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A Square-Constellation-Based M -Ary DCSK Communication System

GUOFA CAI¹, YI FANG¹, (Member, IEEE), GUOJUN HAN¹, (Senior Member, IEEE), FRANCIS C. M. LAU², (Senior Member, IEEE), AND LIN WANG³, (Senior Member, IEEE)

¹School of Information Engineering, Guangdong University of Technology, Guangzhou 510006, China

²Department of Electronic and Information Engineering, The Hong Kong Polytechnic University, Hong Kong

³Department of Communications Engineering, Xiamen University, Xiamen 361005, China

Corresponding author: Y. Fang (fangyi@gdut.edu.cn)

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ABSTRACT In this paper, a square-constellation-based M -ary differential chaos-shift-keying (M -DCSK) communication system framework is proposed. The symbol-error-rate expression of the proposed system is derived over additive white Gaussian noise as well as multipath Rayleigh fading channels. Moreover, the peak-to-average-power-ratio (PAPR) performance is carefully analyzed for the newly proposed system. Based on the square-constellation-based M -DCSK framework, a simple least-square estimator is designed to demodulate the information exploiting pilot-assistant symbols. Analytical and simulated results show that the proposed system benefits from a lower energy consumption, but suffers from a higher PAPR compared with the circle-constellation-based M -DCSK system.

INDEX TERMS M -ary differential chaos-shift-keying (M -DCSK), symbol error rate, peak-to-average-power ratio (PAPR), multipath Rayleigh fading channel.

I. INTRODUCTION

Differential chaos-shift-keying (DCSK) [1] has drawn much attention in the past two decades because it is more robust to multipath fading channels than differential phase shift keying (DPSK) and can be applied to ultra-wideband (UWB) systems [2], [3]. DCSK systems do not require a chaotic synchronization and a channel estimator at the receiver. At the same time they offer excellent performance over multipath fading channels [4]. Performance analysis on DCSK systems has been studied in a variety of transmission scenarios over multipath fading channels, such as point-to-point scenario [5], cooperative scenario [6], [7], and multi-input multi-output (MIMO) scenario [8], [9].

Noticeably, the traditional DCSK system has two drawbacks: i) half of the bit duration is spent on sending non-information-bearing reference signal, which results in lower data rate and energy efficiency; and ii) a radio frequency (RF) delay line is needed at both the transmitter and receiver, and is very difficult to be realized in practical applications. To address the RF-delay-line problem at the receiver, a code-shifted DCSK (CS-DCSK) system was proposed

in [10] by using Walsh codes. Moreover, two improved versions of the high spectral efficiency DCSK system were presented in [11] and [12]. Recently, a new multicarrier DCSK (MC-DCSK) modulation [13] together with its analog network coding counterpart [14] were proposed, which not only can provide a parallel transmission without the need of the RF delay line at both the transmitter and receiver ends, but also can dramatically save energy and increase the spectral efficiency. In order to further boost the spectral and energy efficiency, a new multiresolution M -ary DCSK (MR M -DCSK) modulation based on the principle of circle constellation was proposed in [15], which can support different kinds of data services for different bit error rate (BER) requirements and can be easily extended to MC version. As an extension of quadrature chaos shift keying (QCSK) system [16], the circle-constellation-based M -DCSK system enables more desirable error performance than the M -PSK-DCSK system [17].

However, when a different frequency selective fading channel corresponding to each subcarrier channel is considered, the schemes in [13]–[15] become inappropriate.

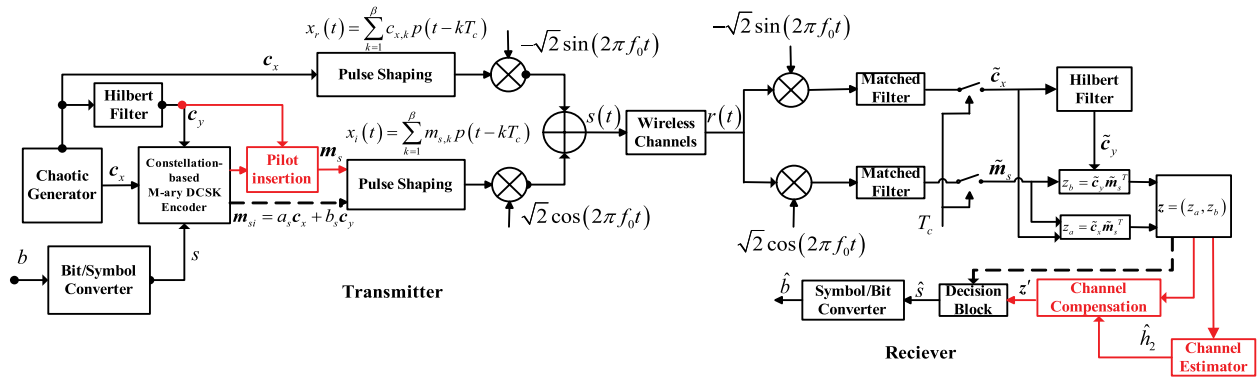


FIGURE 1. Block diagram of a generalized constellation-based M -DCSK communication system.

Orthogonal frequency division multiplexing DCSK (OFDM-DCSK) system was then formulated to mitigate this drawback [18] and a phase-separated DCSK system in [19] can well avoid this problem. Moreover, although M -ary DCSK based on Walsh codes [20] and its improved version [21] can obtain better performance, the systems require RF delay lines and larger bandwidth. Based on Walsh codes and Hilbert transform, an orthogonal multi-level DCSK system without any RF delay line at the receiver was developed in [22], where the performance is very promising with an increasing level M . The main drawback of the orthogonal multi-level DCSK system is that it requires larger bandwidth. Although the circle-constellation-based M -DCSK system is a bandwidth-efficient system, the system performance decreases as the level M becomes larger.

With the aforementioned motivations, we propose a square-constellation-based M -DCSK communication system framework in this paper in order to achieve a better performance while keeping the bandwidth-efficiency property unchanged. Furthermore, pilot-assistant symbols are inserted into the information symbols every P symbol intervals. Then a simple least-square (LS) estimator is designed for the proposed system to demodulate the information over multipath fading channels with the assistance of the pilot-assistant symbols. The proposed system framework can be easily generalized to other constellations, i.e., rectangular, cross, and star constellations. Besides, the performance of the proposed system is analyzed and its advantages and disadvantages are discussed. In summary, the major contributions of this work are described as follows.

- 1) A new square-constellation-based M -DCSK communication system framework is proposed, which can be generalized to other constellations, i.e., rectangular, cross and star constellations. In order to address the demodulation problem under fading conditions, a simple LS estimator assisted by pilot-assistant symbols, is conceived for the new system.
- 2) The symbol-error-rate (SER) expression of the new system is derived and analyzed over AWGN as well as multipath Rayleigh fading channels. Both the theoretical results

and simulation results show that the square-constellation-based M -DCSK system accomplishes a notable performance gain over the circle-constellation-based M -DCSK system under the same average power.

- 3) The peak-to-average-power ratio (PAPR) performance of the new system is intensely analyzed. Results show that the new system has a higher PAPR with respect to the circle-constellations-based M -DCSK system.

The remainder of this paper is organized as follows. The generalized constellation-based M -DCSK communication system framework is proposed in Section II. The performance analysis of square-constellation-based M -DCSK system is reported in Section III. Main results and discussions are provided in Section IV. Finally, some concluding remarks are given in Section V.

II. GENERALIZED CONSTELLATION-BASED M -DCSK COMMUNICATION SYSTEM FRAMEWORK

In this section, we briefly introduce the generalized constellation-based M -DCSK communication system shown in Fig. 1. If circle constellations are adopted, the red-color part is removed. For square, rectangular, cross, and star constellations, a channel-estimator module is required. Here, circle and square constellations are adopted as two instances.¹

A. CHAOTIC GENERATOR

In this paper, the logistic map, i.e.,

$$c_{k+1} = 1 - 2c_k^2, \tag{1}$$

¹It should be noted that, although the square-constellation-based DCSK system needs a channel estimator, its BER performance is much better than that of the circle-constellation-based DCSK system (see Section IV). In other words, the square-constellation-based DCSK system is more efficient than the circle-constellation-based one in terms of energy consumption. Moreover, other types of constellations (e.g., rectangular, cross, and star) can also be used in the generalized system framework, thus offering a broader application prospect. Besides, based on the multiresolution M -DCSK framework in [15], a multiresolution square-constellation-based M -DCSK system can be devised, which provides a more flexible BER requirement by adjusting the distance d because both the amplitude and phase carry message in this case.

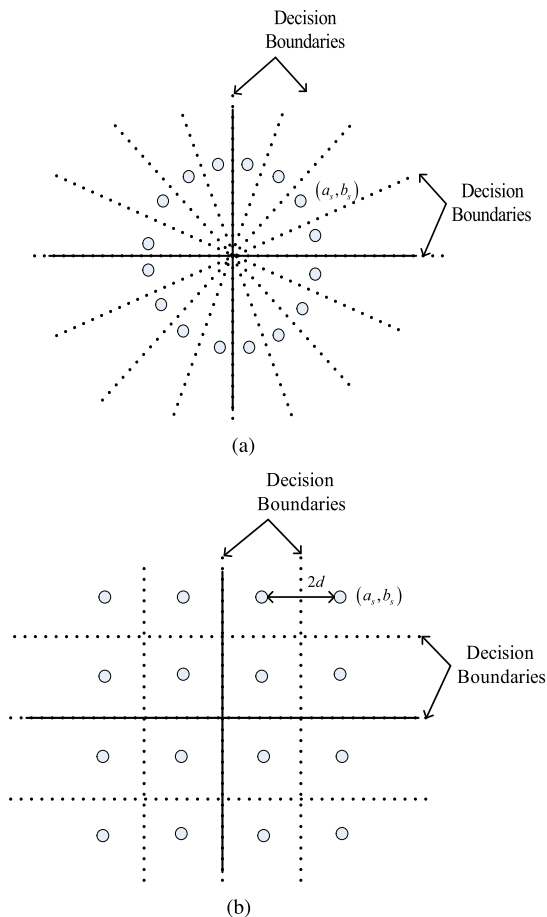


FIGURE 2. Constellations of the (a) circle and (b) square 16-DCSK, and their decision boundaries.

is employed where c_k ($k = 1, 2, \dots$) denotes the k -th chip of a chaotic signal. Thereby, two independent chaotic signals (or even different segments of the same chaotic waveform) each of length β can be generated by (1). Here β is defined as the spreading factor. These two chaotic signals have a very low cross-correlation value for a large spreading factor (for details see [9]). For a normalized chaotic signal with zero mean, the variance is equal to one.

B. TRANSMITTER

A chaotic reference signal $\mathbf{c}_x = (c_{x,1}, \dots, c_{x,\beta})$ is generated by the chaotic generator. In order to obtain two independent orthogonal chaotic signals, the chaotic reference signal is transformed into another quadrature signal $\mathbf{c}_y = (c_{y,1}, \dots, c_{y,\beta})$ by using the Hilbert transform [15], [16], [22], where the Hilbert transform is easily implemented by means of the finite-impulse-response (FIR) filter. Thus we have $\sum_{k=1}^{\beta} c_{x,k}c_{y,k} = 0$.

For constellation-based M -DCSK, a circle 16-DCSK and a square 16-DCSK constellation ($2d$ is the distance between adjacent points.) are adopted as two instances and shown in Fig. 2. Moreover, the encoded information is denoted by $\mathbf{m}_s = a_s\mathbf{c}_x + b_s\mathbf{c}_y$, where the index s ($s = 1, \dots, M$)

presents the symbol value, and a_s and b_s denote the corresponding x-axis coordinate value and y-axis coordinate value, respectively, of the constellation point. If circle constellation is adopted, one has $a_s^2 + b_s^2 = 1$ for all s . On the other hand, if square constellation is applied, the above equality will not hold.

When the chaotic signals are passed through the pulse shaping block, where a square-root-raised-cosine filter $p(t)$ is adopted, the reference and information signals are written, respectively, as $x_r(t) = \sum_{k=1}^{\beta} c_{x,k}p(t - kT_c)$ and $x_i(t) = \sum_{k=1}^{\beta} m_{s,k}p(t - kT_c)$, where T_c is the chip duration and $\int_{kT_c}^{(k+1)T_c} p^2(t - kT_c) dt = E_p$. Moreover, all chaotic segments \mathbf{c}_x and \mathbf{c}_y have been normalized to 1, and hence $\sum_{k=1}^{\beta} c_{x,k}^2 = \sum_{k=1}^{\beta} c_{y,k}^2 = 1$. Finally, the transmitted signal of constellation-based M -DCSK system can be written as

$$s(t) = \sqrt{2}x_i(t) \cos(2\pi f_0 t) - \sqrt{2}x_r(t) \sin(2\pi f_0 t), \quad (2)$$

where f_0 is the frequency of the sinusoidal carriers and is assumed to be a multiple of $1/T_c$ satisfying $f_0 \gg 1/T_c$.

C. CHANNEL MODEL

Without loss of generality, we consider a multipath slow Rayleigh-fading channel, which is commonly used in spread-spectrum systems [13], [15]. The impulse response function of such a channel model can be defined as

$$h(t) = \sum_{l=1}^L \alpha_l \delta(t - \tau_l), \quad (3)$$

where L denotes the number of paths, τ_l and α_l represent the time delay and the channel coefficient of the l -th path, respectively.

In such a channel model, the probability density function (PDF) of the channel coefficient α_l is given by [23]

$$f(\alpha_l) = \frac{\alpha_l}{\sigma^2} \exp\left(-\frac{\alpha_l^2}{2\sigma^2}\right), \quad \alpha_l > 0, \quad (4)$$

where $\sigma > 0$ is the scale parameter of the distribution. Particularly, the channel is reduced to the AWGN channel when the number of paths and channel coefficient are respectively set to $L = 1$ and $\alpha_1 = 1$.

D. RECEIVER

When the transmitted signals are passed through the wireless channels, the signal at the receiver is given by

$$r(t) = h(t) \otimes s(t) + n(t), \quad (5)$$

where \otimes is the convolution operator, and $n(t)$ is a wideband AWGN with zero mean and variance $N_0/2$.

The received signal is then separated into the in-phase and quadrature-phase components which are further passed through two corresponding matched filters. Subsequently, two filtered signals are obtained and they will be sampled every T_c time unit. At this step, one obtains the desired signals, i.e., chaotic reference signal and chaotic information

signal denoted by $\tilde{\mathbf{c}}_x$ and $\tilde{\mathbf{m}}_s$, respectively. The chaotic reference signal $\tilde{\mathbf{c}}_x$ is also transformed into another quadrature reference signal $\tilde{\mathbf{c}}_y$ by using the Hilbert transform. To demodulate the received symbol, the decision vector $z = (z_a, z_b)$ where $z_a = \tilde{\mathbf{c}}_x \tilde{\mathbf{m}}_s^T$ and $z_b = \tilde{\mathbf{c}}_y \tilde{\mathbf{m}}_s^T$ is computed.

E. CHANNEL ESTIMATOR AND DECISION

For circle-constellation-based M -DCSK system, the decision vector is made according to the decision boundaries shown in Fig. 2(a). However, other-constellations-based M -DCSK system requires a channel estimator. In the proposed system, one should only attain the channel power value defined in [24] rather than acquiring every channel coefficient, which is different from the conventional channel estimation in coherent communication systems. We assume that the estimated channel state information is \hat{h}_2 . Then, the normalized decision vector becomes $z' = \frac{z}{E_p \hat{h}_2}$ which will be used to decode the received symbol according to the decision boundaries shown in Fig. 2(b).

III. PERFORMANCE ANALYSIS OF SQUARE-CONSTELLATION-BASED M -DCSK SYSTEM AND CHANNEL ESTIMATION

In this section, we firstly derive the transmission energy and SER expression of the square-constellation-based M -DCSK communication system, where both AWGN and multipath Rayleigh fading channels are considered. Afterwards, the PAPR performance of the corresponding system is analyzed. Finally, a simple channel estimator is designed for the proposed system. In the following, the square-constellation-based M -DCSK system is referred to as S- M -DCSK system while circle-constellation-based M -DCSK system is referred to as C- M -DCSK system for simplicity.

A. TRANSMISSION ENERGY

Here, we briefly analyze the average symbol energy for the S- M -DCSK system. To begin with, the transmission energy for the transmitted signal $s(t)$ is computed as

$$\begin{aligned}
 E_T &= \int_{(k-1)\beta T_c}^{k\beta T_c} s^2(t) dt \\
 &= \int_{(k-1)\beta T_c}^{k\beta T_c} \left[x_i^2(t) + x_r^2(t) + x_i^2(t) \cos(4\pi f_0 t) \right. \\
 &\quad \left. - x_r^2(t) \cos(4\pi f_0 t) - 2x_i(t)x_r(t) \sin(4\pi f_0 t) \right] dt \\
 &= \int_{(k-1)\beta T_c}^{k\beta T_c} \left[x_i^2(t) + x_r^2(t) \right] dt \\
 &= \sum_{i=(k-1)\beta}^{k\beta-1} \left(m_{s,i}^2 + c_{x,i}^2 \right) \int_{iT_c}^{(i+1)T_c} p^2(t - iT_c) dt \\
 &= \left(a_s^2 + b_s^2 + 1 \right) E_p. \tag{6}
 \end{aligned}$$

In the M -order ($M = 4, 16, 64, \dots$) square constellation, the amplitudes in both axes are $\pm d, \pm 3d,$

$\pm 5d, \dots, \pm(\sqrt{M}-1)d$. Thus, the average energy per symbol (i.e., E_s) can be calculated as

$$\begin{aligned}
 E_s &= \frac{1}{M} \sum_{m=1}^{\sqrt{M}} \sum_{n=1}^{\sqrt{M}} \left(a_m^2 + b_n^2 + 1 \right) E_p \\
 &= \frac{E_p}{M} \times \frac{2M(M-1)}{3} d^2 + E_p \\
 &= \frac{2(M-1)d^2 + 3}{3} E_p. \tag{7}
 \end{aligned}$$

B. SER PERFORMANCE

In practical applications, the largest multipath delay is generally much shorter than the symbol duration [13], [15], i.e. $0 < \tau_{l,max} \ll \beta$. In the following analysis, we assume that this statement holds and thus the inter-symbol interference (ISI) is negligible. Since the channel is slowly fading, we also assume the channel coefficients are constant during each transmission period.

The decision vector $z = (z_a, z_b)$ can be approximated as

$$\begin{aligned}
 z_a &\approx \sum_{k=1}^{\beta} \left[\left(\sum_{l=1}^L \alpha_l \sqrt{E_p} c_{x,k-\tau_l} + n_k \right) \right. \\
 &\quad \left. \times \left(\sum_{l=1}^L \alpha_l (a_s \sqrt{E_p} c_{x,k-\tau_l} + b_s \sqrt{E_p} c_{y,k-\tau_l}) + n_{k+\beta} \right) \right], \tag{8}
 \end{aligned}$$

$$\begin{aligned}
 z_b &\approx \sum_{k=1}^{\beta} \left[\left(\sum_{l=1}^L \alpha_l \sqrt{E_p} c_{y,k-\tau_l} + \tilde{n}_k \right) \right. \\
 &\quad \left. \times \left(\sum_{l=1}^L \alpha_l (a_s \sqrt{E_p} c_{x,k-\tau_l} + b_s \sqrt{E_p} c_{y,k-\tau_l}) + n_{k+\beta} \right) \right]. \tag{9}
 \end{aligned}$$

For a large spreading factor, one has [13], [15]

$$\sum_{k=1}^{\beta} c_{x,k-\tau_i} c_{x,k-\tau_j} \approx 0, \quad i \neq j. \tag{10}$$

Then, define the following approximated expression by

$$E[a_s^2 + b_s^2 + 1] E_p \approx E_s, \tag{11}$$

where $E[\cdot]$ denotes the expectation operator.

By exploiting (10), (8) and (9) can be respectively reduced to

$$\begin{aligned}
 z_a &\approx \sum_{l=1}^L \alpha_l^2 a_s E_p + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_p} c_{x,k-\tau_l} n_{k+\beta} \\
 &\quad + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_p} (a_s c_{x,k-\tau_l} + b_s c_{y,k-\tau_l}) n_k + \sum_{k=1}^{\beta} n_{k+\beta} n_k, \tag{12}
 \end{aligned}$$

$$\begin{aligned}
 z_b &\approx \sum_{l=1}^L \alpha_l^2 b_s E_p + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_p} c_{y,k-\tau_l} n_{k+\beta} \\
 &\quad + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_p} (a_s c_{x,k-\tau_l} + b_s c_{y,k-\tau_l}) \tilde{n}_k + \sum_{k=1}^{\beta} n_{k+\beta} \tilde{n}_k. \tag{13}
 \end{aligned}$$

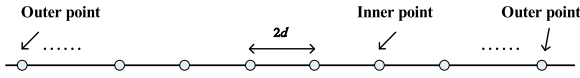


FIGURE 3. Each row/column of the square M -DCSK constellation.

Since the chaotic sequence and the Gaussian noise are independent, the means of z_a and z_b can be calculated in a similar way as [15], resulting in

$$E[z_a] = a_s E_p \sum_{l=1}^L \alpha_l^2, \quad (14)$$

$$E[z_b] = b_s E_p \sum_{l=1}^L \alpha_l^2. \quad (15)$$

For an S- M -DCSK encoder, the variances of z_a and z_b can be easily obtained by using (11), i.e.,

$$\begin{aligned} \text{var}[z_a] &= \sum_{l=1}^L \alpha_l^2 E[a_s^2 + b_s^2 + 1] E_p \frac{N_0}{2} + \frac{\beta N_0^2}{4} \\ &\approx \sum_{l=1}^L \alpha_l^2 \frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}, \end{aligned} \quad (16)$$

$$\text{var}[z_b] \approx \sum_{l=1}^L \alpha_l^2 \frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}, \quad (17)$$

where $\text{var}[\cdot]$ denotes the variance operator.

1) AWGN CHANNEL

For the AWGN case, the number of paths and channel coefficient α_l are equal to one. The mean and variance of the decision variables are, respectively, given as

$$\begin{cases} E[z'_a] = a_s \\ E[z'_b] = b_s \end{cases} \quad (18)$$

and

$$\text{var}[z'_a] = \text{var}[z'_b] = \frac{\frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}}{E_p^2} \quad (19)$$

Next, we consider the constellation shown in Fig. 3, which is the same as \sqrt{M} -ary PAM (pulse amplitude modulation). For this constellation, there exists two types of points, i.e., $\sqrt{M} - 2$ inner points and 2 outer points. If an inner point is transmitted, there is an error in detection if $|N| > d$, where, N is a zero-mean Gaussian random variable with variance $\frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}$. If an outer point is transmitted, the probability of error is one half of the error probability of an inner point because noise can cause error only in one direction. Therefore, for the inner points, one has

$$P_{ei} = \Pr(|N| > d) = 2Q\left(\frac{dE_p}{\sqrt{\frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}}}\right), \quad (20)$$

where $\Pr(\cdot)$ is the probability of an incorrect symbol detection and $Q(x) = (1/\sqrt{2\pi}) \int_x^\infty \exp(-t^2/2) dt$ for $x \geq 0$. On the other hand, for the outer points, one has

$$P_{eo} = \frac{P_{ei}}{2} = Q\left(\frac{dE_p}{\sqrt{\frac{E_s N_0}{2} + \frac{\beta N_0^2}{4}}}\right). \quad (21)$$

With the help of Eqs. (7), (20) and (21), the SER for this constellation is given by

$$\begin{aligned} P_{e,\sqrt{M}} &= \frac{1}{\sqrt{M}} [(\sqrt{M} - 2)P_{ei} + 2P_{eo}] \\ &= \frac{2(\sqrt{M} - 1)}{\sqrt{M}} Q\left(\frac{6d}{\sqrt{2\gamma_s + \beta}} \gamma_s\right), \end{aligned} \quad (22)$$

where $\gamma_s = E_s/N_0$ is the average symbol signal-noise-rate (SNR). Accordingly, the SER of the S- M -DCSK system is derived as

$$P_e = 1 - (1 - P_{e,\sqrt{M}})^2. \quad (23)$$

2) MULTIPATH RAYLEIGH FADING CHANNEL

Assume that perfect channel state information is obtained at the receiver. For L independent Rayleigh fading channels with equal power delay profile, the PDF of the instantaneous symbol-SNR γ_s can be written as [23, p. 854]

$$f(\gamma_s) = \frac{\gamma_s^{L-1}}{(L-1)! \bar{\gamma}_c^L} \exp\left(-\frac{\gamma_s}{\bar{\gamma}_c}\right), \quad (24)$$

where $\bar{\gamma}_c$ is the average symbol-SNR per channel, defined by

$$\bar{\gamma}_c = \left(\frac{E_s}{N_0}\right) E[\alpha_j^2] = \left(\frac{E_s}{N_0}\right) E[\alpha_l^2], \quad j \neq l, \quad (25)$$

and

$$\gamma_s = \left(\sum_{l=1}^L \alpha_l^2\right) \frac{E_s}{N_0}, \quad (26)$$

with $\sum_{l=1}^L E[\alpha_l^2] = 1$.

For dissimilar channels, the PDF of the instantaneous γ_s can be written as [23, p. 876]

$$f(\gamma_s) = \sum_{l=1}^L \frac{\pi_l}{\bar{\gamma}_l} \exp\left(-\frac{\gamma_s}{\bar{\gamma}_l}\right), \quad (27)$$

where

$$\pi_l = \prod_{j=1, j \neq l}^L \frac{\bar{\gamma}_l}{\bar{\gamma}_l - \bar{\gamma}_j}.$$

The SER expression of the S- M -DCSK system over multipath Rayleigh fading channels is expressed by

$$P_e = \int_0^{+\infty} P_{e,\sqrt{M}}(\gamma_s) f(\gamma_s) d\gamma_s, \quad (28)$$

where

$$P_{e,\sqrt{M}}(\gamma_s) = 1 - \left[1 - \frac{2(\sqrt{M} - 1)}{\sqrt{M}} Q\left(\frac{6d}{\sqrt{2\gamma_s + \beta}} \gamma_s\right)\right]^2.$$

C. PAPR PERFORMANCE

The peak-to-average-power ratio (PAPR) of the transmitted signal can be written as

$$PAPR = \frac{\max_{1 \leq k \leq \beta} \{(a_s c_{x,k} + b_s c_{y,k})^2 + c_{x,k}^2\}}{\frac{1}{\beta} \sum_{k=1}^{\beta} \{(a_s c_{x,k} + b_s c_{y,k})^2 + c_{x,k}^2\}}. \quad (29)$$

Since the PAPR of a transmitted signal is a random variable, one may consider the probability of PAPR exceeding a certain level $PAPR_0$, i.e.,

$$CCDF = Pr(PAPR > PAPR_0), \quad (30)$$

where CCDF represents the complementary cumulative distribution function (CCDF). If the CCDF curve of system A is lower than that of system B, the PAPR of the former is lower than that of the latter.

D. LS-BASED CHANNEL ESTIMATION

Since the channel is slowly fading, it is reasonable to assume that the channel coefficients keep constant during the transmission time of some symbols. Hence, a pilot symbol is inserted into the information symbols every P symbol intervals. As seen from (14) and (15), once the parameter $h_2 = \sum_{l=1}^L \alpha_l^2$ is obtained, the transmitted information can be estimated. Moreover, assume that the pilot symbol (a_q, b_q) is transmitted. Then, because of the symmetry of decision variables, one can restrict the analysis to

$$z_{aq} \approx \sum_{l=1}^L \alpha_l^2 a_q E_p + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_p} [c_{x,k-\tau_l} n_{k+\beta} + (a_q c_{x,k-\tau_l} + b_q c_{y,k-\tau_l}) n_k] + \sum_{k=1}^{\beta} n_{k+\beta} n_k. \quad (31)$$

Hence, if the LS estimator is adopted, the channel parameter is estimated as

$$\hat{h}_2 = \frac{z_{aq}}{E_p a_q}. \quad (32)$$

Similarly to [25], the LS estimator provides good estimates in high SNR regime, but in low SNR regime it suffers from having a high mean-square error. Here, the symbol (1, 0) is adopted as the pilot symbol to minimize noise.

IV. NUMERICAL RESULTS AND DISCUSSIONS

In this section, the simulated SER and BER performance of the S- M -DCSK system is performed over AWGN and Rayleigh fading channels. To be specific, the simulated SER results are used to verify the accuracy of the proposed analytical method, while the simulated BER result are used to validate the superiority of the proposed system. Here, $d = \sqrt{\frac{3}{2(M-1)}}$, $M = 4, 16, 64$, is set in simulation, where the average transmitted power of the S- M -DCSK system is the same as that of the C- M -DCSK system. Additionally, in simulation, a pilot symbol is inserted at the end of every 4 information-bearing symbols, and thus the transmission rate reduces to 4/5.

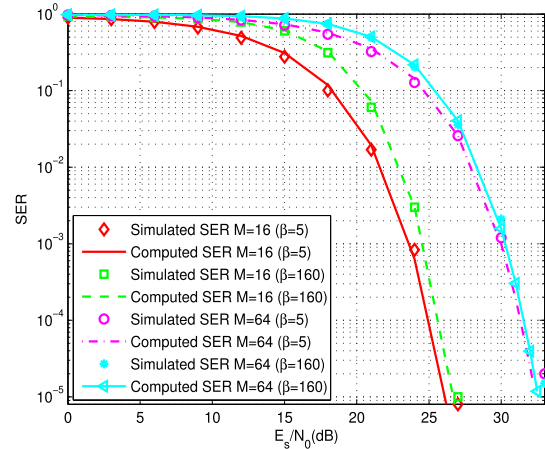


FIGURE 4. SER performance of S-16/64-DCSK system over an AWGN channel.

A. PERFORMANCE EVALUATION OF THE PROPOSED SCHEME

Fig. 4 compares the computed SER in (23) with the simulated SER of S-16/64-DCSK system over an AWGN channel, where the spreading factor β is set to 5 and 160. It can be observed that the computed SERs well match the simulated SERs for larger spreading factors, while for a small spreading factor of 5, the gap between the theoretical and simulated results is about 1dB in the high SNR regime.

In order to verify the performance of the proposed scheme over multipath Rayleigh fading channels, we assume that the receiver knows the perfect channel state information. Moreover, the simulated parameters are set as follows: number of paths $L = 3$, power delay profiles $E[\alpha_1^2] = 1/2$, $E[\alpha_2^2] = 2/5$, $E[\alpha_3^2] = 1/10$, time delays $\tau_1 = 0$, $\tau_2 = 2$, $\tau_3 = 5$, and spreading factor $\beta = 80, 160, 320$. Fig. 5 compares the computed SER in (28) with the simulated SER of S-16/64-DCSK system over a multipath Rayleigh fading channel. Referring to this figure, the computed SERs are well consistent with the simulated SERs when $M = 16$. When $M = 64$, however, there is a relative large discrepancy between the theoretical and simulation results at the high E_s/N_0 regime with $\beta = 80, 160$.

To further validate the LS estimator, the SER performance of the S-16/64-DCSK systems with perfect CSI (p-CSI) and estimated CSI over a multipath Rayleigh fading channel is shown in Fig. 6. Here, the system parameters are assumed as $E[\alpha_1^2] = E[\alpha_2^2] = E[\alpha_3^2] = 1/3$, $\tau_1 = 0$, $\tau_2 = 2$, $\tau_3 = 5$, and $\beta = 320$. As seen from this figure, the proposed LS-estimator scheme possesses a performance loss of about 1 dB as compared to the p-CSI scheme.

Fig. 7 plots the simulated SERs against the value of d over a multipath Rayleigh fading channel. The simulated parameters are set as follows: number of paths $L = 3$, power delay profiles $E[\alpha_1^2] = 1/2$, $E[\alpha_2^2] = 2/5$, $E[\alpha_3^2] = 1/10$, time delays $\tau_1 = 0$, $\tau_2 = 2$, $\tau_3 = 5$, spreading factor $\beta = 320$, $N_0 = 0.05$ and $E_p = 1$. One can observe that the SER performance is relevant to d , but it tends to be

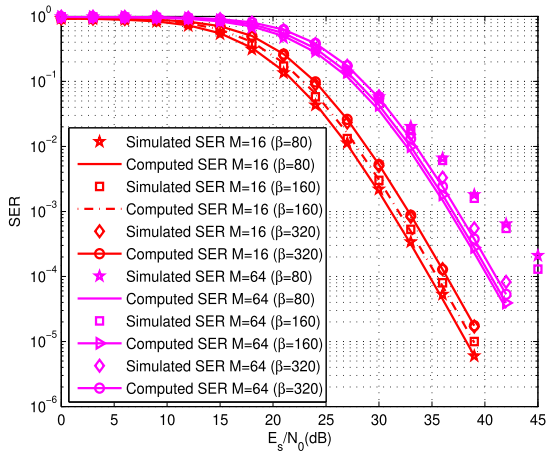


FIGURE 5. SER performance of S-16/64-DCSK system over a multipath Rayleigh fading channel.

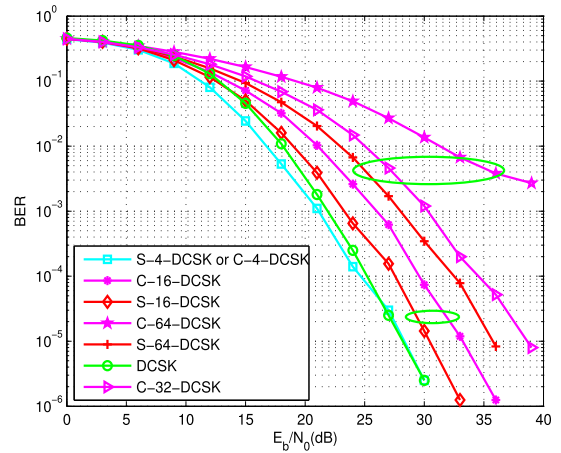


FIGURE 8. BER results of the traditional DCSK, C- M -DCSK, and S- M -DCSK systems over a multipath Rayleigh fading channel. The parameters used are $M = 4, 16, 32, 64$, and $\beta = 80$.

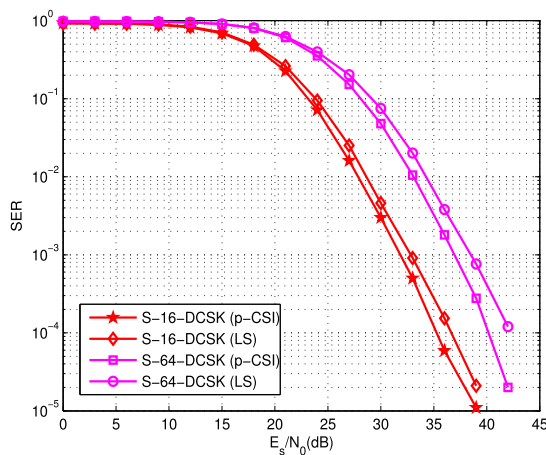


FIGURE 6. SER performance of the S-16/64-DCSK systems with perfect CSI and estimated CSI over a multipath Rayleigh fading channel.

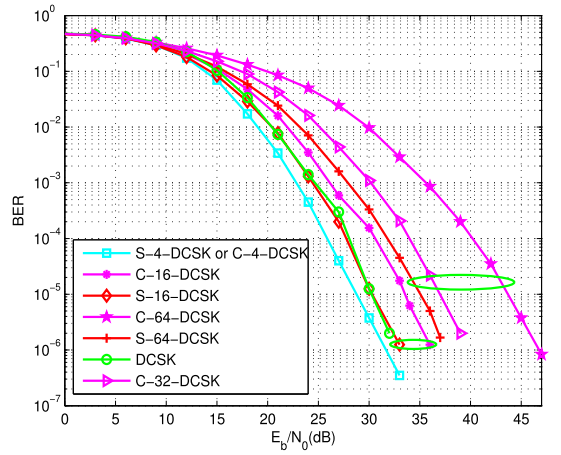


FIGURE 9. BER results of the traditional DCSK, C- M -DCSK, and S- M -DCSK systems over a multipath Rayleigh fading channel. The parameters used are $M = 4, 16, 32, 64$, and $\beta = 320$.

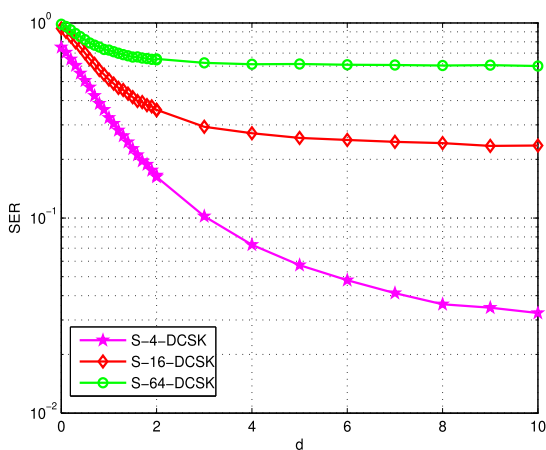


FIGURE 7. SER performance vs. the value of d over a multipath Rayleigh fading channel.

stable when d exceeds a certain threshold. Hence, one should carefully select the value of d to accomplish a good tradeoff between the performance and transmitted power. Note that,

in this paper, we choose $d = \sqrt{\frac{3}{2(M-1)}}$ ($M = 4, 16, 64$) to ensure that the transmitted power is kept the same for both S- M -DCSK and C- M -DCSK systems.

B. PERFORMANCE COMPARISON BETWEEN C- M -DCSK SYSTEM AND S- M -DCSK SYSTEM

In this subsection, we assume that the perfect CSI is not available at the receiver and hence the LS estimator is adopted. Moreover, the parameters of a multipath fading channel are assumed as $E[\alpha_1^2] = E[\alpha_2^2] = E[\alpha_3^2] = 1/3$, $\tau_1 = 0$, $\tau_2 = 2$, $\tau_3 = 5$.

Fig. 8 and Fig. 9 show the BER results of the traditional DCSK, C- M -DCSK ($M = 4, 16, 32, 64$), and S- M -DCSK ($M = 4, 16, 64$) systems over a multipath Rayleigh fading channel, where the spreading factors are set to $\beta = 80$ and 320 . In comparison with the C-16/64-DCSK systems, the S-16/64-DCSK systems somewhat decrease the transmission rate because some symbols are used in estimating the CSI. In contrast, the S-16/64-DCSK systems significantly improve the error performance as compared with

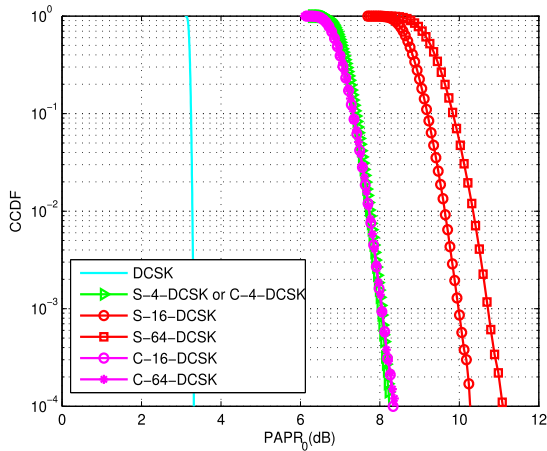


FIGURE 10. PAPR performance of the traditional DCSK, C- M -DCSK, and S- M -DCSK systems ($M = 4, 16, 64$). The spreading factor is set as $\beta = 100$.

the circle counterparts. For example, at a BER of 10^{-6} , the S-16-DCSK system achieves a gain of about 2 dB over the C-16-DCSK system. In particular, the S-64-DCSK system has a more than 10 dB gain as compared with the C-64-DCSK system and accomplishes a gain of about 2 dB over the C-32-DCSK system. It can be observed that the C-64-DCSK system is susceptible to ISI for a low spreading factor. Beside, we can also observe that the performance of the S-16-DCSK system is about 2 dB away from that of the traditional DCSK system for a low spreading factor and the gap vanishes as β becomes larger (e.g., $\beta = 320$). It should be noted that the transmission rate of the proposed S-16-DCSK system is nearly four times of the traditional DCSK system.

Fig. 10 depicts the PAPR performance of the traditional DCSK, C- M -DCSK, and S- M -DCSK systems, where the spreading factor is set to $\beta = 100$. It is observed that compared with the C-16/64-DCSK systems, the S-16/64-DCSK systems enable lower energy consumption but suffer from a higher PAPR. Furthermore, the PAPR of the S- M -DCSK system increases as M becomes larger, while that of the C- M -DCSK system is almost unchanged. Hence, the C- M -DCSK system is suitable for more efficient utilization of the power amplifiers.

V. CONCLUSIONS

In this paper, we have proposed a generalized low-complexity constellation-based M -DCSK communication system. To facilitate practical applications, we have also designed a simple LS estimator by using pilot-assistant symbols. The theoretical SER expression of the square-constellation-based M -DCSK system has been estimated over AWGN as well as multipath Rayleigh fading channels, which are reasonably consistent with the corresponding simulated results. The performance evaluation shows that the proposed system can realize lower energy consumption in comparison with the C- M -DCSK system under multipath fading condition, but possesses relative higher PAPR. Hence, these results will be very useful for a myriad of practical applications. For instance, the C- M -DCSK system is

suitable for wireless-body-area-network (WBAN) [26] [27] applications due to the varying channels and the small size requirements, but the S- M -DCSK system can be applied in wireless-sensor-network (WSN) [27] applications because of the static sensor nodes. To overcome the high-PAPR weakness, our future work will focus on conceiving some strategies to reduce PAPR of this new system.

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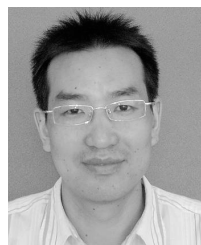
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GUOFA CAI received the Ph.D. degree in communication engineering, Xiamen University, Xiamen, China, in 2015, the M.Sc. degree in circuits and systems from Fuzhou University, Fuzhou, China, in 2012, and the B.Sc. degree in communication engineering from Jimei University, Xiamen, China, in 2007. He is currently a Post-Doctoral Fellow with the School of Information Engineering, Guangdong University of Technology, China. His primary research interests include information theory, chaotic communications, channel coding, UWB, and MIMO communications.



YI FANG (M'15) received the B.Sc. degree in electronic engineering from East China Jiaotong University, China, in 2008, and the Ph.D. degree in communication engineering, Xiamen University, China, in 2013. From 2012 to 2012, he was a Research Assistant in electronic and information engineering with The Hong Kong Polytechnic University, Hong Kong. From 2012 to 2013, he was a Visiting Scholar in electronic and electrical engineering with University College London, U.K. From 2014 to 2015, he was a Research Fellow with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore. He is currently an Assistant Professor with the School of Information Engineering, Guangdong University of Technology, China. His primary research interests are information and coding theory, chaotic modulation, UWB, and MIMO communications.



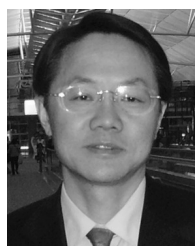
GUOJUN HAN (M'12–SM'14) received the Ph.D. degree from Sun Yat-sen University, Guangzhou, China, and the M.E. degree in electronic engineering from the South China University of Technology, Guangzhou, China. From 2011 to 2013, he was a Research Fellow with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore. From 2013 to 2014, he was a Research Associate with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology. He is currently a Full Professor and the Vice Dean with the School of Information Engineering, Guangdong University of Technology, Guangzhou, China. His research interests include wireless communications, and coding and signal processing for data storage.



FRANCIS C. M. LAU (M'93–SM'03) received the B.Eng. degree (Hons) in electrical and electronic engineering and the Ph.D. degree from King's College London, University of London, U.K., in 1989 and 1993, respectively. He is currently a Professor and an Associate Head with the Department of Electronic and Information Engineering, The Hong Kong Polytechnic University, Hong Kong. He is a fellow of IET. His research interests include chaos-based digital communica-

tions, channel coding, cooperative networks, wireless sensor networks, applications of complex-network theories, and wireless communications. He has co-authored over two research monographs—Chaos-Based Digital Communication Systems and Digital Communications with Chaos: Multiple Access Techniques and Performance Evaluation. He co-holds two U.S. patents related to chaos-based communications and two other U.S. patents related to channel encoders/decoders. He has authored over 270 research papers, including over 120 journal publications. His publications received the "Best Paper Award" at the International Conference on Advanced Technologies for Communications in 2011 and 2015 and the "Outstanding Paper Awards" at the International Conference on Advanced Communication Technology in 2012 and 2013.

He serves as a Technical Committee Program Member, the Session Chair, and a Reviewer of many international conferences. He served as an Associate Editor for several journals, including the IEEE TRANSACTIONS ON CIRCUITS and SYSTEMS II in 2004 and 2005, the IEEE TRANSACTIONS ON CIRCUITS and SYSTEMS I in 2006 and 2007, and the IEEE Circuits and Systems Magazine in 2012 and 2015. He has been a Guest Associate Editor of the International Journal and Bifurcation and Chaos since 2010 and an Associate Editor of the IEEE TRANSACTIONS ON CIRCUITS and SYSTEMS II again since 2016. He was also the Chair of Technical Committee on Nonlinear Circuits and Systems, the IEEE Circuits and Systems Society from 2012 to 2013 and the Co-Track Chair of Nonlinear Circuits and Systems, and the 2010 IEEE International Symposium on Circuits and Systems.



LIN WANG (S'99–M'03–SM'09) received the M.Sc. degree in applied mathematics from the Kunming University of Technology, China, in 1988, and the Ph.D. degree in electronics engineering from the University of Electronic Science and Technology of China, China, in 2001.

From 1984 to 1986, he was a Teaching Assistant with the Mathematics Department, Chongqing Normal University. From 1989 to 2002, he was a Teaching Assistant, a Lecturer, and an Associate Professor in applied mathematics and communication engineering with the Chongqing University of Post and Telecommunication, China. From 1995 to 1996, he spent one year with the Mathematics Department, University of New England, Australia. In 2003, he spent three months as a Visiting Researcher with the Center for Chaos and Complexity Networks, Department of Electronic Engineering, The City University of Hong Kong. From 2013 to 2013, he was a Senior Visiting Researcher with the Department of Electronics and Communication Engineering, UC Davis. From 2003 to 2012, he was a Full Professor and the Associate Dean with the School of Information Science and Technology, Xiamen University, China. He has been a Distinguished Professor since 2012. He holds 12 patents in the field of physical layer in digital communications. He has authored over 100 journal and conference papers. His current research interests are in the area of channel coding, joint source-and-channel coding, chaos modulation, and their applications to wireless communication and storage systems.

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