wide attention in the field of wireless communications. The number of carriers is typically chosen to be large enough so that the signal on each subcarrier is propagated through a channel which behaves in a nonselective manner. The system realizes a diversity gain due to the fact that the fading processes on each carrier are independent, with sufficient separation of the carriers [1].

In MC-CDMA wireless communication systems, the system performance is limited by multiple-access interference (MAI) and intersymbol interference (ISI). In order to solve the problem of MAI and ISI in MC-CDMA multiuser systems, multiuser detection which requires the full knowledge of parameters of MAI can be more computationally complex compared with that of single-user detection [2]. Besides, coherent receivers require exact knowledge of the channel phase for optimum performance. Thus, complex carrier phase and frequency synchronization circuits have to be implemented in the receiver. However, in fading environments or when low-cost local oscillators are employed, acquisition and tracking of the carrier phase may be difficult or even impossible [3]. Noncoherent single-user receiver is a favorable choice.

There only exists few literatures to discuss MC-CDMA systems with differential encoding [4]-[8]. The DPSK is performed at a subcarrier level to cause significant performance gain for downlink MC-CDMA in [4].

It also uses a joint phase estimation approach for all users’ signals to allow the implementation of an effective equal gain combining receiver. In [5], an uplink MC-CDMA system based on chip-level differential encoding was proposed. This scheme can simplify the hardware structure and the channel estimation algorithms as well as to reduce the sensitivity to possible phase noise in the frequency down conversion circuit. In [6], an MC-CDMA system using symbol-level differential encoding was proposed and analyzed in frequency selective static and fast Raleigh fading channels. The closed-form expressions for the probability of error are also derived and evaluated. In [7], a less complex system by integrating noncoherent differential amplitude and phase shift keying modulation with multi-carrier code division multiple access (DAPSK-MC-CDMA) was introduced. The DAPSK-MC-CDMA system can reduce the system complexity and also maintain a good performance as well as high spectral efficiency. In [8], the authors also adopt DAPSK modulation in MC-CDMA systems. It was found that the 16/64 DAPSK-MC-CDMA scheme can significantly improve the system performance and is very robust to frequency offset compared with that of coherent modulation. In this paper, we propose four adaptive single-user MC-CDMA receivers with decision feedback differential detection (DF-DD) in phase unknown environment [9]. The proposed noncoherent receivers are composed of a linear equalizer and either a nonrecursive or a recursive DF-DD scheme. We show that the performances of the proposed receivers with DPSK signal can approach that of the coherent receiver with PSK signal if an infinite number of feedback symbols is used. For adaptation of the equalizer coefficients, a modified LMS algorithm with low complexity and fast convergence speed is used. Computer simulations confirm that even for a finite number of feedback symbols, the proposed receivers perform considerably better than a comparable scheme. Three kinds of COST 207 multipath channel are considered: rural area, typical urban and bad urban [10].

II. SYSTEM MODEL

Consider a downlink M-ary DPSK MC-CDMA system in multipath fading channels. Transmitted data bits are differentially encoded before being serial-to-parallel converted to a number of parallel streams. On each stream each encoded bit is DS spread spectrum modulated and transmitted with certain number of carriers. Successive carriers of the system are allowed to overlap with various overlapping percentages. The baseband structure of the DPSK
MC-CDMA transmitter is presented in Fig. 1. Assume that there are $U$ active users accessing this MC system simultaneously. At the $k$th MC block symbol interval (of duration $T$), the $M$-ary DPSK signals $a(u)[k] \in \{e^{\jmath 2\pi n/M} | n \in \{0,1,\ldots,M-1\}\}$ of the $u$th user are differentially encoded. The PSK signal $b^{(u)}[k]$ results from

$$b^{(u)}[k] = a^{(u)}[k] y^{(u)}[k-1]$$

(1)

The PSK symbol is then spread by code vector $c^{(u)} = [c_0^{(u)}, c_1^{(u)}, \ldots, c_{Q-1}^{(u)}]$ of length $Q$, where $[\bullet]^T$ denotes the transposition of a vector. All the $Q$ spreading chips corresponding to one particular symbol are serial-to-parallel converted to maintain a longer symbol duration. Prior to the orthogonal frequency-division multiplexing (OFDM) modulation, the spreading signal from each user must be combined firstly. The complex bandpass representation of the transmitted signal without guard interval is thus given as

$$s(t) = \sum_{q=0}^{Q-1} \sum_{k=-\infty}^{T-1} b^{(u)}[k] y^{(u)}_q e^{\jmath 2\pi f_q t + q_\delta} p(t-kT)$$

where $f$ is the carrier frequency, $f_\delta$ is the subcarrier separation and $p(t)$ is a rectangular signalling pulse. Since the OFDM modulation can be implemented by taking an inverse fast Fourier transform (IFFT), the matrix form of the $k$th OFDM block symbol for the $u$th user can be expressed as follows,

$$s[k] = F^H C h[k] = [s_0[k], s_1[k], \ldots, s_{Q-1}[k]]^T$$

where $F$ indicates a Fourier matrix with entries given by $e^{-\jmath 2\pi k q/Q}$ and $[\bullet]^H$ denotes the Hermition transposition of a vector. The code matrix $C$ is composed of the spreading codes of all active users, where the $u$th column of $C$ is equal to $c^{(u)}$. The data vector $b[k]$ is composed of PSK symbol of all active users within the $k$th OFDM symbol block, where the $u$th element of $b[k]$ is $b^{(u)}[k]$. A guard interval in the form of a cyclic prefix (CP) is added in front of the CP converted samples of the OFDM block to avoid interference between consecutive OFDM symbols [11]. The transmission channel is modeled with $L$ Raleigh fading paths, each of which is characterized by path gain factor $h_l$ and relative delay time $\tau_l$. The channel impulse response can be characterized by $h(t) = \sum_{l=0}^{L-1} h_l \delta(t-\tau_l)$. Assume that the entries of $h_l$ are i.i.d. (independent and identically distributed) Gaussian random variables with unit variance.

III. The Proposed Noncoherent MC-CDMA Receivers

If the guard interval is longer than the multipath spread and the synchronization is perfect, there is no ISI between any two successive OFDM blocks. Thus we just need to concentrate on one OFDM block after removing the CP. For simplicity of notation, we ignore the section of the CP contained in the OFDM block, keeping in mind that the CP block still exists. The use of CP converts the linear convolution of the transmitted data block and the channel impulse response to a circular convolution [12]. Therefore, the received signal after removing CP in the time-domain can be described as

$$r[k] = e^{\jmath \theta} \sum_{l=0}^{L-1} h_l s[k;l] + g[k]$$

(2)

$$s[k;l] = \left[ s_{Q-l}[k], \ldots, s_{Q-1}[k], s_0[k], \ldots, s_{Q-(l+1)}[k] \right]^T$$

where $\theta$ is the unknown constant phase shift in noncoherent case, perfect constant phase shift in coherent case. $s[k;l]$ denotes the signal of the $l$th path delayed $l$ sampling time, and $g[k]$ is the additive noise vector containing zero-mean, uncorrelated complex Gaussian noise (AWGN) samples during the $k$th MC block interval. The delayed path is equivalent to the cyclic shift of the direct path because the guard interval in the form of CP is appended in front of the data block. From the expression (2), the distorted signal through a dispersive channel can be viewed as a linear combination of contents of all the attenuated paths. After discarding the samples corresponding to the CP, an FFT of size $Q$ is performed, i.e. $y[k]=Frw[k]$. The basic concept of MC-CDMA differential detector is shown in Fig. 2 where $q[k]$ is the symbol rate output of the proposed adaptive receiver. In the next stage of the receiver, the decision variable $d[k]$ is obtained by noncoherent processing of $q[k]$. For this purpose, the reference symbol $q_{ref}[k]$ is generated. It can be done nonrecursively as follows,

$$q_{ref}[k-1] = \frac{1}{N-1} \sum_{n=1}^{N-1} q[k-n] \prod_{z=1}^{n} \hat{a}[k-z]$$

(3)

where $N\geq 2$, is the number of adaptive filter output symbols used for calculation of $d[k]$ and "cleaning" the reference symbol. This concept will be explained in detail later. The recursive form of the reference symbol can be obtained as

$$q_{ref}[k-1] = (1-\alpha)q[k-1] + \alpha \hat{a}[k-1] q_{ref}[k-2]$$

(4)

where $0 \leq \alpha \leq 1$ is a forgetting factor. The decision variable $d[k]$ is then obtained from

$$d[k] = q[k] q_{ref}[k-1]$$

(5)

where $(\bullet)^*$ denotes as complex conjugation. $\hat{a}[k]$ in (3) and (4) can be the estimated DPSK symbols based on hard decision of $d[k]$, that is $\hat{a}[k] = \text{sgn}(d[k])$. By LMS algorithm, the equalizer coefficients can be adjusted to minimize the error variance $\sigma_e^2 = E\{e[k]^2\}$, where the error signal $e[k]$ is given as

$$e[k] = \hat{a}[k] - d[k]$$

(6)

For convenience, we assume that $\alpha=0$ is the desired user, and $c = e^{\jmath \theta}$ used to despread the signal in the receiver.

A. Q-tap Receiver

The received signals first go through FFT followed by a transversal $Q$-tap filter. The equalizer output of the $k$th symbol time signal is given as

$$q[k] = w^H[k] Frw[k]$$

(7)
where $w[k]$ is the weight of adaptive filter with length $Q$ which is initialized as the spreading code of the desired user. This filter not only despreads the received signal but also equalizes the MAI and channel effect. Substituting (7) into (5), the error signal of (6) can be written as follows,

$$ e[k] = \tilde{a}[k] - w^{H}[k]\mathbf{F}[k]\mathbf{g}_{ref}[k-1] $$

(8)

Applying the steepest descent method, the instantaneous gradient vector can be obtained as follows,

$$ \frac{\partial}{\partial w}[k]e[k] = -\mathbf{F}[k]\mathbf{g}_{ref}^*[k-1]\mu^*[k] $$

So the adaptive filter weight can be updated as

$$ w[k+1] = w[k] + \mu e^*[k]\mathbf{F}[k]\mathbf{g}_{ref}^*[k-1] $$

where $\mu$ is the step size.

**B. CHRA Receiver**

Let the FFT output $\mathbf{F}[k]$ as

$$ \hat{r}[k] = \mathbf{F}[k] = [\hat{r}_0[k], \hat{r}_1[k], ..., \hat{r}_{Q-1}[k]]^T $$

and define a $G \times Q$ matrix as follows,

$$ \mathbf{R}[k] = \begin{bmatrix} \hat{r}_0[k] & \cdots & \hat{r}_{Q-1}[k] \\ \hat{r}_1[k] & \cdots & \hat{r}_{Q-1}[k] \\ \vdots & \cdots & \vdots \\ \hat{r}_{Q-1-G}[k] & \cdots & \hat{r}_{Q-1-G}[k] \end{bmatrix} $$

where $G$ is the number of filter taps and $G \leq Q$ in general. Note that $\hat{r}_x[k] = \hat{r}_{Q-x}[k-1]$ when $x \in \{1, ..., Q\}$. The received matrix is first equalized in chip level and then despreaded at the adaptive filter output. Contrary to the symbol-level estimation in the $Q$-tap receiver, CHRA is a chip-level signal estimation method. The chip-level equalizer output is $\hat{q}[k] = w^{H}[k]\mathbf{R}[k]$, where $w[k]$ is a $G \times 1$ weight vector.

C. **CSFB Receiver**

The first transversal filter in the nonadaptive filter bank is the code matched filter of the desired user, and its output is the despreaded signal. The other $D-1$ filters are cyclically shifted versions of the code matched filter with successive shifts being spaced by one sampling time. The shifted version filter not only operate the function of despread but also match to the correlation results of the transmitted signal and channel. To estimate the desired signal, we can combine the filter outputs of the nonadaptive filters linearly and adaptively as follows,

$$ q[k] = w^{H}[k]r_{ref}[k] $$

(11)

where $w[k]$ is a $D \times 1$ vector. The elements of $r_{ref}[k]$ can be described as follows,

$$ r_{ref1} = c_1^{(t)}[k] $$

$$ r_{ref2} = c_2^{(t)}[k] $$

$$ r_{refD} = c_D^{(t)}[k] $$

Note that $c_1$ is the first code matched filter of the nonadaptive filter bank which consists of the spreading code of the desired user. The other code matched filters, $c_2, ..., c_D$, are cyclically version of the spreading code of the desired user. Substituting (11) into (5), the error signal of (6) can be obtained as follows,

$$ e[k] = \hat{a}[k] - w^{H}[k]r_{ref}[k-1] $$

The instantaneous gradient vector can be formulated as

$$ \frac{\partial}{\partial w}[k]e[k] = -r_{ref}[k]g_{ref}^*[k-1] $$

The adaptive filter weight thus can be updated as

$$ w[k+1] = w[k] + \mu e^*[k]r_{ref}[k-1] $$

**D. DFAR Receiver**

The DFAR uses both forward filter and feedback filter. The filter output of the forward filter is written as $q_{1}[k] = w_{1}^{H}[k]\mathbf{R}[k]$, where $w_{1}[k]$ is a $G \times 1$ weight vector of the forward filter. The output signal $q_{1}[k]$ can be described as $q_{1}[k] = q_{1}[k]c$. Before differential decoding, the estimated symbol is obtained by subtracting the decision feedback interference from the output signal, that is

$$ q[k] = q_{1}[k] - \hat{d}[k] $$

(12)

where $\hat{d}[k]$ is the output signal of the feedback filter,

$$ \hat{d}[k] = w_{b}^{H}[k]q_{b}[k] $$

(13)

and $w_{b}[k]$ is a $G_b \times 1$ weight vector of the feedback filter. The vector $q_{b}[k]$ consists of the reference signals already known to the receiver as follows,

$$ q_{b}[k] = [g_{ref}[k-G_b], ..., g_{ref}[k-1]] $$

Since the reference symbols are not differentially decoded, the time delay reference signal can be applied to estimate the interference in the feedback filter. Substituting (12) and (13) into (5), the error signal of (6) can be written as

$$ e[k] = \hat{a}[k] - w^{H}[k]\mathbf{R}[k]g_{ref}^*[k-1] + w_{b}^{H}[k]q_{b}[k]g_{ref}^*[k-1] $$

The instantaneous gradient vector of the forward filter can be formulated as follows,

$$ \frac{\partial}{\partial w}[k]e[k] = -e^*[k]\mathbf{R}[k]g_{ref}^*[k-1] $$

The instantaneous gradient vector of the feedback filter can be formulated as
\[
\frac{\partial}{\partial w(k)} \| e(k) \|^2 = e^\ast[k] R[k] e^\ast_{ref}[k-1]
\]

The updated equations of the forward filter and the feedback filter can be described as

\[
w_f[k+1] = w_f[k] + \mu e^\ast[k] R[k] e^\ast_{ref}[k-1]
\]
\[
w_b[k+1] = w_b[k] - \mu e^\ast[k] q_b[k] e^\ast_{ref}[k-1]
\]

IV. SIMULATION RESULTS

In the simulations, the 32 Hadamard spreading code is selected. The modulated signals are DQPSK for the noncoherent detection and QPSK for the coherent detection throughout the simulations. The step size \( \mu = 0.005 \). The phase shift \( \theta \) is randomly generated between 0 and \( 2\pi \) in each experiment, unknown in the noncoherent receiver, perfect known in the coherent case. The filter lengths of the \( Q \)-tap, the CHRA and the CSFB receivers are 32, 16 and 16, respectively. Besides, the filter lengths of the forward filter and the feedback filter in DFAR are both 16. The simulation channels are COST 207 environments including typical urban, bad urban and rural area.

Fig. 3 shows the learning curves of the proposed adaptive noncoherent \( Q \)-tap, CHRA, CSFB and DFAR receivers in nonrecursive mode with \( N=25 \). Obviously, the \( Q \)-tap receiver has the best MMSE performance but worst convergence rate. This is because the filter length of the \( Q \)-tap receiver is larger than that of the others. In general, larger filter length can be more effective in compensation for interference. Besides, larger filter length cause larger initial error, the receiver thus spends more time on convergence.

The BER performance comparisons of the noncoherent \( Q \)-tap receiver with nonrecursive mode and the coherent one are shown in Fig. 4. We can observe that the BER of the noncoherent method decreases considerably as \( N \) increases. This might be attributed to the fact that the reference symbol \( q_{\text{ref}}[k-1] \) is less noisy for larger \( N \) (the ensemble average effect in (3)). There is about 2 dB loss between noncoherent receiver with \( N=3 \) and the coherent one at BER=10^-2. However, the power loss reduces to about 0.5 dB for \( N=8 \). The BER performance of the proposed noncoherent receiver is almost the same as that of the coherent one when the number of feedback symbol is 25. Thus, in the following simulations, we consider \( N=25 \) for all the proposed receivers in nonrecursive mode.

From Figs. 5 to 6, we show the BER comparisons of all the proposed receivers with Hadamard codes. The BER improvement between the DFAR receiver and the CHRA receiver is significant in bad urban, since this environment has richer multipath interference than the other two. Finally, Fig. 7 shows the BERs versus the number of users \( U \) with Hadamard codes in typical urban. From this figure, we can see that when the system is half-loaded, the \( Q \)-tap receiver and the DFAR receiver still can work well (BER < 10^-7). Besides, when the number of users is increased, the BER degradation of the CSFB receiver is more serious than the others.

V. CONCLUSIONS

In this paper, we propose four adaptive single-user MC-CDMA receivers with DF-DD in phase unknown environment. The noncoherent receivers are composed of a linear equalizer and either a nonrecursive or a recursive DF-DD scheme. The performances of the proposed receivers with MDPSK signal can approach that of the coherent receiver with MPSK signal if an infinite number of feedback symbols is used. Computer simulations confirm that even for a finite number of feedback symbols, the proposed receivers perform well.

REFERENCES


Fig. 1. The block diagram of MC-CDMA transmitter with differential encoder.
Fig. 2. The basic form of MC-CDMA differential detector

Fig. 3. Learning curves of four receivers with Hadamard codes in bad urban.

Fig. 4. BER comparison of noncoherent and coherent Q-tap receivers.

Fig. 5. BER performances of the four noncoherent receivers with Hadamard codes in typical urban.

Fig. 6. BER performances of the four noncoherent receivers with Hadamard codes in bad urban.

Fig. 7. BER performances versus number of users in typical urban channel.