

Differential Parallel FSOK Transceiver Design with Low PAPR for MCSS Uplink System over High Mobility Fading Channels

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Abstract—Recently, a new multi-carrier spread spectrum (MC-SS) systems with frequency-shift orthogonal keying (FSOK) has been proposed and shown to be more spectral and power efficient than the conventional MCSS systems. In this paper, an extension for multiuser and multi-stream, we develop a novel differential parallel QPSK-FSOK MCSS uplink system over high mobility multipath fading channels while preserving the low peak-to-average-power ratio (PAPR) property. The proposed differential transceiver can overcome the mobile fading channel and carrier frequency offset (CFO) effect. Finally, simulation results confirm that the ML rule designed for M -ary despreading and demapping can acquire the full multipath frequency diversity gain for the proposed system¹.

I. INTRODUCTION

Multi-carrier spread spectrum (MC-SS) systems are proposed based on the combination of a direct-sequence spread-spectrum and multi-carrier transmission technology to support high data rate transmission [1]-[4]. Similar to orthogonal frequency division multiplexing (OFDM) system, it can combat multipath fading and inter-symbol interference (ISI) via cyclic prefix (CP) and FFT-based frequency domain equalizer (FDE). However, the traditional Walsh-coded MCSS system, like OFDM, suffers from a high peak-to-average-power ratio (PAPR) [5]. In order to simultaneously achieve low PAPR and full frequency diversity gain, a novel cyclic-shift orthogonal keying (CSOK) scheme has been proposed for MC-SS systems [6]-[7]. Moreover, based on [6]-[7], Deng *et al* proposed a novel extension called the differential M -ary phase shift keying (MPSK) FSOK MC-SS system to support robust communication over high mobility multipath channels while preserving the low PAPR property [8].

In this paper, we extend the differential QPSK-FSOK transceiver [8] to the multiuser and multi-substream structure and consider the carrier frequency offset (CFO) effect. The development of transceiver involves the following stages. First, the data substream of each user is mapped onto the QPSK-FSOK symbol and spreaded by a frequency-shift orthogonal sequence at the same time. Second, the substream of each user

is differentially encoded through a “block feedback delay” operation, in which the Chu sequence [9] used for the initial differential encoding can preserve low PAPR property. Furthermore, let the differential encoded symbol of each substream repeated P times and multiplied by the frequency shift. Then, the substream mapping, user interleaving, IFFT and CP insertion are applied to the transmitted signal. At receiver, the received data is differentially equalized by the maximum ratio combining (MRC) method, and then despreaded by the set of frequency shift orthogonal sequences, which can overcome the mobile fading channel and CFO effects. Finally, the ML rule designed for M -ary despreading and demapping can acquire the full multipath frequency gain for the proposed system.

The remainder of this chapter is organized as follows: In Section II, we present the transmitted signal model of the differential parallel QPSK-FSOK MCSS system. Section III introduces the multipath fading channel and CFO effects. In Section IV, the receiver structure with the corresponding differential decoding scheme is developed. Section V illustrates simulation results. Section VI draws conclusions.

II. TRANSMITTED SIGNAL MODEL

The transmitter block diagram of the differential parallel QPSK-FSOK MCSS uplink system with K users is shown in Fig. 1. The transmitted data of each user undergoes a process consisting of QPSK-FSOK symbol mapping/spreading, initial selecting of Chu sequence, differential block encoding, data repeating, frequency shifting, IFFT, and CP insertion. The details of these procedures are described and analyzed as follows.

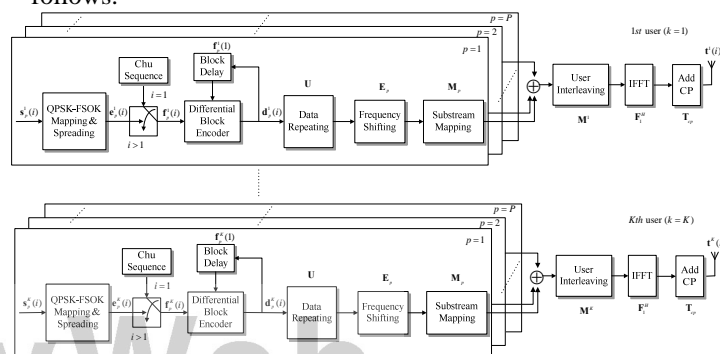


Fig. 1 Transmitter block diagram of the proposed differential parallel QPSK-FSOK MCSS system.

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A. QPSK-FSOK Symbol Mapping and Spreading

As shown in Fig. 2, assuming there are P substreams, let $\mathbf{s}_p^k(i) = [s_{p,0}^k(i) s_{p,1}^k(i) \cdots s_{p,R+1}^k(i)]^T$, $p = 1, 2, \dots, P$ denote the p th substream of the i th transmitted symbol block of the k th user, which has $R+2$ bits. The first R bits, $[s_{p,0}^k(i) s_{p,1}^k(i) \cdots s_{p,R-1}^k(i)]^T$, can be mapped into the i th FSOK symbol block in terms of a set of the $N \times N$ orthogonal FFT matrix

$$\mathbf{Q} = [\mathbf{q}_0 \mathbf{q}_1 \cdots \mathbf{q}_m \cdots \mathbf{q}_{N-1}], \text{ where } \mathbf{q}_m = \left[1 e^{-j2\pi \frac{m}{N}} \cdots e^{-j2\pi \frac{m}{N}(N-1)} \right]^T \text{ is}$$

the N -point frequency shifted sequences vector. Therefore, the first R bits can be converted into the decimal number of mapping index m , $m = 0, 1, \dots, N-1$. Then, the FSOK symbol $\mathbf{q}_{m_{p,i}^k}$ of the p th substream of the i th transmitted symbol block of the k th user can be selected from \mathbf{Q} . Next, the last two bits $[s_{p,R}^k(i) s_{p,R+1}^k(i)]^T$ can be mapped into QPSK symbol $d_p^k(i) = s_{p,R}^k(i) + j \cdot s_{p,R+1}^k(i)$. Finally, the hybrid QPSK-FSOK symbol block $\mathbf{e}_p^k(i)$ can be constructed by

$$\mathbf{e}_p^k(i) = \mathbf{q}_{m_{p,i}^k} d_p^k(i) \quad (1)$$

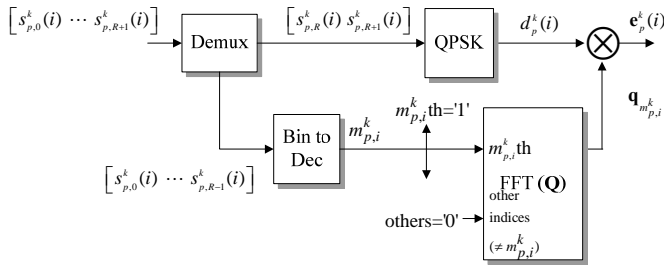


Fig. 2 Block diagram of QPSK-FSOK symbol mapping and spreading.

B. Differential Block Encoder

As shown in Fig. 1, the signal $\mathbf{f}_p^k(1)$ denotes the initial symbol block as the N -point Chu sequence

$$\mathbf{f}_p^k(1) = [c_0 \ c_1 \ \cdots \ c_n \ \cdots \ c_{N-1}]^T \quad (2)$$

where the n th chip of the Chu sequence is $c_n = e^{j\pi n^2 q/N}$, q and N are relatively prime. Next, the input of the differential encoder can be expressed as

$$\mathbf{f}_p^k(i) = \mathbf{e}_p^k(i), \quad i > 1 \quad (3)$$

Then, the output of the differential encoder can be written as

$$\begin{aligned} \mathbf{d}_p^k(i) &= \mathbf{d}_p^k(i-1) \odot \mathbf{f}_p^k(i), \quad i \geq 1 \\ &= \text{diag} \{ \mathbf{d}_p^k(i-1) \} \mathbf{f}_p^k(i) \end{aligned} \quad (4)$$

where \odot is element-wise multiplication, $\mathbf{d}_p^k(i)$ represents the modulated Chu sequence, $\text{diag} \{ \mathbf{d}_p^k(i-1) \}$ denotes the diagonal matrix of $\mathbf{d}_p^k(i-1)$, and we choose the initial

encoded block $\mathbf{d}_p^k(0)$ as the all-one sequence.

C. Resource Allocation and Multi-Carrier Modulation

As shown in Fig. 1, let the differential encoded symbol of each substream repeated P times. Then, the repeated substream can be multiplied by the frequency shift. Following the resource allocation, CP insertion and IFFT, the base-band transmission signal can be formulated

$$\mathbf{t}^k(i) = \mathbf{T}_{cp} \mathbf{F}_1^H \mathbf{M}^k \sum_{p=1}^P \mathbf{M}_p \mathbf{E}_p \mathbf{U} \mathbf{d}_p^k(i) \quad (5)$$

where $\mathbf{U} = [\mathbf{I}_N \cdots \mathbf{I}_N]^T$ is a $NP \times N$ repeating matrix formed by the $N \times N$ identity matrix of \mathbf{I}_N , which lets the input symbol $\mathbf{d}_p^k(i)$ repeated P times; $\mathbf{E}_p = \text{diag} \left\{ 1, e^{-j2\pi \frac{p}{NP}}, \dots, e^{-j2\pi \frac{p(NP-1)}{NP}} \right\}$ denotes the diagonal frequency shift matrix; $\mathbf{T}_{cp} = [\mathbf{I}_{N_g \times N_1}^T, \mathbf{I}_{N_1}^T]^T$ is a matrix with adding a cyclic prefix, where $\mathbf{I}_{N_g \times N_1}$ denotes the last N_g columns of \mathbf{I}_{N_1} . Then, \mathbf{F}_1^H is the IFFT with matrix size $N_1 \times N_1$. Then, \mathbf{M}_p is the resource allocation matrix; Therefore, the (m,n) th element of \mathbf{M}_p can be expressed as

$$\mathbf{M}_p(m, n) = \begin{cases} 1, & m = (p-1) + Pn, 0 \leq n \leq NP-1 \\ 0, & \text{otherwise} \end{cases} \quad (6)$$

for $p = 1, 2, \dots, P$. Next, in Eq. (5), \mathbf{M}^k denotes the interleaved subcarrier allocation matrix for the k th user

$$\mathbf{M}^k(m, n) = \begin{cases} 1, & m = (k-1) + Kn, 0 \leq n \leq NP^2-1 \\ 0, & \text{otherwise} \end{cases} \quad (7)$$

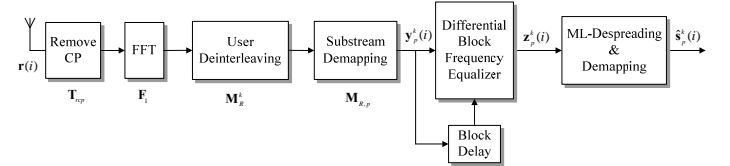


Fig. 3 Receiver block diagram of the proposed parallel QPSK-FSOK MCSS system.

III. CHANNEL AND RECEIVED SIGNAL MODEL

For mobile transmission with a reasonably short packet, the channel response $\mathbf{h}^k(i) = [h_0^k(i), h_1^k(i), \dots, h_{L-1}^k(i)]^T$ can be regarded as invariant within two consecutive blocks. The l th channel tap $h_l^k(i)$ of the i th time instant of the k th user is a time-varying Gaussian complex fading process. The timing of received signal is assumed to be perfect synchronization. Next, the base-band received signal in the presence of CFO with K users can be expressed as

$$\mathbf{r}(i) = \sum_{k=1}^K \mathbf{G}^k(i) \mathbf{H}^k(i) \mathbf{t}^k(i) + \mathbf{w}(i) \quad (8)$$

where $\mathbf{H}^k(i)$ is the $(N_1 + N_g) \times (N_1 + N_g)$ channel matrix of i th received block of the k th user, and $\mathbf{w}(i)$ denotes an $(N_1 + N_g) \times 1$ uncorrelated noise vector with zero mean and variance σ_n^2 . Then, the diagonal CFO matrix $\mathbf{G}^k(i)$ of k th user can be expressed as

$$\mathbf{G}^k(i) = g(i)\mathbf{G}(\Delta f_k) \quad (9)$$

where $g(i) = e^{j2\pi\Delta f_k(i-1)(N_g+N_1)T_s}$ is the initial phase of the i th block of CFO, $\mathbf{G}(\Delta f_k) = \text{diag}\{1, e^{j2\pi\Delta f_k T_s}, \dots, e^{j2\pi\Delta f_k(N_g+N_1-1)T_s}\}$ denotes the CFO matrix with T_s being the sampling period, and Δf_k denotes the unknown CFO of user k between the transmitter and receiver.

IV. PROPOSED RECEIVERS OF DIFFERENTIAL PARALLEL QPSK-FSOK MCSS SYSTEM

A. Frequency Domain Deinterleaving and Demapping

As shown in Fig. 3, the received signal after removing CP, FFT, user deinterleaving and substream demapping can be formulated

$$\mathbf{y}_p^k(i) = \mathbf{M}_{R,p} \mathbf{M}_R^k \mathbf{F}_1 \mathbf{T}_{rcp} \mathbf{r}(i) \quad (10)$$

where $\mathbf{T}_{rcp} = [\mathbf{0}_{N_1 \times N_g}, \mathbf{I}_{N_1}]$ is a matrix that removes the CP; \mathbf{M}_R^k denotes the subcarrier deinterleaving for the k th user, i.e., $\mathbf{M}_R^k = \mathbf{M}^{kT}$. $\mathbf{M}_{R,p}$ is the demapping matrix for the parallel substreams. Then, substituting $\mathbf{t}^k(i)$, $\mathbf{r}(i)$ and $\mathbf{G}^k(i)$ in Eqs. (5),(8),(9) into Eq. (10) yields the received signal of the k th user as

$$\begin{aligned} \mathbf{y}_p^k(i) &= \mathbf{M}_{R,p} \bar{\mathbf{\Lambda}}^k(i) \mathbf{M}_p \mathbf{E}_p \mathbf{U} \mathbf{d}_p^k(i) + \bar{\mathbf{w}}_p^k(i) \\ &= \bar{\mathbf{\Lambda}}_p^k(i) \mathbf{E}_p \mathbf{U} \mathbf{d}_p^k(i) + \bar{\mathbf{w}}_p^k(i) \end{aligned} \quad (11)$$

where $\bar{\mathbf{\Lambda}}^k(i)$ is a composite diagonal matrix. The composite noise $\bar{\mathbf{w}}_p^k(i)$ with interference terms can be expressed as

$$\begin{aligned} \bar{\mathbf{w}}_p^k(i) &= \mathbf{M}_{R,p} \mathbf{M}_R^k \tilde{\mathbf{\Lambda}}_{ICI}^k(i) \mathbf{M}^k \sum_{p=1}^P \mathbf{M}_p \mathbf{E}_p \mathbf{U} \mathbf{d}_p^k(i) \\ &+ \sum_{j=1, j \neq k}^K \mathbf{M}_{R,p} \mathbf{M}_R^j \tilde{\mathbf{\Lambda}}_{ICI}^j(i) \mathbf{M}^j \sum_{p=1}^P \mathbf{M}_p \mathbf{E}_p \mathbf{U} \mathbf{d}_p^j(i) + \mathbf{w}_p^k(i) \end{aligned} \quad (12)$$

where $\tilde{\mathbf{\Lambda}}_{ICI}^k(i)$ is the ICI part of the composited channel $\tilde{\mathbf{\Lambda}}^k(i)$, i.e., $\tilde{\mathbf{\Lambda}}^k(i) = \tilde{\mathbf{\Lambda}}_{diag}^k(i) + \tilde{\mathbf{\Lambda}}_{ICI}^k(i)$ in which $\tilde{\mathbf{\Lambda}}^k(i)$ can be expressed by $\tilde{\mathbf{\Lambda}}^k(i) = \mathbf{F}_1 \mathbf{T}_{rcp} g(i) \mathbf{G}(\Delta f_k) \mathbf{H}^k(i) \mathbf{T}_{cp} \mathbf{F}_1^H$.

B. Differential Block Decoder

To simplify the formula of differential block decoder, we assume the channel is fixed within two consecutive blocks and

ignore the initial phase of CFO, the composite channel matrix can be regarded as

$$\begin{aligned} \mathbf{H}^k(i-1) &= \mathbf{H}^k(i) \\ \bar{\mathbf{\Lambda}}_p^k(i-1) &= \bar{\mathbf{\Lambda}}_p^k(i) \end{aligned} \quad (13)$$

Based on the assumption of Eq. (13) and MRC concept, the simple and efficient differential decoding scheme can be obtained by applying the conjugate product of the $(i-1)$ th and i th received data blocks. The differential decoded block data of the MRC decoder can be expressed as

$$\begin{aligned} \mathbf{z}_p^k(i) &= \mathbf{y}_p^k(i) \odot \mathbf{y}_p^k(i-1) \\ &= \hat{\mathbf{\Lambda}}_p^k(i) \mathbf{U} \mathbf{q}_{m_p,i}^k \mathbf{d}_p^k(i) + \tilde{\mathbf{w}}_p^k(i) \end{aligned} \quad (14)$$

where $\hat{\mathbf{\Lambda}}_p^k(i) = \bar{\mathbf{\Lambda}}_p^k(i-1) \bar{\mathbf{\Lambda}}_p^k(i)$ is a $NP \times NP$ diagonal matrix with absolute value.

C. QPSK-FSOK ML Despreading and Demapping

After the MRC differential block decoder in Eq. (14), the decoded block data can be despreading by applying ML rule

$$\begin{aligned} \tilde{z}_{p,m}^k(i) &= (\mathbf{U} \mathbf{q}_m)^H \mathbf{z}_p^k(i) \\ &= \mathbf{q}_m^H \hat{\mathbf{\Lambda}}_p^k(i) \mathbf{q}_{m_p,i}^k \mathbf{d}_p^k(i) + \tilde{\mathbf{w}}_p^k(i) \end{aligned} \quad (15)$$

where $\tilde{\mathbf{w}}_p^k(i) = (\mathbf{U} \mathbf{q}_m)^H \tilde{\mathbf{w}}_p^k(i)$, $\hat{\mathbf{\Lambda}}_p^k(i) = \mathbf{U}^H \hat{\mathbf{\Lambda}}_p^k(i) \mathbf{U}$ is a $N \times N$ diagonal matrix. Then, we describe the QPSK-FSOK symbol detection and demapping operation. The decision $R+2$ bits can be expressed as $[\hat{s}_{p,0}^k(i), \hat{s}_{p,1}^k(i), \dots, \hat{s}_{p,R+1}^k(i)]$. Then, a two-step QPSK-FSOK symbol detection scheme can be obtained by applying the ML rule. First, to determine the maximum index of the FSOK sequence, the first R bits $[\hat{s}_{p,0}^k(i), \dots, \hat{s}_{p,R-1}^k(i)]$ can be detected by

$$\begin{aligned} \hat{m} &= \arg \max_m \{ |\text{Re}(\tilde{z}_{p,m}^k(i))| + |\text{Im}(\tilde{z}_{p,m}^k(i))| \} \\ [\hat{s}_{p,0}^k(i), \dots, \hat{s}_{p,R-1}^k(i)] &= \text{dec2bin}\{\hat{m}\}, \quad 0 \leq m \leq N-1 \end{aligned} \quad (16)$$

where the function ‘‘dec2bin’’ denotes the conversion of the unsigned decimal number into binary digits. Note that if \hat{m} is a correct decision in terms of the frequency shift, i.e., $\hat{m} = m_{p,i}^k$. The Next step is a simple QPSK slicer to detect the other two bits $[\hat{s}_{p,R}^k(i), \hat{s}_{p,R+1}^k(i)]$ for the maximum \hat{m} th despreading data in Eq. (16), i.e.,

$$\begin{aligned} \hat{s}_{p,R}^k(i) &= \text{dec}\{\text{Re}(\tilde{z}_{p,\hat{m}}^k(i))\} \\ \hat{s}_{p,R+1}^k(i) &= \text{dec}\{\text{Im}(\tilde{z}_{p,\hat{m}}^k(i))\} \end{aligned} \quad (17)$$

where $\text{dec}\{\cdot\}$ denotes the decision operation. Thus, based on Eqs. (16) and (17), the i th QPSK-FSOK symbol of the p th substream can be detected. Moreover, the ML detector can be equally applied to detect the other substream of each user.

V. SIMULATION RESULTS

In this section, simulation results are demonstrated to confirm the performance of the proposed differential parallel QPSK-FSOK MCSS system. In all the Monte-Carlo simulation, the system bandwidth is 10 MHz, the center carrier is set at 2.5 GHz, the mobile fading channel is simulated by Jake's model [10]. The independent Rayleigh fading paths (L) are generated with delays from $[0 \sim 7 T_s]$, where T_s is the sample duration and the delay interval is smaller than the guard interval $N_g = 32$. The length of FSOK symbol sequences is $N = 32$.

In the first simulation, Fig. 4 shows the BER performance of the proposed differential parallel QPSK-FSOK MCSS system over mobile velocity of 300 Km/Hr fading channel with different multipath number $L=2, 4, 6, 8$. The number of substream $P=2$ and number of users $K=4$ are set, and the normalized carrier frequency offset of each user is assumed $\Delta f_k T_s = 0.2$. In Fig. 4, the frequency diversity gain and the M -ary gain make the proposed system superior to the theoretical QPSK fading system. As L increases, the frequency diversity gain improves significantly.

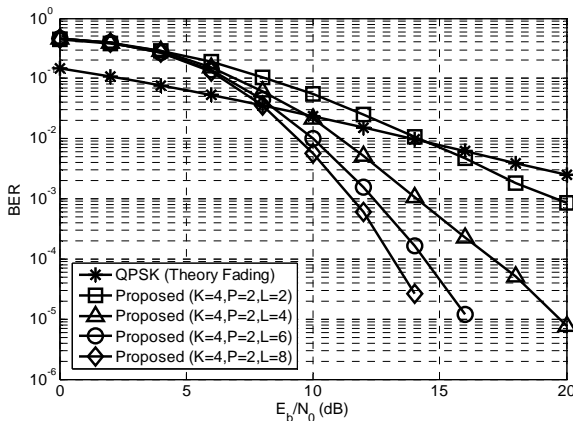


Fig. 4 BER performance versus different multipath number L .

Finally, we evaluate the PAPR property of the transmitted signal samples interpolated by an oversampling ratio of 4. The PAPR (in dB) is defined as:

$$\text{PAPR}_{\text{dB}} = 10 \log_{10} \left(\frac{\max \{|x_k|^2\}}{\langle |x_k|^2 \rangle} \right) \quad (18)$$

where x_k is the k th interpolated sample, and $\langle \bullet \rangle$ denotes the time-average operation. For the proposed differential systems, the CCDF of the PAPR is plotted in Fig. 5. The number of user $K=4$ and number of substream $P=2$ are set. It is observed that the proposed differential systems exhibit a much lower PAPR performance, as shown in Eq. (5), than the conventional WH MC-SS system. Thus the proposed system is proved to be much less demanding in power amplifier linearity.

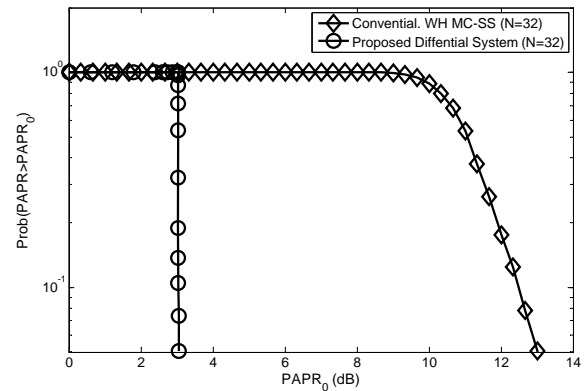


Fig. 5 The PAPR CCDF for the proposed system.

VI. CONCLUSIONS

In this paper, we have proposed a new differential parallel QPSK-FSOK MCSS system over fast changing multipath fading channels. Based on the transceiver design, the ML rule is employed for symbol demapping and decision. Finally, simulation results confirm the proposed low-PAPR differential system robust against the CFO effect and outperform the theoretical QPSK fading system due to the frequency diversity gain and M -ary gain.

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