A High-Rate Low-PAPR Multicarrier Spread Spectrum System Using Cyclic Shift Orthogonal Keying

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***Abstract*—A bandwidth and power efficient multi-code multicarrier spread spectrum (MCSS) system is proposed in terms of a new cyclic shift orthogonal keying (CSOK) scheme. The system can operate in either the Hybrid CSOK (HCSOK) mode which combines PSK/QAM with CSOK, or the Quadrature CSOK (QCSOK) mode which provides higher data rate. Under dispersive channel, efficient maximum likelihood receivers are derived for both modes. It is shown that the CSOK MCSS system can outperform the traditional single-code MCSS system in terms of bandwidth efficiency, PAPR, and bit error rate. Furthermore, under indoor multipath fading channel, the QCSOK receiver performs only slightly inferior to the QPSK-CSOK receiver, while achieving almost twice the data rate of the later.**

Keywords-multicarrier system; spread spectrum; orthogonal keying; maximum likelihood receiver; peak-to-average power ratio.

I. INTRODUCTION

It is well known that spread spectrum (SS) technique has inherent anti-jamming capability which makes it suitable for unlicensed-band wireless applications. However, traditional direct-sequence spread spectrum (DSSS) system is not spectrally efficient and involves complex RAKE processing under multipath fading channel. On the other hand, MCSS can be regarded as a combination of orthogonal frequency division multiplexing (OFDM) and frequency domain spreading [1]. Hence, it improves the bandwidth efficiency, and can combat inter-symbol interference (ISI) via cyclic prefix (CP) and FFT-based frequency domain equalizer (FDE). In fact, as the spreading factor (SF) equals to one, the single-code MCSS system degenerates to OFDM system, having no processing gain and multipath diversity effect. However, traditional Walsh-coded MCSS (WC-MCSS) system, like OFDM, suffers from a high peak-to- average-power ratio (PAPR). Besides, the number of bits per symbol is reduced as the SF is increased, making it insufficient for high data rate application. To reduce the PAPR, different spreading codes have been studied [2,3]. For increasing the throughput, multi-code SS systems have been proposed [4-6]. In [6], a novel cyclic-shift *M*-ary bi-orthogonal (CS-MBOK) MCSS scheme was proposed by the authors. In this paper, a new multi-code MCSS system structure and receiver design are proposed to improve the bandwidth efficiency as well as error rate, while reducing the PAPR at the same time. The key to its success lies in the so called cyclic shift orthogonal keying (CSOK), which is used to map data bits into *N* orthogonal spreading codes generated by cyclically shifting an *N*-point base code. In this paper, the Chu sequence is exploited due to its perfect cyclic autocorrelation property [7]. Beginning with the basic CSOK scheme, we propose two operation modes for flexible improvements on bandwidth efficiency, BER performance, and PAPR. The first is called the Hybrid CSOK (HCSOK) mode which combines PSK/QAM symbol with CSOK symbol. To attain higher data rate, the second mode, called quadrature CSOK (QCSOK) mode, is proposed, which can be regarded as two parallel but quadrature BPSK-CSOK systems. Efficient maximum likelihood receivers are then derived for both modes under dispersive channel. Low-complexity FFT-based receiver structure can be obtained to implement the maximum ratio combiner and CSOK symbol demapping. Compared to the WC MCSS system, it is shown that the proposed CSOK MCSS system can significantly improve the error performance, bandwidth efficiency, and PAPR in a flexible manner by choosing the operating mode and code length *N*.

This paper is organized as follows. In Section II, the CSOK scheme and two associated operating modes are described. In Section III, the maximum likelihood receivers are derived. Section IV includes the simulation results. Finally, conclusions are made in Section V.

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II. SYSTEM AND SIGNAL MODEL

A. Cyclic Shift Orthogonal Keying

As mentioned before, the proposed CSOK system is in nature a multi-code system where *N* mutually orthogonal codes are used for frequency-domain spreading of the information data. These codes can then be regarded as symbols in orthogonal keying, and each one can carry $R = \log_2 N$ bits. To generate such mutually orthogonal codes easily, the idea of cyclic code shifted keying [8] is used. First, we choose an *N*-point base code $\mathbf{c}_{(0)} = [c_0\ c_1 \dots c_{N-1}]^T$, and then generate the code set $C_N = \{c_{(k)}, k=0,1,\dots, N-1\}$, where

$$\mathbf{c}_{(m)} = [c_m \ \cdots \ c_{N-1} \ c_0 \ \cdots \ c_{m-1}]^T \qquad (1)$$

Note that the desired mutually orthogonality property of the *N* CSOK symbols can be verified via the periodic auto-correlation function (PACF) of $\mathbf{c}_{(0)}$ which is expressed as

$$\rho(\lambda) = \frac{1}{N}\mathbf{c}^H_{(m)}\mathbf{c}_{(m-\lambda)} \ \ ; \ m, \lambda = 0 \sim N-1 \qquad (2)$$

If $\mathbf{c}_{(0)}$ has an ideal impulse-like PACF, the code set C_N forms an alphabet for orthogonal keying. The desired property can be satisfied by some poly-phase sequences, such as the Chu sequence [7] and Frank-Zadoff sequence [9].

Here, we choose the N-point Chu sequence as the base code. For even N, the sequence is given by

$$c_n = e^{j\pi n^2 q/N}, \qquad 0 \le n \le N-1 \tag{3}$$

where q is an integer with gcd(q,N)=1. Besides the ideal PACF property, the constant modulus feature of the Chu sequence can greatly alleviate the annoying PAPR problem occurred in most multi-carrier systems.

B. Two CSOK Transmission Modes

Now, an R-bit CSOK symbol can be mapped to a code in C_N and then transmitted on N sub-carriers. In addition, we can combine a CSOK symbol with basic digital modulation symbol like PSK or QAM to carry more information bits. This idea leads to the hybrid CSOK (HCSOK) mode, whose block diagram is shown in Fig. 1. At the i^{th} symbol epoch, the HCSOK symbol can be represented as:

$$\mathbf{c}_i = d_i \mathbf{c}_{(m_i)} \tag{4}$$

where d_i and m_i denote the i^{th} digital modulation symbol and the index (cyclic shift) of the CSOK symbol, respectively. To calculate the bandwidth efficiency, first we denote P as the number of bits carried by d_i. Then one HCSOK symbol corresponds to μ=R+P bits/symbol. Specifically, if $d_i = \pm 1$, we have the BPSK-HCSOK case which also corresponds to M-ray bi-orthogonal keying (MBOK) [6] for M=2N, and μ=R+1. In general, MPSK-HCSOK symbol can be used to increase μ without destroying the constant envelope property. On the other hand, QAM-HCSOK symbol is more efficient to raise the data rate, but deteriorates the PAPR. Higher modulation levels of MPSK/QAM will also result in significant BER degradation.

To achieve higher throughput without losing performance, we propose another CSOK transmitted mode as shown in Fig.2. It is called quadrature CSOK (QCSOK) mode. This mode transmits two BPSK-HCSOK signals simultaneously over two phase-quadrature I/Q channels using the same Chu sequence. Therefore, the number of bits per codeword can be raised to μ=2(R+1), at the price of a moderate PAPR increase.

To represent the QCSOK signal, let $\mathbf{c}_i^I$ and $\mathbf{c}_i^Q$ be the two BPSK-HCSOK codewords transmitted on the I- and Q-channel, respectively. Then the QCSOK codeword is expressed as:

$$\mathbf{c}_i = \mathbf{c}_i^I + j\mathbf{c}_i^Q = d_i^I \mathbf{c}_{(m_i^I)} + j d_i^Q \mathbf{c}_{(m_i^Q)} \tag{5}$$

where d_i^I (d_i^Q) is the BPSK symbol and m_i^I (m_i^Q) $\in \{0, \cdots, N-1\}$ is the cyclic shift index of the I- (Q-) channel CSOK symbol.

After taking the IFFT of the composite codeword $\mathbf{c}_i$, a cyclic prefix with duration T_{CP} is inserted, resulting in the symbol duration of $T_{sym} = T_{FFT} + T_{CP}$, where T_{FFT} denotes the FFT duration, and T_{CP} is chosen to be larger than the maximum delay spread of the multipath channel. Hence, the estimated channel bandwidth for the system remains the same as that of conventional MCSS system. It can be easily shown that the bandwidth efficiency is $(P + \log_2 N) T_{FFT} / (N+1) T_{SYM}$ (bps/Hz) for the HCOSK mode, and $(2 + 2\log_2 N) T_{FFT} / (N+1) T_{SYM}$ for the QCSOK mode. Being an MCSS system, the proposed system inherits the advantages of low receiver complexity for achieving multipath diversity, and processing gain to suppress co-channel interference (CCI). Being an orthogonal keying system too, it can choose different N for flexible tradeoff between bit error rate and bandwidth efficiency, while the traditional single-code MCSS system does not offer such flexibility.

With the flexible combination of the two modes and different N, Table I summarize many cases for the value of μ in bits/symbol. It is clear that the QCSOK mode can achieve almost twice the throughput as the QPSK-HCSOK mode as N is large. Moreover, since the QCSOK mode transmits two BPSK-HCSOK streams over the I/Q channels simultaneously, it can attain the same BER performance in AWGN channel as the single-channel BPSK-HCSOK mode does. But in multipath channel, frequency-selective channel response will destroy the orthogonality between the I/Q channels, and cause multi-code interference (MCI). To tackle the channel problem, maximum likelihood receivers will be designed in section III.

III. Maximum Likelihood Receiver Design

A. ML Receiver for HCSOK Mode

Consider a HCSOK MCSS system with a single user, and let the received signal be sampled at the chip rate at the receive filter output. After removing the CP and taking FFT, the discrete-time $N \times 1$ received vector for the i^{th} HCSOK symbol can be represented as follows

$$\mathbf{r}_i = \mathbf{H}\mathbf{c}_i + \mathbf{F}\mathbf{w} = d_i \mathbf{H}\mathbf{c}_{(m_i)} + \mathbf{v} \tag{6}$$

where $\mathbf{H} = \mathrm{diag}\left[H_0 \ \cdots \ H_{N-1} \right]$ is a diagonal matrix formed by channel frequency response on the N sub-carriers, $\mathbf{F}$ denotes the $N \times N$ Fourier matrix, and $\mathbf{w}$ is the AWGN noise. Recalling the HCOSK codeword in (4), two constituent symbols d_i and m_i are to be detected simultaneously in order to retrieve the μ data bits. Hence, we apply the maximum likelihood criterion [10] to obtain the optimum decisions of d_i and m_i as follows:

$$\left(\hat{d}_i, \hat{m}_i \right) = \arg\min_{d_i, m_i} \left[\left\| \mathbf{r}_i - \mathbf{H}\mathbf{c}_i \right\|^2 \right] \tag{7}$$

Since the detection rule is symbol-by-symbol, for simplicity, we can neglect the symbol index i in following discussion. The cost function in (7) can be further simplified as follows:

$$\left\| \mathbf{r} - \mathbf{H}\mathbf{c} \right\|^2 = \left\| \mathbf{r} \right\|^2 - 2\,\mathrm{Re}\left\{ \left(d\mathbf{H}\mathbf{c}_{(m)} \right)^H \mathbf{r} \right\} + \left\| \mathbf{H}\mathbf{c}_{(m)} \right\|^2 \tag{8}$$

In RHS of (8), the first and third terms are constants. As a result, the ML symbol decision can be obtained as

$$\left(\hat{d}, \hat{m} \right) = \arg\max_{d,m} \left[\mathrm{Re}\left\{ d^* \mathbf{c}_{(m)}^H \mathbf{H}^H \mathbf{r} \right\} \right] \tag{9}$$

where the superscripts * and ^{H} denote complex conjugate and Hermitian transposition, respectively, and the notation $\mathrm{Re}\{x\}$

means taking the real part of $x \in Z$.

A block diagram shown in Fig. 3 depicts the HCSOK ML receiver corresponding to the decision rule (9). First, the received vector **r** is passed through a one-tap FDE with the coefficient vector $\mathbf{H}^H = \text{diag}\left[H_0^* \ \cdots \ H_{N-1}^*\right]$, yielding the output

$$\mathbf{y} = \mathbf{H}^H \mathbf{r} \tag{10}$$

In fact, the FDE can be regarded as a maximum ratio combiner (MRC) [10], which attempts to coherently collect all the symbol energy on N sub-carriers. Second, the equalized output vector **y** needs to be correlated with all possible HCSOK codewords with different cyclic shift m and modulation symbol $d = d_I + jd_Q$. To do so, one can use a correlator bank for the code set C_N, and denote its output vector as $\mathbf{z} = \left[z_0 \ \cdots \ z_{N-1}\right]^T$, where

$$z_m = z_m^I + jz_m^Q = \mathbf{c}_{(m)}^H \mathbf{y}, \quad m = 0 \sim N\text{-}1 \tag{11}$$

However, due to the cyclic shift property of C_N, we can utilize an equivalent but more efficient FFT-based code correlator to obtain the same output vector **z** as

$$\mathbf{z} = \mathbf{F}^{-1}\left(\mathbf{c}_F^* \odot \mathbf{F}\mathbf{y}\right) \tag{12}$$

where $\odot$ denotes pairwise multiplication, and $\mathbf{c}_F$ is the FFT of the base Chu sequence $\mathbf{c}_{(0)}$. As N gets larger, more computational saving can be obtained. Finally, the ML detection rule (9) can be rewritten as

$$\begin{aligned}\left(\hat{d}, \hat{m}\right) &= \arg\max_{d,m}\left[\text{Re}\left\{d^* \mathbf{z}\right\}\right] \\ &= \arg\max_{d_I, d_Q, m}\left[d_I z_m^I + d_Q z_m^Q\right]\end{aligned} \tag{13}$$

For the important QPSK-HCSOK case, $d \in \pm 1 \pm j$, and the ML rule can be further simplified to the following two steps:

$$\begin{aligned}\hat{m} &= \arg\max_m\left[\left|z_m^I\right| + \left|z_m^Q\right|\right] \\ \hat{d} &= \text{sgn}\left[z_{\hat{m}}^I\right] + j \times \text{sgn}\left[z_{\hat{m}}^Q\right]\end{aligned} \tag{14}$$

For the simplest BPSK-HCSOK case, the decision rule deals with only the real part of **z**, and we have

$$\hat{m} = \arg\max_m\left[\left|z_m^I\right|\right] \quad ; \quad \hat{d} = \text{sgn}\left[z_{\hat{m}}^I\right] \tag{15}$$

B. ML Receiver for QCSOK Mode

Since the transmitted QCSOK symbol described in (5) consists of two independent BPSK-HCSOK symbols in I/Q channels, the received signal vector can be expressed as

$$\mathbf{r} = \left[d^I \mathbf{H}\mathbf{c}_{\left(m^I\right)} + jd^Q \mathbf{H}\mathbf{c}_{\left(m^Q\right)}\right] + \mathbf{v} \tag{16}$$

By using maximum likelihood criterion again, the following decision rule arises

$$\left(\hat{\mathbf{d}}, \hat{\mathbf{m}}\right) = \arg\min_{\{\mathbf{d},\mathbf{m}\}}\left[\left\|\mathbf{r} - \left(d^I \mathbf{H}\mathbf{c}_{\left(m^I\right)} + jd^Q \mathbf{H}\mathbf{c}_{\left(m^Q\right)}\right)\right\|^2\right] \tag{17}$$

where $\hat{\mathbf{d}} = \left[\hat{d}^I \ \ \hat{d}^Q\right]$, $\hat{\mathbf{m}} = \left[\hat{m}^I \ \ \hat{m}^Q\right]$. Following the derivation in (8)-(9), the ML decision rule of QCSOK mode can also be rewritten as follows:

$$\begin{aligned}\left(\hat{\mathbf{d}}, \hat{\mathbf{m}}\right) = \arg\max_{\{\hat{\mathbf{d}},\hat{\mathbf{m}}\}}\Big[& d^I \,\text{Re}\left\{\mathbf{c}_{\left(m^I\right)}^H \mathbf{H}^H \mathbf{r}\right\} + d^Q \,\text{Im}\left\{\mathbf{c}_{\left(m^Q\right)}^H \mathbf{H}^H \mathbf{r}\right\} \\ & + d^I d^Q \,\text{Im}\left\{\mathbf{c}_{\left(m^I\right)}^H \left|\mathbf{H}\right|^2 \mathbf{c}_{\left(m^Q\right)}\right\}\Big]\end{aligned} \tag{18}$$

In the objective function of (18), note that the third term is an additional multi-code interference (MCI) term introduced by the channel response. Let $\eta\left(m^I, m^Q\right) = \text{Im}\left\{\mathbf{c}_{\left(m^I\right)}^H \left|\mathbf{H}\right|^2 \mathbf{c}_{\left(m^Q\right)}\right\}$ denote this MCI term between $\mathbf{c}_{\left(m^I\right)}$ and $\mathbf{c}_{\left(m^Q\right)}$, where $\left\{m^I, m^Q\right\} \in 0 \sim N-1$, and the detection rule (18) can be simplified as

$$\left(\hat{\mathbf{d}}, \hat{\mathbf{m}}\right) = \arg\max_{\{\mathbf{d},\mathbf{m}\}}\left[d^I z_{m^I}^I + d^Q z_{m^Q}^Q + d^I d^Q \eta\left(m^I, m^Q\right)\right] \tag{19}$$

where $z_{m^I}^I = \text{Re}\left\{\mathbf{c}_{\left(m^I\right)}^H \mathbf{y}\right\}$ and $z_{m^Q}^Q = \text{Im}\left\{\mathbf{c}_{\left(m^Q\right)}^H \mathbf{y}\right\}$ are the real-part and imaginary-part of the m^I-th element and m^Q-th element of the code correlator output **z**, respectively. Under the AWGN channel, **H** = **I**, and the MCI term will be zero. Thus, (19) can be decoupled into two BPSK-HCSOK decisions as in (15) by using the real and imaginary parts of **z,** respectively. That is

$$\begin{aligned}\hat{\mathbf{m}} &= \left[\arg\max_{m^I}\left[\left|z_{m^I}^I\right|\right] \ \ \arg\max_{m^Q}\left[\left|z_{m^Q}^Q\right|\right]\right] \\ \hat{\mathbf{d}} &= \left[\text{sgn}\left[z_{\hat{m}^I}^I\right] \ \ \text{sgn}\left[z_{\hat{m}^Q}^Q\right]\right]\end{aligned} \tag{20}$$

Eq. (20) is called 1-D ML scheme. Under frequency selective channel, the ML rule in (19) needs a two-dimensional (2-D) search in the $\{m^I, m^Q\}$ space. But the computational complexity can be greatly reduced by using a suboptimal searching scheme which incurs only slight performance degradation. It is called an ML cross-road search (ML-CS) scheme which has two passes. As is shown in Fig. 5, the first pass is to use the 1-D ML scheme in (20) for finding an initial point $\hat{\mathbf{m}}_0 = \left[\hat{m}_0^I, \hat{m}_0^Q\right]$ in the search space, without considering the MCI term. Then in the second pass, we search along the "cross road" with the intersection at $\hat{\mathbf{m}}_0$. On the "cross road", the global maxima of (19) can often be found without resorting to the brute force 2-D search. Hence, the searching complexity can be reduced by a factor of $2/N$.

IV. PERFORMANCE EVALUATION RESULTS

To compare the bandwidth efficiency (BWE) and performance among various MCSS systems and for different N and operating modes, we set a common sub-carrier spacing at Δf = 625 kHz and the ratio T_{FFT}/T_{SYM} =0.75. Then the corresponding chip rate is $R_c = \Delta f \times N$. First, under the AWGN channel, Fig. 6 compares the proposed QPSK-HCSOK and QCSOK modes and the traditional QPSK WC-MCSS system in terms of BWE and the required E_b/N_0 at a BER of P_e=10^{-5} for different codeword length N=2^n, where n=1,2,…, 8. It is seen that the QCSOK mode becomes more and more bandwidth and power efficient than the MC-MCSS system, as N gets larger. Moreover, for the proposed two CSOK operating modes with the same N, it is noted that the QCSOK mode achieves the best BWE, while the QPSK-HCSOK mode requires the lowest E_b/N_0. However, for large N, say N =256, the QCSOK mode requires a little more E_b/N_0 (0.3 dB for N=256) than the QPSK-HCSOK mode, while almost doubling the data rate and BWE.

Next, the complementary cumulative distribution function (CCDF) of PAPR is shown in Fig. 7 for several MCSS systems at the codeword length N=32. To generate the CCDF, all the transmitted signals are interpolated by the over-sampling ratio OVR=4. It is noted that by using cyclically-shifted Chu sequence, the HCSOK and QCSOK signals have much lower PAPR than the QPSK WC-MCSS signal.

Thirdly, we evaluate the BER performance of the ML-based receivers for both CSOK modes under multipath fading channel. Considering indoor application, we assume that the channel response has an exponential power delay profile [11] as follows:

$$E\left[\left|h_n\right|^2\right]=\left[1-e^{-1/(R_c\tau_{rms})}\right]e^{-n/(R_c\tau_{rms})}, \qquad (21)$$

where h_n denotes the n^{th} chip-spaced channel tap, and the rms delay spread τ_{rms} is set at 70 ns. Fig. 8 shows the average BER curves of the conventional QPSK WC-MCSS, QPSK-HCSOK, and QCSOK systems over 200 channel realizations. For N=32, the QPSK-HCSOK mode outperforms the QPSK WC-MCSS systems by about 2 dB at P_e=10^{-5}, while the QCSOK mode has only an SNR degradation within 1 dB of QPSK-HCSOK. When the codeword length N is increased to 256, all systems get better BER due to more path diversity gain. But the performances of QPSK-HCSOK and QCSOK modes become closer, and they can achieve a higher SNR gain of about 4 dB as compared to the traditional QPSK WC-MCSS system. This gain is resulted from orthogonal keying.

Finally, in Fig. 9 we compare the BER performances of the original 2-D ML, the decoupled 1-D ML, and the simplified ML-CS algorithms for the cases of N=32 and 256. It is seen that the ML-CS scheme can achieve almost the same performance as the 2-D ML for N=256, although for N=32, it suffers from an error floor at about P_e=10^{-4}.

V. CONCLUSIONS

We have proposed a new multi-code MCSS system in terms of a variety of CSOK transmission modes, which exploit the Chu sequence as spreading code due to its ideal PACF and constant envelope properties. The proposed CSOK MCSS system has significant improvements in bandwidth efficiency, BER performance, and PAPR, as compared with conventional single code MCSS system. To combat the multipath fading channel, ML-based receivers have been derived for both the HCSOK and QCSOK modes, respectively. A low-complexity ML-CS algorithm has also been proposed for reducing the 2-D search space of the QCSOK mode. Finally, the system performance in the presence of co-channel interference is under study and will be presented in future work.

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TABLE I
Comparison of bits/symbol (μ) for different CSOK modes

Code Length / Mode	**N=32 (R=5)**	**N=256 (R=8)**	**N=1024 (R=10)**
QPSK-HCSOK (P=2)	7	10	12
8PSK-HCSOK (P=3)	8	11	13
16QAM-HCSOK (P=4)	9	12	14
QCSOK (P=1×2)	12	18	22

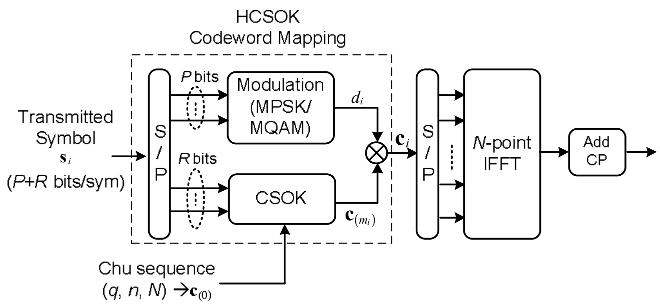

Figure 1. The transmitter structure of HCSOK MCSS system.

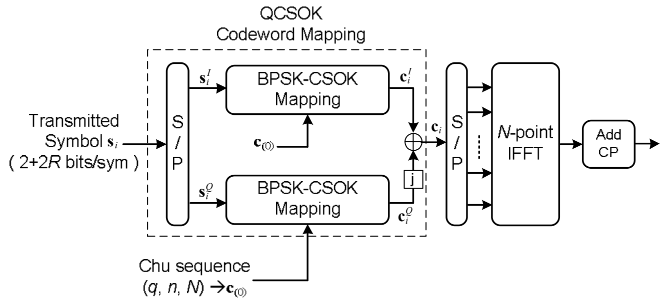

Figure 2. The transmitter structure of QCSOK MCSS system.

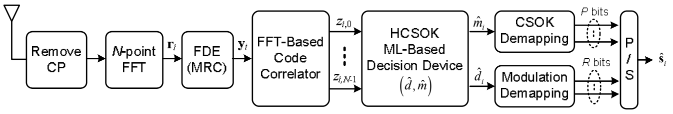

Figure 3. The ML receiver block diagram for the HCSOK MCSS system.

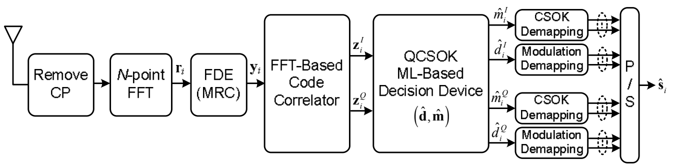

Figure 4. The ML receiver block diagram for the QCSOK MCSS system.

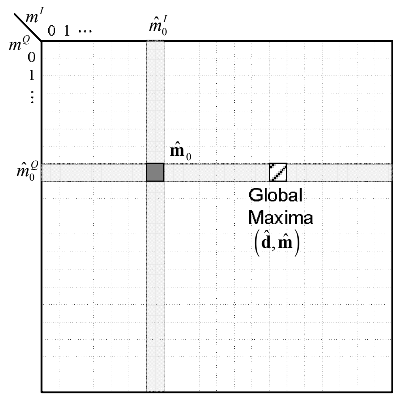

Figure 5. Illustration of the ML cross-road search (ML-CS) algorithm.

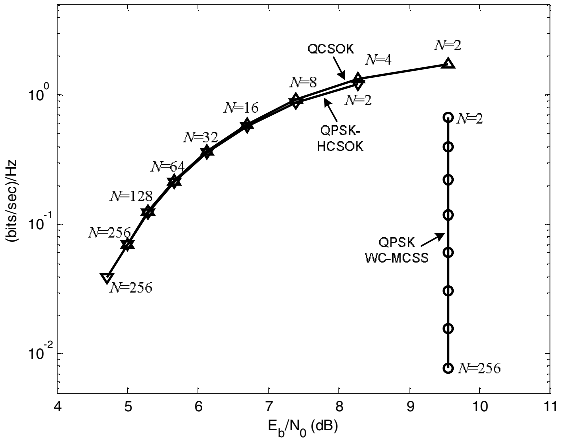

Figure 6. The BWE vs. E_b/N_0 comparison of QPSK-HCOSK, QCSOK, and QPSK WC-MCSS systems at BER P_e=10^{-5}.

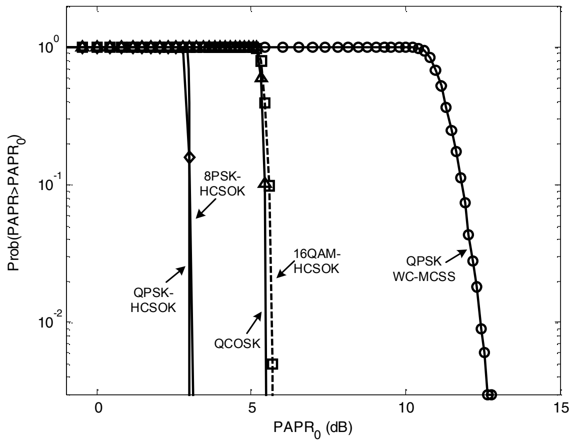

Figure 7. The PAPR CCDF comparison of QPSK/8PSK/16QAM-HCSOK, QCOSK, and QPSK WC-MCSS signals for N=32.

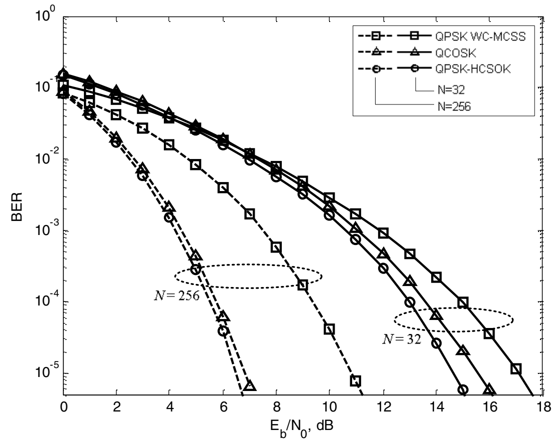

Figure 8. BER performance comparison of QPSK-HCSOK, QCOSK, and QPSK WC-MCSS system for N=32 and 256, under multipath fading channel.

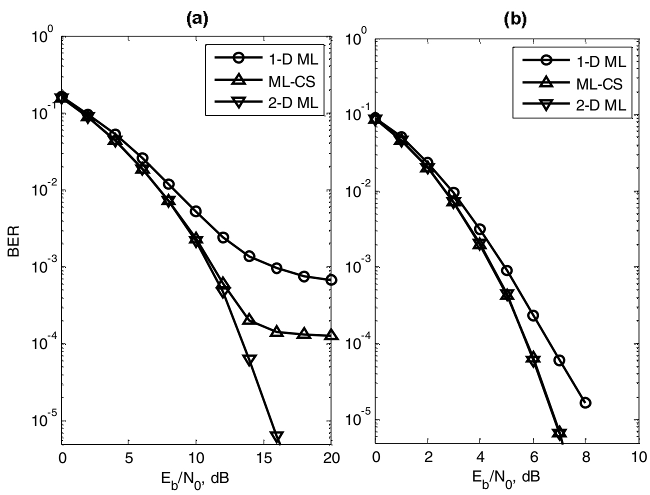

Figure 9. For QCSOK mode, the BER performances under multipath fading channel for 2-D ML, ML-CS, and decoupled 1-D ML receivers: (a) N=32 , (b) N=256.