Proceedings of

The 10th Annual Joint Workshop

On

Control and Power Electronics Technology

The School of Electrical Engineering and Automation
Harbin Institute of Technology, China

And

The School of Electrical Engineering and Computer Science
Kyungpook National University, Korea

Sponsored by

The School of Electrical Engineering and Automation,
Harbin Institute of Technology, China

Co Sponsored by

The School of Electrical Engineering and Computer Science
Kyungpook National University, Korea

July 14, 2009
Harbin, Heilongjiang, China
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Workshop Program

14 July 2009 Room 604, Yufushao Science Hall

Chairman: Prof. Dianguo Xu (Harbin Institute of Technology, China)

Prof. Cho, Jinho (Kyungpook National University, Korea)

9:00-9:10 Opening Ceremony

Chairman: Dr. Xue Sun (Harbin Institute of Technology, China)

No.1 Workshop

Chairman: Prof. Cho, Jinho (Kyungpook National University, Korea)

9:10-9:35

Minyoung Kim, Autonomous Map-Building for Mobile Robots Using A Visual Perception Guidance, KNU, Korea

9:35-10:00

Kai Song, ICA Based Pattern Analysis for Electronic Noses, HIT, China

10:00-10:25

Sekwang Park, Glass Properties for Electrostatic Bonding in Process of the Packaging of MEMS Devices, KNU, Korea

10:25-10:40

Coffee Break

No. 2 Workshop

Chairman: Prof. Xiaohua Zhang (Harbin Institute of Technology, China)

10:40-11:05

Mingcheng Zhang, Analysis and Optimization of Thrust Characteristics
of Tubular Linear Electromagnetic Launcher for Space-Use, HIT, China

11:05-11:30

Vyacheslav Tuzlukov, Performance of DS-CDMA Downlink Systems with Fading Channel under Employment of Generalized Detector, KNU, Korea

11:30-11:55

Haiming Qi, Research of Pipeline Robot Tracking and Locating Technology

HIT, China

12:00-12:15

Taking photos in front of Dianji building

Closing Ceremony

12:15

Lunch in Xiyuan Hotel
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Performance of DS-CDMA Downlink Systems with Fading Channel under Employment of Generalized Detector

Vyacheslav P. Tuzlukov

Abstract: This paper investigates a transmitted signaling technique using orthogonal unified complex Hadamard transform spreading sequences and the generalized detector constructed on the basis of the generalized approach to signal processing in noise direct-sequence code-division multiple access (DS-CDMA) downlinks to maintain the orthogonality between users and reduce the effect of multipath fading and interference from other users. A general multipath fading model is assumed. System performance is evaluated by means of signal-to-interference-plus-noise ratio (SINR) at the generalized receiver output. It is shown that the SINR of the system employing the orthogonal unified complex Hadamard transform spreading sequences and the generalized receiver is independent of the phase offsets between different paths, while the SINR of the same system using the Walsh-Hadamard (WH) spreading sequences is related to the squared cosine of path phase offsets. As a result, the bit-error ratio (BER) performance of the DS-CDMA downlink systems employing the generalized receiver is better than that of the system with the WH spreading sequences at high SINRs. Comparative analysis between the BER performance of DS-CDMA system using the generalized receiver and Rake one, which consists of a bank of correlation receivers, with each individual receiver correlating with a different arriving multipath component, shows a superiority of the first receiver over the second one both at high SINRs and at low SINRs.

Index Terms: Direct-sequence code-division multiple access (DS-CDMA), generalized detector (GD), frequency-selective fading, signal-to-interference-plus-noise ratio (SINR), unified complex Hadamard transform spreading sequences, Walsh-Hadamard spreading sequences (WH).

1. INTRODUCTION

DIRECT-SEQUENCE code-division multiple access (DS-CDMA) is a well-known communication transmission technique that allows multiple subscribers to share the same spectrum simultaneously. It provides spectrum efficiency, high system capacity, robustness against interference, and improved quality of service [1], [2]. Especially, in DS-CDMA downlink systems, signals of multiple users are transmitted simultaneously by means of orthogonal spreading codes generated by con-catena-tion of orthogonal Walsh-Hadamard (WH) channelization sequences and pseudo-noise (PN) random scrambling sequences. In a cellular environment however, the presence of multi-paths introduces non-zero time delays that destroy orthogonality between codes, and leads to interference among the signals in the downlink. The scrambling sequences aim to randomized signals? unsatisfactory and inhomogeneous behavior at non-zero time delays. The performance of the concatenated WH/PN spreading sequences was analyzed in [3].

Recently, wideband DS-CDMA has become the focus of current research interest, for example, for future mobile communication systems. The use of complex spreading sequences is an important feature of wideband CDMA. Complex spreading is a term used in reference to dual-channel spreading [4]. It can be implemented either by a complex-valued sequence [5][7] or by two binary sequences. As was shown in [5], the complex sequences can be 3 dB better than binary Gold sequences in the maximum periodic correlation parameters, and in addition to that, larger sets are available in complex sequences. All the complex sequences mentioned above, however, are nonorthogonal. They can be categorized as complex-valued PN spreading sequences. The application of nonorthogonal complex sequences in DS-CDMA was investigated in [8], where performance bounds were derived for DS-CDMA systems with complex signature sequences over additive white Gaussian noise (AWGN) channels.

In this paper, we investigate the generalized receiver constructed on the basis of the generalized approach to signal processing in noise [9][13] in DS-CDMA downlinks to maintain the orthogonality between users and reduce the effect of multi-path fading and interference from other users. We consider the orthogonal 4-phase complex sequences for DS-CDMA down-link system. The sequences are derived from the unified complex Hadamard transform matrix introduced in [14]. The correlation properties of unified complex Hadamard transform sequences have been studied in [15]. The unified complex Hadamard transform sequences provide better autocorrelation pro-perties than WH sequences that are characterized with very poor autocorrelation properties.

V. P. Tuzlukov is with the School of Electrical Engineering and Computer Science, Kyungpook National University, Daegu, Korea (phone: 82-53-950-5599; fax: 82-53-950-5505; e-mail: tuzlukov@ee.knu.ac.kr).
The orthogonal unified complex Hadamard transform spreading sequences are used at the transmitter as channelization spreading codes scrambled by a long PN sequence to maintain the orthogonality between the users, and at the same time, reduce the effect of multipath fading and interference from other users. A coherent generalized receiver [10] is used to combat the adverse effects of short-term multipath fading in mobile radio propagation environments. Because the signal-to-interference-plus-noise ratio (SINR) expression is computationally much simpler compared with the error-probability expression, the SINR is mostly used for evaluating and selecting code sequences among the several candidates. Therefore, in this paper, we mainly investigate the SINR at the generalized receiver output when the DS-CDMA downlink transmitter employs the unified complex Hadamard transform spreading sequences and compare this with the SINR when the WH real sequences are used. It is shown that the SINR of the system employing the generalized receiver and using the unified complex Hadamard transform spreading sequences is independent of the phase offsets between different paths, while the SINR of the same system using the WH sequences is related to the squared cosine of path phase offsets. Therefore, as a direct result, the BER performance of the DS-CDMA downlink system employing the unified complex Hadamard transform spreading sequences is better than that of the system with the WH sequences under Gaussian approximation. Also, we carry out a BER performance comparison for the DS-CDMA system employing the generalized receiver with the same system using the Rake receiver [16]. Comparative analysis shows us a great superiority in BER performance under employment of the generalized receiver.

II. HADAMARD TRANSFORM SEQUENCES

The considered DS-CDMA system uses the orthogonal unified complex Hadamard transform spreading sequences. These so-called orthogonal unified complex Hadamard transform sequences are easy to generate. Larger sets of complex sequences are also available. They were categorized into two groups: the half-spectrum property orthogonal unified complex Hadamard transform spreading sequences and the non-half-spectrum property orthogonal unified complex Hadamard transform spreading sequences. Consider briefly how these sequences can be generated and note the main definitions and remarks discussed in [14] and [15].

A unified complex Hadamard transform matrix $U$ of order $N = 2^n$ is a square matrix with elements $\pm 1$ and $\pm j$, and is constructed by [14], [15]:

$$U_n = \bigotimes_{i=1}^{n} U_1 = U_1 \otimes U_{n-1} = U_1 \otimes \cdots \otimes U_1,$$  
(1)

where $\otimes$ denotes the Kronecker product, and $U_1$ is defined as

$$U_1 = \begin{bmatrix} \mu_1 & \mu_2 \mu_3 \\ \mu_2 & \mu_2 \mu_3 \end{bmatrix}$$  
(2)

with $\mu_1, \mu_2, \mu_3 \in \{1, -1, j, -j\}$ and $j = \sqrt{-1}$.  
(3)

There are 64 different matrices for $U_1$ satisfying (2) with elements $\pm 1$ and $\pm j$,

$$U_1 U_1^* = U_1^* U_1 = 2I_2$$  
(4)

and

$$|\det(U_1)|^2 = 2^2,$$  
(5)

where $\ast$ indicates complex conjugate. Hence, the unified complex Hadamard transform matrix is orthogonal. Furthermore, the unified complex Hadamard transform matrices contain a WH transform matrix as a special case, with

$$\mu_1 = \mu_2 = \mu_3 = 1$$  
(6)

in matrix $U_1$.

The unified complex Hadamard transform matrices have two categories of 32 basic matrices, depending on whether $\mu_3$ in (2) is imaginary or not [15]. If $\mu_3$ is imaginary, the matrix group is called the half-spectrum property unified complex Hadamard transform. Otherwise, the group is called the non-half-spectrum property unified complex Hadamard transform. The unified complex Hadamard transform spreading sequence $c_k$, $k = 1, \ldots, N$, is defined by the $k$-th row of the unified complex Hadamard transform matrix.

It has been shown in [15] that the non-half-spectrum property unified complex Hadamard transform sequences have very similarly poor autocorrelation properties as WH sequences, and some of the half-spectrum property unified complex Hadamard transform sequences exhibit a reasonable compromise between the autocorrelation and cross-correlation functions. In this paper, we just consider the half-spectrum property unified complex Hadamard transform sequences, i.e., $\mu_3 = j$ or $-j$.

III. SYSTEM MODEL

In this paper, we consider a single-cell environment system model. This is sufficient to assess the effects of multipath fading and interference components from other users on the performance. In particular, we analyze a complex baseband-equivalent model with binary phase-shift keying (BPSK) data and complex sequence sequences over a multipath fading channel for a DS-CDMA downlink communication system. The baseband representation of the total signal transmitted on the downlink is given by

$$a(t) = \sum_{k=0}^{K-1} a_k(t) = \sum_{k=0}^{K-1} \sqrt{E_b} b_k(t) c_k(t),$$  
(7)

where $K$ is the number of users;

$$a_k(t) = \sqrt{P_b} b_k(t) c_k(t)$$  
(8)

is the transmitted signal of the $k$-th user; $P_b$ is the power of the $k$-th transmitted signal;
The orthogonal unified complex Hadamard transform spreading sequences are used at the transmitter as channelization spreading codes scrambled by a long PN sequence to maintain the orthogonality between the users, and at the same time, reduce the effect of multipath fading and interference from other users. A coherent generalized receiver [10] is used to combat the adverse effects of short-term multipath fading in mobile radio propagation environments. Because the signal-to-interference-plus-noise ratio (SINR) expression is computationally much simpler compared with the error-probability expression, the SINR is mostly used for evaluating and selecting code sequences among the several candidates. Therefore, in this paper, we mainly investigate the SINR at the generalized receiver output when the DS-CDMA downlink transmitter employs the unified complex Hadamard transform spreading sequences and compare this with the SINR when the WH real sequences are used. It is shown that the SINR of the system employing the generalized receiver and using the unified complex Hadamard transform spreading sequences is independent of the phase off-sets between different paths, while the SINR of the system using the WH sequences is related to the squared cosine of path phase offsets. Therefore, as a direct result, the BER performance of the DS-CDMA downlink system employing the unified complex Hadamard transform spreading sequences is better than that of the system with the WH sequences under Gaussian approximation. Also, we carry out a BER performance comparison for the DS-CDMA system employing the generalized receiver with the same system using the Rake receiver [16]. Comparative analysis shows us a great superiority in BER performance under employment of the generalized receiver.

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A unified complex Hadamard transform matrix \( U \) of order \( N = 2^s \) is a square matrix with elements \( \pm 1 \) at \( \pm j \), and is constructed by [14], [15]

\[
U_n = \bigotimes_{i=1}^{s} U_i = U_1 \otimes U_2 \otimes \cdots \otimes U_s ,
\]

where \( \otimes \) denotes the Kronecker product, and \( U_i \) is defined as

\[
U_i = \begin{bmatrix} \mu_1 - \mu_2 - \mu_3 \\ \mu_1 + \mu_2 - \mu_3 \\ \mu_1 - \mu_2 + \mu_3 \\ \mu_1 + \mu_2 + \mu_3 \end{bmatrix}
\]

with \( \mu_1, \mu_2, \mu_3 \in \{1, -1, j, -j\} \) and \( j = \sqrt{-1} \).

There are 64 different matrices for \( U_i \) satisfying (2) with elements \( \pm 1 \) and \( \pm j \),

\[
U_i U_i^* = U_i^* U_i = 2I_2
\]

and

\[
| \det (U_i) |^2 = 2^s
\]

where \( * \) indicates complex conjugate. Hence, the unified complex Hadamard transform matrix is orthogonal. Furthermore, the unified complex Hadamard transform matrices contain a WH transform matrix as a special case, with

\[
\mu_1 = \mu_2 = \mu_3 = 1
\]

in matrix \( U_1 \).

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\[
a(t) = \sum_{k=0}^{K-1} a_k(t) \sum_{k=0}^{K-1} b_k(t) c_k(t)
\]

where \( K \) is the number of users;

\[
a_k(t) = \left( P_{k} b_k(t) c_k(t) \right)
\]

is the transmitted signal of the \( k \)-th user; \( P_{k} \) is the power of the \( k \)-th transmitted signal;
\[ b_k(t) = \sum_{n=-\infty}^{\infty} b_k^{(n)} p_k(t-nT) \]  
(9)
is the data signal of the k-th user;

\[ b_k^{(n)} \in \{-1,+1\} \]  
(10)
denotes the n-th data bit value of the k-th user; the function \( p_k(t) \) is the rectangular pulse of symbol duration \( T \); \( c_k(t) \) is the complex spreading signal defined by

\[ c_k(t) = \sum_{m=-\infty}^{\infty} c_k^{(m)} \phi(t-mT_c) \]  
(11)
and \( c_k^{(m)} \) denotes the m-th complex chip value of the k-th user.
The function \( \phi(t) \) is a chip waveform that is time-limited to \([0,T_c] \) with

\[ \int_0^{T_c} \phi^2(t)dt = T_c \]  
(12)
including the rectangular pulse of duration \( T_c \), and \( T_c \) is called the chip duration. Throughout this paper, we assume that \( T = NT_c \).

Power control is assumed to be perfect, and the transmitted signal power \( P_b \) is also assumed known.

We also assume

\[ c_k^{(m)} = d^{(m)} u_k^{(m)} \]  
(14)
where

\[ d = \{d^{(m)}\} \text{ with } d^{(m)} \in \{+1,-1\} \]  
(15)
is random scrambling code commonly used by all users, and

\[ u = \{u_k^{(m)}\} \]  
(16)
is a user-specific orthogonal unified complex Hadamard transform spreading sequence with period \( N \). Thus,

\[ c^{(k)} = \{c_k^{(m)}\} \]  
(17)
is a random sequence with

\[ M \{c^{(m)}(c^{(m)})^*\} = 0, \quad m \neq n \]  
(18)
for all \( k \) and \( k' \), where \( M \{ \cdot \} \) denotes the expectation. The final results of our analysis in this paper are applicable to systems that use long PN scrambling sequences such as m-sequences and Gold sequences. This is because the periods of these long scrambling codes are much larger than that of the spreading factor \( N \), and have correlation properties similar to those of the random scrambling sequences. At the base station transmitter in mobile communication system, the signals of all \( K \) users are symbol synchronously added before passing through a frequency-selective multipath fading channel. The complex baseband equivalent impulse response of the multipath channel can be presented in the following form:

\[ h(t) = \sum_{l=1}^{L} a_l \exp(j\theta_l) \delta(t-\tau_l) \]  
(19)
where \( L \) is the number of resolvable propagation paths, and \( a_l \exp(j\theta_l) \) and \( \tau_l \) are the complex fading factor and propagation delay of the \( l \)-th path, respectively. Note, that \( a_l \) may be Rayleigh-, Rician-, or Nakagami-distributed, depending on a specific channel model. All random variables in (19) are assumed independent for \( l \). The channel parameters, such as \( a_l \), \( \theta_l \), and \( \tau_l \) are here assumed to be known in the despreading and demodulation process, although in practice, the impulse response of the channel is typically estimated using pilot symbols or a pilot channel. Moreover, we assume that the multipaths at the input of the generalized receiver are resolvable and chip-synchronized, i.e., they are spaced at least one chip duration apart in time, and the relative delays are multiples of the chip duration.

Without loss of generality, the resolved paths are assumed to be numbered such that

\[ 0 \leq \tau_0 < \tau_1 < \cdots < \tau_{L-1} < T \]  
(20)

Hence, the baseband complex representation of the signal input to the generalized receiver of any user is given by

\[ x(t) = \sum_{k=0}^{K-1} \sum_{l=0}^{L-1} \sqrt{P_b} a_k \exp(j\theta_l) b_k(t-\tau_l) c_k(t-\tau_l) + n(t) \]  
(21)
where \( n(t) \) is the complex background additive white Gaussian noise with zero mean and one-sided power spectral density \( N_0 \) and variance \( \sigma^2 \).

In order to mitigate the multipath fading effect, the generalized receiver with coherent demodulation is employed. The generalized receiver structure is presented in Fig.1, where the number of fingers is equal to the number of resolvable paths. Since the symbols \( b_k^{(m)} \) are independent and identically distributed, from one symbol duration to another and from one user to another, without loss of generality, we focus our attention on the generalized receiver output of the user 0 for the zeroth transmitted symbol. The complex generalized receiver output of the \( i \)-th finger of user 0 in accordance to the generalized approach to signal processing in noise [9]-[13] is given by

\[ z_i = Re \left \{ \int_{t_i}^{t_i+T} 2x(t) \sqrt{P_b} c_0^{*}(t-\tau_l) \exp(-j\theta_l) dt \right \} \\
- \int_{t_i}^{t_i+T} x(t) x^*(t-\tau_l) dt \right \} + \Delta_i(t) \]  
(22)
where \( \sqrt{P_b} c_0^{*}(t-\tau_l) \exp(-j\theta_l) \) is the reference signal generated by the generalized receiver [10], [13]; \( r_l \) is a delay factor that can be neglected for simplicity of analysis; \( \Delta_i(t) \) is the background noise forming at the output of the generalized receiver, which is determined in the following manner:

\[ \Delta_i(t) = \int_{t_i}^{t_i+T} \eta(t) \phi(t-\tau_l) dt - \int_{t_i}^{t_i+T} \xi(t) \phi(t-\tau_l) dt \]  
(23)
Here \( \xi(t) \) is the noise forming at the output of the preliminary filter of input linear tract of the generalized detector; \( \eta(t) \) is the noise forming at the output of the additional filter of input linear tract of the generalized detector; the noise \( \xi(t) \) and \( \eta(t) \)
are uncorrelated between each other and have the same statistical characteristics as the noise \( n(t) \) at the input of the generalized receiver [13]. Further analysis is carried out under the main condition of functioning the generalized receiver? equality be-tween the power of information signal \( P_{s_k} \) and power of refer-ence signal (model signal) \( P_{s_k}^* \) [10], i.e.,

\[
P_{s_k} = P_{s_k}^*.
\]

\( R_{\theta}(x) \) is the real part of \( x \). Later, we will use \( \text{Im}(x) \) to denote the imaginary part of \( x \). Finally, the combiner output of the generalized receiver that produces a decision statistic is given by

\[
Z = \sum_{i=0}^{L-1} w_i z_i,
\]

where the selection of the combining weights \( w_i \) ||| determines the specific diversity-combining technique.

\[\begin{align*}
\tau & \rightarrow z_0(\tau) \\
& \quad \overset{GD_0}{\rightarrow} \quad \overset{X}{\rightarrow} \\
\tau & \rightarrow z_1(\tau) \\
& \quad \overset{GD_1}{\rightarrow} \quad \overset{X}{\rightarrow} \\
& \quad \vdots \\
\tau & \rightarrow z_{L-1}(\tau) \\
& \quad \overset{GD_{L-1}}{\rightarrow} \quad \overset{X}{\rightarrow} \\
\tau & \rightarrow Z
\end{align*}\]

Fig. 1. Generalized receiver with \( L \) fingers.

IV. PERFORMANCE ANALYSIS

A. SINR at the Generalized Receiver Output

In this section, we investigate SINR by considering the generalized receiver shown in Fig.1. It follows from (21)\( ^\text{?} \) 24 that the output of the \( i \)-th generalized receiver finger for user 0 can be presented in the following form

\[
z_i = T \sigma_{\theta} \alpha_i b_i^{(0)} + I_{MMP}^{(i)} + I_{MP}^{(i)} + \Delta_i(t),
\]

where the first term is the signal component; the second term is the multiple-user interference component determined by

\[
I_{MMP}^{(i)} = \text{Re} \left\{ \sum_{k=1}^{K-1} P_{s_k} \alpha_k b_k^{(0)} \int_{\tau_i}^{\tau} \tau_{\sigma_{\theta}} \alpha_i b_i^{(0)}(t-\tau) \alpha_i^*(t-\tau) dt \right\};
\]

the third term is the multipath interference component given by

\[
I_{MP}^{(i)} = \text{Re} \left\{ \sum_{k=1}^{K-1} \sum_{l=0, l \neq i}^{K-1} P_{s_k} \alpha_i \exp[j(\theta_i - \theta_l)] \right\}
\]

and the fourth term \( \Delta_i(t) \) is the background noise at the generalized receiver output given by (23). It can be seen that the multiple-user interference component \( I_{MMP}^{(i)} \) and the multipath interferences component \( I_{MP}^{(i)} \) are due to the interference from the \( i \)-th path of other users’ signals and from the remaining \( L-1 \) paths from all users’ signals, respectively, and the background noise at the generalized receiver output \( \Delta_i(t) \) is independent and identically distributed random variables obeying to the asymptotic Gaussian distribution with zero mean and variance of \( 4T \sigma_{\theta}^2 \).

Define aperiodic correlation function \( R_{\theta,n}(\tau) \) by

\[
R_{\theta,n}(\tau) = \begin{cases} 
\sum_{p=0}^{N-1} c_{\theta}^{(n)}(c_{\theta}^{(n+p)})^*, & 0 \leq q \leq N-1 \\
\sum_{p=0}^{N-1} c_{\theta}^{(n+p)}(c_{\theta}^{(n)})^*, & 1 \leq q \leq N \\
0, & q > N.
\end{cases}
\]

Then, let

\[
\tau_i - \tau_i = \hat{q}_i T_c.
\]

With the assumption of chip synchronization, it can be obtained that \( q_i \) is an integer and the multiple-user interference and multipath interference components can be presented in the following form:

\[
I_{MMP}^{(i)} = \text{Re} \left\{ T \sigma_{\theta} \sum_{k=1}^{K-1} P_{s_k} \alpha_k \exp[j(\theta_i - \theta_l)] R_{\theta,n}(0) \right\},
\]

\[
I_{MP}^{(i)} = R_{\theta,n}(q_i - N),
\]

where

\[
R_{\theta,n}(q_i - N) = b_k^{(0)} b_k^{(0)}(q_i - N) + b_k^{(0)} b_k^{(0)}(q_i).
\]

In the following, we shall compare the SINR at the generalized receiver output using the unified complex Hadamard transform spreading codes with that using the WH spreading sequences. Note, that when the orthogonal spreading codes are employed in downlink, such as the WH real spreading codes and the orthogonal unified complex Hadamard transform spreading codes in this paper, we have

\[
R_{\theta,n}(q_i - N) = 0, \ k \neq 0.
\]

Hence, the multiple-user interference component is equal to zero, i.e.,

\[
I_{MMP}^{(i)} = 0.
\]
When the WH spreading codes are used in the DS-CDMA downlink system, we can obtain that the multipath interference component takes the following form:

\[
I_{M}^{(1)} = \sum_{i=0}^{L-1} \sum_{k=0}^{K-1} P_{k} a_{i} \cos(\theta_{i} - \theta_{k}) R_{0}
\]

(37)

Due to the mutually independent random variables \( a_{i} \) for \( 0 \leq k \leq K - 1 \), the multipath interference component \( I_{M}^{(1)} \) has zero mean. In the light of (18), it can be easily shown via straightforward computation that, for the WH spreading sequences we have

\[
M\{R_{0}^{2}\} = M\{\hat{R}_{0}^{2}\} = N.
\]

(38)

Hence, the variance of the multipath interference component \( I_{M}^{(1)} \) is denoted by \( \sigma_{M}^{2} \) and can be determined in the following form:

\[
\sigma_{M}^{2} = N \int_{0}^{L} \sum_{i=0}^{L-1} a_{i}^{2} \cos^{2}(\theta_{i} - \theta_{k}) \sum_{k=0}^{K-1} P_{k}^{2}.
\]

(39)

Therefore, the SINR at the generalized receiver i-th finger output for the WH spreading codes is determined by

\[
SINR^{(i)}_{M} = \frac{E_{b} a_{i}^{2}}{\sqrt{1 + \sum_{j=0}^{L-1} a_{j}^{2} \cos^{2}(\theta_{j} - \theta_{k}) \sum_{k=0}^{K-1} E_{b}^{2} + 4 \sigma_{M}^{2}}},
\]

where

\[
E_{b} = P_{b} T_{b}, \quad k = 0, 1, \ldots, K - 1
\]

(41)

is the energy per data symbol of the m-th user.

Similarly, when the unified complex Hadamard transform spreading codes are used in the DS-CDMA downlink system, we can obtain that the multipath interference component \( I_{M}^{(1)} \) takes the following form:

\[
I_{M}^{(1)} = \sum_{i=0}^{L-1} \sum_{k=0}^{K-1} P_{k} a_{i} \cos(\theta_{i} - \theta_{k}) \Re\{R_{0}\}
\]

(42)

Note that the multipath interference component \( I_{M}^{(1)} \) has zero mean. If the half-spectrum property unified complex Hadamard transform spreading sequences are employed, it can be easily shown from (18) via straightforward computation that

\[
M\{\Re\{R_{0}\}\}^{2} = M\{\Im\{R_{0}\}\}^{2} = \frac{N}{2}.
\]

(43)

Hence, the variance of the multipath interference component \( I_{M}^{(1)} \) is denoted by \( \sigma_{M}^{2} \) and can be determined in the following form:

\[
\sigma_{M}^{2} = \frac{N}{2} \int_{0}^{L} \sum_{i=0}^{L-1} a_{i}^{2} \cos^{2}(\theta_{i} - \theta_{k}) \sum_{k=0}^{K-1} E_{b}^{2}.
\]

(44)

B. Generalized Receiver Finger Weights

It is noted that the interference plus thermal noise power se-n by different fingers of the generalized receiver is different. Under the assumption that the multipath interference signals are uncorrelated from one finger to another, the optimal weight in terms of the maximizing generalized receiver SINR is dependent on the multipath interference and is called modified maximal ratio combining (MMRC). Note that for a traditional maximal ratio combining (MRC), the weights are chosen as

\[
w_{i} = a_{i}.
\]

(47)

Here if the MMRC is employed as the combiner, the combining weights for DS-CDMA with WH codes and DS-CDMA with the unified complex Hadamard transform spreading sequences employing the generalized receiver are different. This is a direct consequence of the difference in the generalized receiver finger output SINR values of the two systems.

For the WH spreading sequences in DS-CDMA system employing the generalized receiver, MMRC weights take the following form:

\[
w_{i}^{M} = \frac{E_{b} a_{i}^{2}}{\sqrt{1 + \sum_{j=0}^{L-1} a_{j}^{2} \cos^{2}(\theta_{j} - \theta_{k}) \sum_{k=0}^{K-1} E_{b}^{2} + 4 \sigma_{M}^{2}}},
\]

(48)

For the unified complex Hadamard transform spreading sequences in DS-CDMA system employing the generalized receiver, MMRC weights have the following form:

\[
w_{i}^{M} = \sqrt{\frac{E_{b} a_{i}^{2}}{\sqrt{1 + \sum_{j=0}^{L-1} a_{j}^{2} \cos^{2}(\theta_{j} - \theta_{k}) \sum_{k=0}^{K-1} E_{b}^{2} + 4 \sigma_{M}^{2}}}}.
\]

(49)

After the MMRC scheme, the generalized receiver SINR is equal to the sum of all fingers? SINRs, as in (40) or (46). Using the unified complex Hadamard transform spreading sequences in DS-CDMA system employing the generalized
receiver ensures that the SINR is independent of the phase offset between different paths, while using the WH real sequences causes the SINR to be related to the squared cosine of the phase offset between paths, as seen by comparing (40) and (46). Even though the average SINR per finger of the WH codes is the same as that of the unified complex Hadamard transform spreading sequences, owing to
\[
M\{\cos^2(\theta_i - \theta_j)\} = 0.5,
\] (50)
the distribution of SINR over the random variable \(\cos^2(\theta_i - \theta_j)\) can cause degradation in bit error rate (BER) performance under some conditions. This is analogous to the case of a transmission over flat Rayleigh fading channels: even when the Rayleigh gain has a mean of 1, the performance is far worse than in an additive white Gaussian noise channel.

Now, we give the BER performance comparison of systems employing the unified complex Hadamard transform and WH spreading sequences. To analyze the performance of spreading sequences and the diversity-combining schemes considered, we adopt a Gaussian approximation approach based on the central limit theorem [18]. The Gaussian approximation is known not only to give accurate estimations of the error probability in the region of practical interest, but also to offer insights into the effects of various sequence and system parameters and interference sources on the performance of the generalized receiver [12]. For simplicity, here we only consider the single finger generalized receiver case, i.e., the generalized receiver has only one demodulating finger, and the finger is locked onto an arbitrary path, say, the \(i\)-th multipath component. Under the Gaussian approximation, for the WH spreading sequences, a straightforward derivation based on the decision process indicates that the conditional symbol error probability (SEP) for a given
\[
\theta = (\theta_0, \theta_1, \ldots, \theta_{L-1})
\] (51)
takes the following form:
\[
\text{SEP}_{\text{WH}}(\theta) = \Phi\left(\sqrt{\text{SINR}_{\text{WH}}^{(s)}}\right),
\] (52)
where \(\text{SINR}_{\text{WH}}^{(s)}\) is as in (40), and
\[
\Phi(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} \exp\left(-\frac{y^2}{2}\right) dy
\] (53)
is the error integral. Averaging (52) with respect to the associated random variables, the average error probability may be determined in the following form:
\[
\text{SEP}_{\text{WH}} = M\left\{\text{SEP}_{\text{WH}}(\theta)\right\}.
\] (54)

The averaging may most efficiently be carried out via the Monte Carlo or Matlab techniques. For the unified complex Hadamard transform spreading sequences, under the Gaussian approximation, the conditional symbol error probability can be determined in the following form:
\[
\text{SEP}_{\text{WH}} = \Phi\left(\sqrt{\text{SINR}_{\text{WH}}^{(s)}}\right),
\] (55)
where \(\text{SINR}_{\text{WH}}^{(s)}\) is as in (46).

At this point, we should compare (54) and (55). If any function \(f(x)\) is a convex function and \(X\) is a random variable, then
\[
M\{f(X)\} \geq f\{M\{X\}\}
\] (56)
is satisfied. To apply Jensen's inequality, first define
\[
X = \cos^2(\theta_i - \theta_j).
\] (57)
Then since \(\theta_i, i = 0, 1, 2, \ldots, L-1\) are uniformly distributed within the limits of the interval \([0, 2\pi]\), straightforward calculations give
\[
M\{X\} = 0.5.
\] (58)
Moreover, the function \(f(x)\) here has the following form:
\[
f(x) = \Phi\left(\frac{1}{a + bx}\right),
\] (59)
where \(a > 0\) and \(b \geq 0\). Calculating the second derivative of the function \(f(x)\) with respect to \(x\), we find that \(f(x)\) is a convex function if
\[
a + bx \leq \frac{1}{3}.
\] (60)
A sufficient condition
\[
E_{\text{b}}^2 \geq 3 \sigma_{\text{b}}^2 \left[ N \sum_{i=L_0}^{L-1} a_i^2 \sum_{k=0}^{K-1} E_{\text{b}}^2 + 4\sigma_a^4\right]
\] (61)
satisfies this inequality and allows applications of Jensen's inequality successively to each component in \(\theta\) by using (52) and (54). Then we obtain
\[
\text{SEP}_{\text{WH}} \geq \Phi\left(\sqrt{\text{SINR}_{\text{WH}}^{(s)}}\right),
\] (62)
where
\[
\text{SINR}_{\text{WH}}^{(s)} = \frac{E_{\text{b}}\sigma_{\text{b}}^2}{\sqrt{2N \sum_{i=L_0}^{L-1} a_i^2 \sum_{k=0}^{K-1} E_{\text{b}}^2 + 4\sigma_a^4}}.
\] (63)
Comparing the result of (62) and (63) for the WH spreading sequences and the result of (55) for the unified complex Hadamard transform spreading sequences, we found that
\[
\text{SINR}_{\text{WH}}^{(s)} = \text{SINR}_{\text{H}}.
\] (64)
Thus, (55) is a lower bound on the symbol error probability for the WH spreading sequences when the condition (61) is satisfied at employment of the generalized receiver by DS-CDMA system. This result implies that the DS-CDMA system employing the generalized receiver with the unified complex Hadamard transform spreading sequences is more resistant to multiple-access interference (MAI) than that with WH spreading sequences in the case where only one finger is selected in the generalized receiver. For more fingers and various combining schemes, although simple closed-form bounds for the symbol error probability of the DS-CDMA system employing the generalized receiver with the WH spreading sequences are difficult to obtain, we believe that the same conclusion can be made under some similar conditions obtained by using Jensen's inequality and some intensive calculations. This conclusion can be verified by numerical simulations in the next section. Furthermore, in view of (55)
and (61)? 63, we observe that the DS-CDMA system employing the generalized receiver with the unified complex Hadamard transform spreading sequences can achieve high reliable performance at not only high SNRs \( \frac{E_b}{N_0} \), as in the case of the Rake receiver, but at low SNRs, too.

V. SIMULATION

In this section, the bit error rate (BER) performance of DS-CDMA system employing the generalized receiver is compared among the WH spreading sequences and the unified complex Hadamard transform spreading sequences under different combining schemes for finger weights such as traditional equal gain combining (EGC), MRC and MMRC. Also a comparative analysis of BER performance under employment of the generalized receiver and Rake one is made. Simulations are performed over the Rayleigh, Rician, and additive white Gaussian noise channels, respectively. The spreading sequences of length \( N = 64 \) are considered, and powers are chosen as

\[
P_{b_0} = P_{b_1} = \cdots = P_{b_{k-1}} = 1.
\]

(65)

 Unless stated otherwise, the default system under consideration contains \( K = 10 \) active users, and the number of paths \( L = 4 \) in Rayleigh fading multipath channel with

\[
M\{a_i^n\} = 1, \quad i = 0, 1, \ldots, L - 1,
\]

(66)

and uses the MMRC technique for the generalized receiver. According to Section 2, we choose

\[
\mu_1 = 1, \quad \mu_2 = -j, \quad \text{and} \quad \mu_3 = j
\]

(67)

to construct a set of the half-spectrum property unified complex Hadamard transform spreading sequences. Note that a different sequence assigned to the user-of-interest results in different BER; the BER in the following examples is an average over the sequence subset.

A. Effects of Generalized Receiver Finger Weights

Different combining techniques in the generalized receiver are compared in Fig. 2, including the traditional EGC, MRC and MMRC. Also comparison is made with the Rake receiver.

It is shown that the generalized receiver using the MMRC technique has superiority in comparison with EGC and MRC techniques, especially in the high SNR case.

This result can be explained by the generalized receiver finger weights defined in (48) and (49), where the effect of interference is taken into consideration in MMRC, but it is not considered in the other two combining techniques. Also, Fig. 2 demonstrates a great superiority under employment of the generalized receiver in DS-CDMA system with the unified complex Hadamard transform and WH spreading sequences in comparison with the use of the Rake receiver.

B. Effect of Different Fading Channels

Figure 3 shows the BER performance of DS-CDMA system spread by the unified complex Hadamard transform and WH spreading sequences over different channels, including the additive white Gaussian noise channel, Rayleigh fading channel, and Rician fading channel with factor \( F = 3 \). It can be seen that in all the channels under consideration, the DS-CDMA system employing the generalized receiver with the unified complex Hadamard transform spreading sequences has better performance, especially in the high SNR case. Also, we can see a great superiority under employment of the generalized receiver in DS-CDMA system with the unified complex Hadamard transform and WH spreading sequences over the use of the Rake receiver.
C. Effect of MAI

To show the effect of MAI, Fig. 4 depicts the BER performance as a function of the number of active users $K$. The BER performance of DS-CDMA systems employing the generalized receiver with the unified complex Hadamard transform spreading sequences is better than that of the same system with the WH spreading sequences at high SNR and all the chosen values of $K$, although, the advantage brought by using the unified complex Hadamard transform spreading sequences fades gradually with the increase of MAI, for example, a large number of users and small SNRs, and BER performance of DS-CDMA systems employing the generalized receiver with the unified complex Hadamard transform spreading sequences is comparable to that with the WH spread system in the low SNR case. Also, for all cases, a great advantage under employment of the generalized receiver in comparison with the Rake one is evident. These results in simulations have verified our analytic results.

VI. CONCLUSIONS

In this paper, the orthogonal unified complex Hadamard transform spreading sequences are employed in DS-CDMA downlink systems using the generalized receiver constructed based on the generalized approach to signal processing in noise. The SINR at the generalized receiver output of the DS-CDMA system employing the unified complex Hadamard transform spreading sequences is independent of the phase offsets between different paths, while the SINR of the system using the WH real spreading sequences is related to the squared cosine of path phase offsets. Therefore, the unified complex Hadamard transform spreading sequences provide better performance in terms of BER relative to the system with the WH real spreading sequences at high SNRs. The performance has also been evaluated through simulations, demonstrating the benefits of the unified complex Hadamard transform spreading sequences and MMRC weights in the generalized receiver. Comparative analysis between the generalized receiver constructed on the basis of the generalized approach to signal processing in noise and the Rake receiver employed by DS-CDMA systems with the unified complex Hadamard Transform and WH spreading sequences showed a great superiority of the first receiver over the last one.

REFERENCES


Vycheslav P. Tuzulkov (M'84-SM'07) received B.S. degree in physics and mathematics from the Belarusian State University, Minsk, Belarus in 1973, and the M.S. and Ph.D. degrees in radio physics from the Belarusian State University, Minsk, Belarus, in 1976 and 1990, respectively.

From 1984 to 2004, he was with the Institute of Engineering Cybernetics (the Institute of Informatics Problems from the middle of June 2002), National Academy of Sciences of Belarus, Minsk, where he had served as Research Scientist, Senior Research Scientist, and Chief Research Scientist. From 2000 to April 2002 he has been a Visiting Full Professor with the University of Aizu, Aizu-Wakamatsu City, Japan. From 2003 to 2007 he served as Invited Full Professor at the Ajou University, Suwon, Korea. From 2008 to 2009 he was Professor of the Electrical Engineering Department of the Yeungnam University, Gyeonggian, Korea. Starting in March 2009 he is a Professor of the School of Electrical Engineering and Computer Sciences, Kyungpook National University, Daegu, Korea. His research interests are in signal processing, detection and estimation in radar, communications, satellite communications, wireless and mobile communications, sonar, underwater signal processing, remote sensing, navigational systems and other areas.

Prof. Tuzulkov is a member of SPIE, International Society of Information Fusion, and World Scientific and Engineering Academy and Society.