Dual-Mode Time Domain Multiplexed Chirp Spread Spectrum

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Abstract—In this work, we introduce dual-mode (DM) time domain multiplexed (TDM) chirp spread spectrum (CSS), which offers improved spectral and energy efficiency for low-power wide-area networks (LPWANs). Previous work on LPWANs has proposed DM-CSS modulation using even and odd frequency shifts (FSs) to enhance performance over classical approaches. However, its spectral efficiency is only half that of the in-phase and quadrature (IQ)-TDM-CSS scheme, which employs IQ components with both up and down chirps, resulting in a four times higher SE of Long Range (LoRa). IQ-TDM-CSS only supports coherent detection and is sensitive to the carrier frequency and phase offsets. It makes it less practical for low-cost battery-powered LPWANs in Internet-of-Things (IoT) applications. In contrast, DM-CSS utilizes either an up-chirp or a down-chirp and consists of two chirped symbols that are multiplexed in the time domain. One of these symbols is chirped using an up-chirp and comprises even and odd FSs. The second chirped symbol is also composed of even and odd FSs, which are chirped using a down-chirp. We demonstrate that DM-TDM-CSS achieves a maximum achievable SE close to IQ-TDM-CSS while supporting both coherent and non-coherent detection. Moreover, unlike IQ-TDM-CSS, DM-TDM-CSS is robust against carrier frequency and phase offsets, making it more suitable for low-cost, battery-powered LPWANs in IoT applications.

Index Terms—LoRa, chirp spread spectrum, IoT.

I. INTRODUCTION

The fundamental concept behind Internet-of-Things (IoT) applications is to enable communication between battery-powered devices/sensors while minimizing power consumption to extend the battery life of the terminals. Low-power wide-area networks (LPWANs) are, therefore, essential for such applications. One of the promising technologies for LPWANs is the Long Range (LoRa) wide-area network (LoRaWAN), which employs LoRa as the physical layer modulation scheme [1], [2].

LoRa is a proprietary variant of chirp spread spectrum (CSS) modulation developed by Semtech Corporation. It enables the trade-off of sensitivity with data rates for fixed channel bandwidths [3], [4]. While Semtech has not published the detailed waveform design of LoRa, researchers such as Vangelista in [5] and M. Chiani and A. Elzanaty in [6] have extensively investigated various properties of LoRa modulation, including its waveform design, spectral properties, and low-cost detection process. The number of bits transmitted per LoRa symbol is determined by a scalable parameter known as the spreading factor, \( \lambda = \log_2(M) \), where \( M \) denotes the cyclic time shifts of the chirp that align with the different frequency shifts (FSs) of the down-chirp signal, which is the complex conjugate of the chirp signal. As a result, LoRa is commonly classified as FS chirp modulation [7].

While LoRa has various advantages and widespread adoption, one of its limitations is the low achievable rates in each of the three bands it uses. Therefore, recent research has proposed several spectral-efficient CSS modulation techniques that could serve as potential alternatives to LoRa. A comprehensive explanation of the waveform design of these CSS alternatives can be found in a recent survey [8]. It is worth noting that these CSS variants can have different properties. For example, some possess constant envelope properties and use a single chirp in their symbol structure, while others use multiple chirps and do not retain constant envelope properties. While possessing a constant envelope is desirable, the schemes with a constant envelope generally have low spectral efficiencies, which could be a limiting factor.

When considering alternatives to LoRa using CSS, it’s important to take into account their maximum feasible spectral efficiency (SE). Many of these alternatives focus on improving SE, energy efficiency (EE), or both [9], [10]. Some of the most promising alternatives include IQ-CSS [11], SSK-ICS-LoRa [12], DM-CSS [13], and TDM-CSS [14], among others. It’s worth noting that the literature describes many other CSS schemes for energy-efficient modulations. IQ-CSS encodes bits using the in-phase and quadrature components of the chirp signal. SSK-ICS-LoRa expands the symbol set by using up chirps, down chirps, and their interleaved versions, enabling transmission of two additional bits per symbol compared to LoRa. DM-CSS multiplexes even and odd chirp symbols with different phases and uses either up-chirp or down-chirp signals. TDM-CSS multiplexes two chirps with different slopes in the time domain, while IQ-TDM-CSS uses both the IQ components.
of un-chirped symbols. It’s worth mentioning that the SE of DM-CSS and TDM schemes is higher than that of SSK-ICS-LoRa and classical LoRa. In terms of bit transmission per symbol duration, if LoRa transmits $\lambda$ bits, then SSK-ICS-LoRa, IQ-CSS, TDM-CSS, DM-CSS, and IQ-TDM respectively transmit $\lambda + 2, 2\lambda, 2\lambda, 2\lambda + 1$, and $4\lambda$ bits. For more details on these and other energy-efficient CSS modulations, interested readers can refer to [8].

Despite their advantages, all of these CSS schemes also have some notable drawbacks. According to [8], IQ-CSS is highly sensitive to carrier frequency offset (FO) and has a maximum achievable SE lower than that of DM-CSS and IQ-TDM-CSS. While SSK-ICS-LoRa offers improved EE and supports both coherent and non-coherent detection, it doesn’t significantly improve the SE compared to LoRa and its counterparts. Although TDM-CSS symbols can be detected coherently and incoherently, their maximum possible SE is lower than that of DM-CSS and IQ-TDM-CSS. Additionally, IQ-TDM-CSS is highly sensitive to carrier FO, and only highly complex coherent detection is possible. DM-CSS has better EE than LoRa and supports both coherent and non-coherent detection, but the use of phase shifts (PSs) makes non-coherent detection impractical unless the channel phase rotation is less than $\pi/2$. Furthermore, its highest possible SE is also smaller than that of IQ-TDM-CSS.

Our work introduces the DM-TDM-CSS scheme, which offers a maximum achievable SE comparable to IQ-TDM-CSS. It addresses IQ-TDM-CSS’s limitations by providing coherent and non-coherent detection options while being more resilient to carrier FO and phase offset (PO). DM-TDM-CSS is a fusion of modified versions of DM-CSS and TDM-CSS. Specifically, we utilize the even and odd FSs without PSs, making coherent detection more practical. Additionally, we multiplex two chirped symbols in the time domain, with each un-chirped symbol having unique even and odd FSs. One is chirped with an up-chirp, while the other is chirped with a down-chirp. Unlike DM-CSS, which only uses one type of chirped symbol, DM-TDM-CSS uses both simultaneously. This symbol structure allows for a maximum SE that is only 4 bits lower than IQ-TDM-CSS.

The contributions of this work are as follows:

1) We introduce the DM-TDM-CSS scheme as a viable alternative to existing CSS schemes, such as LoRa. DM-TDM-CSS combines the beneficial features of both DM-CSS and TDM-CSS while mitigating their limitations. This results in a scheme that is energy- and spectral-efficient, as well as robust against carrier frequency and phase offsets.

2) We explain the transceiver design for DM-TDM-CSS, including the waveform generation and the mechanisms for coherent and non-coherent detection.

3) Using mathematical analysis, we determine whether the DM-TDM-CSS symbols are orthogonal to each other. We demonstrate that interference occurs between the even and odd FSs of the up-chirp and down-chirp symbols, resulting in non-orthogonal symbols.

4) Through mathematical analysis, we also estimate the interference caused by the two TDM chirped symbols at the receiver. Our results affirm the conclusion from the orthogonality analysis that the two TDM symbols cause interference with each other.

5) We evaluate the performance of DM-TDM-CSS using various metrics, such as SE versus required signal-to-noise ratio (SNR) per bit for a target bit error rate (BER). BER performance in an additive white Gaussian noise (AWGN) and fading channel, and BER performance considering phase and frequency offsets.

6) We provide closed-form expressions for the interference terms on both the up/down chirped symbols and compute the signal-to-interference ratio (SIR) expressions. We demonstrate that the interference decreases as the value of $\lambda$ increases.

The article is structured as follows: Section II outlines the system model, while Section III introduces the proposed DM-TDM-CSS schemes and analyzes their orthogonality and interference characteristics. In Section IV, we evaluate and compare the performance of DM-TDM-CSS with other schemes. Finally, Section V presents our conclusions based on the results obtained.

II. SYSTEM MODEL

Without loss of generality, we consider a chirped symbol in CSS modulation composed of two components: (i) an un-chirped symbol and (ii) a spreading symbol that spreads the information in the bandwidth, $B$. The un-chirped symbol is a pure sinusoid when only one FS $k$ is activated, or it can be a combination of multiple sinusoids in case multiple FSs are used. When the un-chirped symbol is spread, the FSs have an injective mapping to cyclic time-shift(s). Moreover, the spreading symbol can have different slope rates [15], [16].

We consider that the occupied bandwidth is $B = M/r$, that corresponds to the availability of $M$ FSs implying that $k \in [0, M - 1]$. In the discrete time, we denote the CSS (chirped) symbol consisting of $M$ samples by $s(n) = g(n)c_\xi(n)$, for $n = [0, M - 1]$, where $g(n)$ is the un-chirped symbol, and $c_\xi(n)$ is the spreading symbol given as $c_\xi(n) = \exp(\{j\pi \xi n^2\})$, where $j^2 = -1$ and $\xi$ is the slope rate. Based on the type of CSS modulation, $g(n)$ can have different symbol structures. Moreover, when $\xi = 1$, the spreading symbol corresponds to up-chirp, i.e., $c_\xi(n) = \exp(j\frac{n^2}{2})$. Conversely, if $\xi = -1$, the spreading symbol is a down-chirp denoted as $c_d(n) = \exp(-j\frac{n^2}{2})$. Typically, an up-chirp symbol is used to spread the information in most CSS modulations.

The discrete-time baseband received symbol is given by:

$$y(n) = h s(n) + w(n),$$

where $h$ is the complex channel gain, and $w(n)$ corresponds to AWGN samples. We consider AWGN having single-sided noise power spectral density of $N_0$ and noise variance of $\sigma_n^2 = N_0 B$. It may be noticed that in LPWANs, CSS symbols maintain a narrow bandwidth of 500 kHz or smaller; therefore, a flat fading channel can have a constant attenuation over the entire $B$. Thus, in simplest of cases, it can be considered equal to unity if channel state information (CSI) is known.

III. DUAL-MODE TIME DOMAIN MULTIPLEXED CHIRP SPREAD SPECTRUM

This section will provide a comprehensive analysis of the waveform design for DM-TDM-CSS. Specifically, we will
examine the transceiver architecture of DM-TDM-CSS and demonstrate that its symbols are not orthogonal due to the interference caused by the activation of multiple FSs. Furthermore, we will also conduct an in-depth analysis of the interference characteristics and investigate their impact.

A. Transmission

The transmitter architecture of DM-TDM-CSS is provided in Fig. 1. In DM-TDM-CSS, two chirped symbols are multiplexed in the time domain; therefore, two different un-chirped symbols with \( M \) available frequencies are needed. For each un-chirped symbol, one even and one odd frequency is activated. It may be noticed that among these \( M \) frequencies, \( M/2 \) frequencies are even and \( M/2 \) frequencies are odd. The even activated frequencies for the two un-chirped symbols are \( k_{e,1} \) and \( k_{e,2} \), whereas the odd activated frequencies are \( k_{o,1} \) and \( k_{o,2} \). Note that the even and odd frequencies are determined by the indexes \( k_{e} = 2k_{o} + 1 \), where \( k_{o} \in \{0, M/2 − 1\} \) and \( k_{o} \in \{0, M/2 − 1\} \). \( k_{e,1} \) and \( k_{e,2} \) are determined after binary-to-decimal (bi2de) conversion of bit sequences having lengths \( \lambda_{1} = \lambda - 1 \) and \( \lambda_{2} = \lambda - 1 \), respectively. On the other hand, \( k_{o,2} \) and \( k_{o,2} \) after bi2de conversion of bit sequences having lengths \( \lambda_{3} = \lambda - 1 \) and \( \lambda_{4} = \lambda - 1 \), respectively.

The first un-chirped symbol, \( g_{1}(n) \), comprises two sinusoids. The first sinusoid, \( f_{e,1}(n) \), has an even activated frequency, \( k_{e,1} \), whereas the second sinusoid, \( f_{o,1}(n) \), has an odd activated frequency, \( k_{o,1} \). Then, \( g_{1}(n) \) is given as:

\[
g_{1}(n) = f_{e,1}(n) + f_{o,1}(n)
\]

\[
= \exp \left\{ \frac{j 2\pi}{M} k_{e,1} n \right\} + \exp \left\{ \frac{j 2\pi}{M} k_{o,1} n \right\}.
\]

(2)

Similarly, the second un-chirped symbol, \( g_{2}(n) \), also consists of even frequency, \( k_{e,2} \), and odd frequency, \( k_{o,2} \), activated sinusoids, \( f_{e,2}(n) \), and \( f_{o,2}(n) \). \( g_{2}(n) \) is given as:

\[
g_{2}(n) = f_{e,2}(n) + f_{o,2}(n)
\]

\[
= \exp \left\{ \frac{j 2\pi}{M} k_{e,2} n \right\} + \exp \left\{ \frac{j 2\pi}{M} k_{o,2} n \right\}.
\]

(3)

The next step is to spread the un-chirped symbols, \( g_{1}(n) \), and \( g_{2}(n) \). \( g_{1}(n) \) is then spread using an up-chirp, \( c_{u}(n) \), whereas \( g_{2}(n) \) is spread using a down-chirp, \( c_{d}(n) \) resulting in \( s_{1}(n) \) and \( s_{2}(n) \), i.e.,

\[
s_{1}(n) = g_{1}(n)c_{u}(n)
\]

\[
= \exp \left\{ \frac{j \pi}{M} (2k_{e,1}n + n^{2}) \right\} + \exp \left\{ \frac{j \pi}{M} (2k_{o,1}n + n^{2}) \right\},
\]

(4)

and

\[
s_{2}(n) = g_{2}(n)c_{d}(n)
\]

\[
= \exp \left\{ \frac{j \pi}{M} (2k_{e,2}n - n^{2}) \right\} + \exp \left\{ \frac{j \pi}{M} (2k_{o,2}n - n^{2}) \right\}.
\]

(5)

Afterwards, these two chirped symbols, \( s_{1}(n) \) and \( s_{2}(n) \), are multiplexed in the time domain resulting in \( s(n) \), which is given as:

\[
s(n) = s_{1}(n) + s_{2}(n)
\]

\[
= \exp \left\{ \frac{j \pi}{M} (2k_{e,1}n + n^{2}) \right\} + \exp \left\{ \frac{j \pi}{M} (2k_{o,1}n + n^{2}) \right\}
\]

\[
+ \exp \left\{ \frac{j \pi}{M} (2k_{e,2}n - n^{2}) \right\} + \exp \left\{ \frac{j \pi}{M} (2k_{o,2}n - n^{2}) \right\}.
\]

(6)

DM-TDM-CSS symbol energy is \( E_{s} = \mathbb{E}(|s(n)|^{2}) = \frac{1}{M} \sum_{n=0}^{M-1} |s(n)|^{2} \).

B. Detection

This section presents coherent and non-coherent detection mechanisms for DM-TDM-CSS received symbols, \( y(n) \). For clarity of exposition, we consider the following vectorial representations, \( y = [y(0), y(1), \ldots, y(M − 1)]^{T} \), and \( s = [s(0), s(1), \ldots, s(M − 1)]^{T} \), where \([\cdot]^{T}\) is the transpose operator.

1) Coherent Detection: Fig. 2 depicts the coherent detector architecture for DM-TDM-CSS. Coherent detection involves the estimation of the FSs of the un-chirped symbols, \( k_{e,1}, k_{e,2}, k_{o,1}, \) and \( k_{o,2} \). Assuming that \( h \) is known at the receiver and the transmit symbols are equiprobable, the coherent detection dictates to maximize the probability of receiving \( y \) when \( s \) was sent given \( h \), i.e., \( \text{prob}(y|s, h) \). The likelihood function, \( \text{prob}(y|s, h) \) is given as:

\[
\text{prob}(y|s, h) = \left( \frac{1}{2 \pi \sigma^{2}_{n}} \right)^{M} \exp \left\{ -\frac{||y - hs||^{2}}{2 \sigma_{n}^{2}} \right\}.
\]

(7)

where \( ||\cdot||^{2} \) evaluates Euclidean norm, \( \Re\{\cdot\} \) determines the real component of a complex-valued argument, and

\[
\rho = \left( \frac{1}{2 \pi \sigma^{2}_{n}} \right)^{M} \exp \left\{ -\frac{||y||^{2} + ||hs||^{2}}{2 \sigma_{n}^{2}} \right\}.
\]

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The coherent detection problem in (7) is simplified as:

\[ \hat{k}_{e,1}, \hat{k}_{e,2}, \hat{k}_{o,1}, \hat{k}_{o,2} = \arg \max_{k_e, k_o} \text{prob} \left( y | s, h \right) \]

\[ = \arg \max_{k_e, k_o} \Re \{ \langle y, h s \rangle \}. \tag{8} \]

Considering that \( \langle \cdot \rangle \) evaluates the complex conjugate, \( \langle y, h s \rangle \) can be simplified as:

\[ \langle y, h s \rangle = \bar{y} = \sum_{n=0}^{M-1} y(n) \bar{s}(n) \]

\[ = \bar{y} \left( \sum_{n=0}^{M-1} r_1(n) \bar{g}_1(n) c_1(n) + \sum_{n=0}^{M-1} r_2(n) \bar{g}_2(n) c_2(n) \right) \]

\[ = \bar{y} \left( \sum_{n=0}^{M-1} r_1(n) \bar{T}_c(n) + \sum_{n=0}^{M-1} r_2(n) \bar{T}_o(n) \right) \]

\[ = \bar{y} \left( \Re \{ \mathcal{R}_1(k_{e,1}) \} + \Re \{ \mathcal{R}_1(k_{o,1}) \} + \Re \{ \mathcal{R}_2(k_{e,2}) \} + \Re \{ \mathcal{R}_2(k_{o,2}) \} \right). \tag{9} \]

where \( r_1(n) = y(n)c_1(n) \), and \( r_2(n) = y(n)c_2(n) \). \( \mathcal{R}_1(k) \) and \( \mathcal{R}_2(k) \) is the DFT of \( r_1(n) \) and \( r_2(n) \), respectively. Moreover, \( \mathcal{R}_1(k_{e,1}) \) and \( \mathcal{R}_1(k_{o,1}) \) is the DFT of \( r_1(n) \) evaluated at even and odd indexes, respectively, whereas \( \mathcal{R}_2(k_{e,2}) \) and \( \mathcal{R}_2(k_{o,2}) \) is the DFT of \( r_2(n) \) evaluated at even and odd indexes, respectively. Taking into account the simplification of \( \langle y, h s \rangle \) in (8), the detection problem in (8) becomes:

\[ \hat{k}_{e,1}, \hat{k}_{e,2}, \hat{k}_{o,1}, \hat{k}_{o,2} = \arg \max_{k_e, k_o} \Re \left\{ \bar{y} \left( \mathcal{R}_1(k_{e,1}) + \mathcal{R}_1(k_{o,1}) + \mathcal{R}_2(k_{e,2}) + \mathcal{R}_2(k_{o,2}) \right) \right\}. \tag{10} \]

The FSs evaluated in (10) can also be dis-jointly estimated as:

\[ \hat{k}_{e,1} = \arg \max_{k_e} \Re \left\{ \bar{y} \mathcal{R}_1(k_{e,1}) \right\}, \]

\[ \hat{k}_{e,2} = \arg \max_{k_e} \Re \left\{ \bar{y} \mathcal{R}_2(k_{e,2}) \right\}, \]

\[ \hat{k}_{o,1} = \arg \max_{k_o} \Re \left\{ \bar{y} \mathcal{R}_1(k_{o,1}) \right\}, \]

\[ \hat{k}_{o,2} = \arg \max_{k_o} \Re \left\{ \bar{y} \mathcal{R}_2(k_{o,2}) \right\}. \tag{11} \]

2) Non-Coherent Detection: When the CSI is unavailable, the non-coherent detection mechanism can be used. It is more practical because its computational complexity is considerably lower than the coherent detection, which is better for low-power consumption and low-cost components in LPWANs. The non-coherent detector for DM-TDM-CSS is presented in Fig. 3. For non-coherent detection of DM-TDM-CSS, the FSs from the received symbols in a dis-joint fashion can be identified as:

\[ \hat{k}_{e,1} = \arg \max_{k_e} \Re \{ \mathcal{R}_1(k_{e,1}) \}, \]

\[ \hat{k}_{e,2} = \arg \max_{k_e} \Re \{ \mathcal{R}_2(k_{e,2}) \}, \]

\[ \hat{k}_{o,1} = \arg \max_{k_o} \Re \{ \mathcal{R}_1(k_{o,1}) \}, \]

\[ \hat{k}_{o,2} = \arg \max_{k_o} \Re \{ \mathcal{R}_2(k_{o,2}) \}. \tag{12} \]

From (13), it is observed that the DFT of \( r_1(n) \) and \( r_2(n) \) is first evaluated, which yields \( \mathcal{R}_1(k) \) and \( \mathcal{R}_2(k) \). Subsequently, the FS,
\[ \hat{k}_{c,1} \text{ and } \hat{k}_{o,1} \text{ are determined using } R_1(k) \text{ by separating the even and odd frequency tones, respectively, whereas } \hat{k}_{c,2} \text{ and } \hat{k}_{o,2} \text{ are evaluated using } R_2(k) \text{ by isolating the even and odd frequency tones.} \]

C. Orthogonality Analysis

To analyze whether the DM-TDM-CSS symbols are orthogonal, we evaluate the inner product of the two distinct DM-TDM-CSS symbols, i.e., \( s \) and \( \tilde{s} = [\tilde{s}(0), \tilde{s}(1), \ldots, \tilde{s}(M-1)]^T \) as \( \langle s, \tilde{s} \rangle = \sum_{n=0}^{M-1} s(n)\tilde{s}(n) \). The activated even and odd FSs in the two chirped symbols of \( s \) are \( k_{e,1}, k_{e,2}, k_{o,1}, \text{ and } k_{o,2} \). On the other hand, in \( \tilde{s} \), the even and odd FSs in the two chirped symbols are \( \hat{k}_{e,1}, \hat{k}_{e,2}, \hat{k}_{o,1}, \text{ and } \hat{k}_{o,2} \). The following conditions must hold to determine if the DM-TDM-CSS symbols are orthogonal or not: (i) the even FS in the up-chirp and the down-chirp symbols, \( k_{e,1}, \text{ and } k_{e,2} \) in \( s \) are different from the respective FS in \( \tilde{s} \), \( \hat{k}_{e,1}, \text{ and } \hat{k}_{e,2} \), i.e., \( k_{e,1} \neq \hat{k}_{e,1}, \text{ and } k_{e,2} \neq \hat{k}_{e,2} \); (ii) the same condition also holds for the odd FS of \( s \) and \( \tilde{s} \), i.e., \( k_{o,1} \neq \hat{k}_{o,1}, \text{ and } k_{o,2} \neq \hat{k}_{o,2} \). After some straightforward manipulation, the inner product \( \langle s, \tilde{s} \rangle \) yields:

\[
\langle s, \tilde{s} \rangle = \sum_{n=0}^{M-1} \exp \left\{ j \frac{\pi}{M} (2k_1n + 2n^2) \right\} =: \tau_1
\]
\[
+ \sum_{n=0}^{M-1} \exp \left\{ j \frac{\pi}{M} (2k_2n + 2n^2) \right\} =: \tau_2
\]
\[
+ \sum_{n=0}^{M-1} \exp \left\{ j \frac{\pi}{M} (2k_3n - 2n^2) \right\} =: \tau_3
\]
\[
+ \sum_{n=0}^{M-1} \exp \left\{ j \frac{\pi}{M} (2k_4n - 2n^2) \right\}, \tag{13}
\]

where\( k_1 = k_{e,1} - \hat{k}_{e,2}, k_2 = k_{o,1} - \hat{k}_{o,2}, k_3 = k_{e,2} - \hat{k}_{e,1}, \text{ and } k_4 = k_{o,2} - \hat{k}_{o,1} \). The closed-form expressions for \( \tau_1, \tau_2, \tau_3, \text{ and } \tau_4 \) can be obtained using closed-form expressions for \textit{generalized quadratic Gauss sum}. Firstly, consider \( \tau_1 \), for which \( a = 2, b = 2k_1, \text{ and } c = M \). To this end, we attain:

\[ \tau_1 = \sqrt{\frac{M}{2}} \exp \left\{ j \frac{\pi}{8M} (2M - (2k_1)^2) \right\} \beta_1 \tag{14} \]

where

\[
\beta_1 = \sum_{n=0}^{1} \exp \left\{ -j \frac{\pi}{2} (2k_1n + Mn^2) \right\} = 1 + \exp \left\{ -j \frac{\pi}{2} (M + 2k_1) \right\} \tag{15}
\]

Since \( M \) and \( |k_1| \) are always even; therefore, \( \beta_1 = 2 \) that leads to:

\[ \tau_1 = \alpha \exp \left\{ -j \frac{\pi}{2M} \left( k_{e,1} - \hat{k}_{e,2} \right)^2 \right\}, \tag{16} \]

where \( \alpha = 2\sqrt{M/2} \exp \{ j\pi/4 \} \).

Following the same steps as in (13) and (14), \( \tau_2, \tau_3, \text{ and } \tau_4 \) are obtained as:

\[ \tau_2 = \alpha \exp \left\{ -j \frac{\pi}{2M} \left( k_{o,1} - \hat{k}_{o,2} \right)^2 \right\}, \tag{17} \]
\[ \tau_3 = \alpha \exp \left\{ j \frac{\pi}{2M} \left( k_{e,2} - \hat{k}_{e,1} \right)^2 \right\}, \tag{18} \]

and

\[ \tau_4 = \alpha \exp \left\{ j \frac{\pi}{2M} \left( k_{o,2} - \hat{k}_{o,1} \right)^2 \right\}, \tag{19} \]

respectively. Finally, the closed-form of (13), i.e., \( \langle s, \tilde{s} \rangle \) is given as:

\[ \langle s, \tilde{s} \rangle = \alpha (\theta_1 + \theta_2) + \pi (\theta_3 + \theta_4) \tag{20} \]

where

\[ \theta_1 = \exp \left\{ -j \frac{\pi}{2M} \left( k_{e,1} - \hat{k}_{e,2} \right)^2 \right\}, \tag{21} \]
\[ \theta_2 = \exp \left\{ -j \frac{\pi}{2M} \left( k_{o,1} - \hat{k}_{o,2} \right)^2 \right\}, \tag{22} \]
\[ \theta_3 = \exp \left\{ j \frac{\pi}{2M} \left( k_{e,2} - \hat{k}_{e,1} \right)^2 \right\}, \tag{23} \]

and

\[ \theta_4 = \exp \left\{ j \frac{\pi}{2M} \left( k_{o,2} - \hat{k}_{o,1} \right)^2 \right\}, \tag{24} \]

respectively.

Actuating both even and odd frequency shifts in the two multiplexed chirped symbols results in the loss of orthogonality between the two DM-TDM-CSS symbols, as expressed in eq. (20). Specifically, the even FS of one chirped symbol causes interference with the even FS of the other chirped symbol, and vice versa for the odd FSs.

D. Interference Analysis

The orthogonality analysis has revealed that activating the even FS of one chirped symbol induces interference for the even FS of the other, and the same is the case for the odd FSs. Therefore, it is crucial to analyze this interference quantitatively. This analysis can be done by examining \( r_1(n) \) and \( r_2(n) \) and considering that the received signal \( y(n) = s(n) + w(n) \). In the following analysis, we will focus on \( r_1(n) \) to determine the interference, which is given by:

\[ r_1(n) = y(n)c_d(n) = s(n)c_d(n) + \tilde{w}(n) \]
\begin{equation}
= (g_1(n)c_a(n) + g_2(n)c_a(n))c_d(n) + \hat{w}(n)
= g_1(n) + g_2(n)c_d^2(n) + \hat{w}(n),
\end{equation}

where \( \hat{w}(n) = w(n)c_d(n) \) and \( c_a(n)c_d(n) = 1 \). \( r_1(n) \) in (25) can be re-written as:

\[
r_1(n) = \exp \left\{ \frac{2\pi}{M} k_{\text{e},1} n \right\} + \exp \left\{ \frac{2\pi}{M} k_{\text{o},1} n \right\}
+ \exp \left\{ -\frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{e},2} n \right\}
+ \exp \left\{ -\frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{o},2} n \right\} + \hat{w}(n).
\]

Taking \( M \)-order DFT of \( r_1(n) \) yields \( R_1(k) \), i.e.,

\[
R_1(k) = \sum_{n=0}^{M-1} \exp \left\{ -\frac{2\pi}{2M} \sum_{\xi=\xi_1}^{\xi_2} \exp \left\{ \frac{2\pi}{M} k_{\text{e},1} n \right\}
+ \exp \left\{ -\frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{e},2} n \right\}
+ \exp \left\{ -\frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{o},2} n \right\} + \hat{w}(n),
\]

where \( \tilde{k}_1 = k_{\text{e},1} - k \), \( \tilde{k}_2 = k_{\text{o},1} - k \), \( \tilde{k}_3 = k_{\text{e},2} - k \), and \( \tilde{k}_4 = k_{\text{o},2} - k \).

Now considering \( k \in \tilde{k}_1 \), (27) is ascertained that \( \kappa_1 = M \) when \( \tilde{k}_1 = k_{\text{e},1} \), \( \kappa_2 = 0 \) because \( k_{\text{o},1} \). Moreover, using the closed-form expressions of generalized quadratic Gauss sum, \( \kappa_3 \) is given as:

\[
\kappa_3 = \frac{\pi}{2} \exp \left\{ \frac{j\pi}{2M} \tilde{k}_3^3 \right\} \left( 1 + \exp \left\{ \frac{j\pi}{2} \left( M + 2\tilde{k}_3 \right) \right\} \right) .
\]

If \( M \) is a power of 2 (which in general it is) and if \( k \in \tilde{k}_1 \), then \( \exp\left\{ \frac{j\pi}{2} \left( M + 2\tilde{k}_3 \right) \right\} = 1 \), which leads to

\[
\kappa_3 = \alpha \exp \left\{ \frac{j\pi}{2M} \tilde{k}_3^3 \right\} .
\]

Solving \( \kappa_4 \) yields:

\[
\kappa_4 = \frac{\pi}{2} \exp \left\{ \frac{j\pi}{2M} \tilde{k}_4^3 \right\} \left( 1 + \exp \left\{ \frac{j\pi}{2} \left( M + 2\tilde{k}_4 \right) \right\} \right) .
\]

Since \( \tilde{k}_4 \) is always odd; therefore, \( \exp\left\{ \frac{j\pi}{2} \left( M + 2\tilde{k}_4 \right) \right\} = -1 \) resulting in \( \kappa_4 = 0 \). It is important to note that \( \kappa_2 = \kappa_4 = 0 \) implies that the odd FS of one chirped symbol does not cause any interference with the even FS of the other. Thus, for \( k \in \tilde{k}_1 \), the output of the DFT when \( k_{\text{e},1} = k_{\text{o},1} \) results in

\[
R_1(k_{\text{e},1}) = \frac{M}{\text{signal}} + \alpha \exp \left\{ \frac{j\pi}{2M} \left( k_{\text{e},2} - k_{\text{o},1} \right)^2 \right\} + \hat{W}(k_{\text{e},1}).
\]

Performing similar analysis as done for (31) considering \( k \in \tilde{k}_o \) and \( \tilde{k}_o = k_{\text{o},1} \) yields

\[
R_1(k_{\text{o},1}) = \frac{M}{\text{signal}} + \pi \exp \left\{ \frac{j\pi}{2M} (k_{\text{o},2} - k_{\text{o},1})^2 \right\} + \hat{W}(k_{\text{o},1}).
\]

Note that while evaluating (32), it is observed that \( \kappa_4 = \kappa_3 = 0 \), implying that even FSs of one chirped symbol does not interfere with the odd FS of the other chirped symbol.

\[
r_2(n) = g_1(n)c_a^2(n) + g_2(n) + \hat{w}(n)
= \exp \left\{ \frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{e},1} n \right\}
+ \exp \left\{ \frac{2\pi}{M} n^2 \right\} \exp \left\{ \frac{2\pi}{M} k_{\text{o},1} n \right\}
+ \exp \left\{ \frac{2\pi}{M} k_{\text{e},2} n \right\} + \exp \left\{ \frac{2\pi}{M} k_{\text{o},2} n \right\} + \hat{w}(n),
\]

where \( \hat{w}(n)(n) = w(n)c_a(n) \). \( M \)-order DFT of \( r_2(n) \) results in:

\[
R_2(k) = \sum_{n=0}^{M-1} \exp \left\{ \frac{j\pi}{2M} \left( 2\tilde{k}_1 n + 2n^2 \right) \right\}
+ \sum_{n=0}^{M-1} \exp \left\{ \frac{j\pi}{2M} 2\tilde{k}_2 n + 2n^2 \right\}
+ \sum_{n=0}^{M-1} \exp \left\{ \frac{j\pi}{2M} 2\tilde{k}_3 n \right\} + \hat{W}(k),
\]

where \( \tilde{k}_1 = k_{\text{e},1} - k \), \( \tilde{k}_2 = k_{\text{o},1} - k \), \( \tilde{k}_3 = k_{\text{e},2} - k \), and \( \tilde{k}_4 = k_{\text{o},2} - k \).

For \( k \in \tilde{k}_1 \) when \( \tilde{k}_e = k_{\text{e},2} \), we attain \( \kappa_5 = \alpha \exp \left\{ -j\frac{\pi}{2M} \left( k_{\text{e},1} - \tilde{k}_e \right)^2 \right\} \), \( \kappa_6 = 0 \), \( \kappa_7 = M \), and \( \kappa_8 = 0 \) that leads to

\[
R_2(k_{\text{e},2}) = \frac{M}{\text{signal}} + \alpha \exp \left\{ -j\frac{\pi}{2M} \left( k_{\text{e},1} - k_{\text{e},2} \right)^2 \right\} + \hat{W}(k_{\text{e},2}).
\]

Similarly, for \( k \in \tilde{k}_e \), we have \( \kappa_5 = 0 \), \( \kappa_6 = \alpha \exp \left\{ -j\frac{\pi}{2M} \left( k_{\text{o},1} - \tilde{k}_e \right)^2 \right\} \), \( \kappa_7 = 0 \), and \( \kappa_8 = M \) resulting in

\[
R_2(k_{\text{o},2}) = \frac{M}{\text{signal}} + \alpha \exp \left\{ -j\frac{\pi}{2M} \left( k_{\text{o},1} - k_{\text{o},2} \right)^2 \right\} + \hat{W}(k_{\text{o},2}).
\]

From (31) and (32), we can observe that the activated FSs of the second chirped symbol, i.e., \( k_{\text{e},2} \) and \( k_{\text{o},2} \), cause interference when the activated FSs of the first chirp symbol, i.e., \( k_{\text{e},1} \) and
are to be determined. We can draw similar conclusions from (35) and (36) that the FSs of the first chirped symbol, \( k_{e,1} \) and \( k_{o,1} \), cause interference when we need to determine the FSs of the second chirped symbol, \( k_{e,2} \) and \( k_{o,2} \).

In addition, we can also explicitly obtain the expression for SIR, \( \gamma \) as we have both the signal power and the interference power. Since, the interference power for \( R_1(k_e) \), \( R_1(k_o) \), \( R_2(k_e) \), and \( R_2(k_o) \) is the same, \( \gamma \) is evaluated as:

\[
\gamma = \frac{M^2}{\alpha_1 \exp \left\{ \frac{j \pi}{M} (k_{e,2} - \tilde{k}_e)^2 \right\}} = \frac{M^2}{2M} = \frac{M}{2} \tag{37}
\]

Notice that for the evaluation of \( \gamma \), we have used the interference power of \( R_1(k_e) \); however, as aforementioned, the interference power for \( R_1(k_o) \), \( R_2(k_e) \), and \( R_2(k_o) \) is also the same. Consequently, we will always obtain the same result for \( \gamma \). Furthermore, from (37), we gather that the interference vanishes away with increasing \( M \), i.e., for higher \( \lambda \).

We can also define signal-to-interference plus noise (SINR), \( \Gamma \) as:

\[
\Gamma = \frac{M^2}{2M + \sigma_n^2} = \frac{M}{2 + \frac{\sigma_n^2}{M}} \tag{38}
\]

Again we observe that \( \Gamma \) also increases with an increase in \( M \), i.e., for higher \( \lambda \).

IV. NUMERICAL RESULTS AND DISCUSSIONS

In this section, we present a comprehensive comparison of the proposed DM-TDM-CSS with other classical counterparts in the state-of-the-art using a range of performance metrics obtained via simulations. We consider all the approaches in the literature that can transmit nearly \( 2\lambda \) or more bits per symbol of duration \( T_o \), including IQ-CSS, TDM-CSS, and IQ-TDM-CSS. We also include LoRa in our analysis to provide a benchmark for comparison. We compare the performance of DM-TDM-CSS with other schemes using various metrics, including SE versus EE performance, BER performance in AWGN and frequency-selective fading channels, and BER performance considering phase and frequency offsets. It is important to note that we consider the non-coherent detection for IQ-CSS as proposed in [17]. To ensure consistent interference power levels and bandwidth, we maintain the same value of \( \lambda \) for all schemes while analyzing the BER performance metrics. This is crucial since any changes to this parameter can result in significant interference power and bandwidth variations.

A. Spectral Efficiency of DM-TDM-CSS

The DM-TDM-CSS scheme transmits four different bit sequences of length \( \lambda - 1 \) per symbol, resulting in a total of \( 4\lambda - 4 \) bits per symbol with a duration of \( T_o \). Although DM-TDM-CSS transmits four bits less than IQ-TDM-CSS per symbol, its advantages are demonstrated in subsequent sections. The data rate SE of DM-TDM-CSS, with the given number of bits per symbol, can be calculated as \( R = \frac{4\lambda - 4}{T_o} \) bits/s/Hz. Considering that \( B = \frac{M}{\tau_s} \), the SE of DM-TDM-CSS can be expressed as:

\[
\eta = \frac{R}{B} = \frac{4\lambda - 4}{M} \tag{39}
\]

The spectral efficiencies of other schemes considered in this article are presented in Table I.

B. Spectral Efficiency Versus Energy Efficiency Performance

This section assesses and compares the trade-off between SE and EE of the proposed DM-TDM-CSS with other state-of-the-art schemes. To evaluate the performance at a specific SE, we calculate the EE by determining the \( E_b/N_0 = \frac{E_b}{N_0} \) required to reach a BER of \( 10^{-3} \). Meanwhile, the SE is adjusted by varying \( \lambda = [6, 12] \). We evaluate this performance metric for all the considered schemes in the AWGN channel, considering both coherent and non-coherent detection mechanisms.

According to the SE versus EE performance illustrated in Fig. 4, we observe that the proposed DM-TDM-CSS outperforms all other schemes when coherent detection is employed. Although IQ-TDM-CSS can achieve the maximum achievable SE, the maximum achievable SE of DM-TDM-CSS is slightly less than that of IQ-TDM-CSS. As \( \lambda \) increases above 9, the SE versus EE performance of IQ-TDM-CSS approaches that of DM-CSS, making it a desirable alternative to other approaches. TDM-CSS performs similarly to DM-TDM-CSS. However, IQ-CSS performs better than the proposed DM-TDM-CSS. At first glance, it may seem that the proposed approach is not a good alternative to existing approaches. However, when we consider non-coherent detection, as shown in Fig. 5, it is evident...
that DM-TDM-CSS is one of the best approaches in terms of SE versus EE performance. The reasons are that (i) DM-CSS only offers non-coherent detection when the PSs are 0 and π and requires maximum likelihood detection otherwise, which increases overall detection complexity, and (ii) IQ-TDM-CSS symbols cannot be detected using non-coherent detection, which is a significant limitation.

C. BER Performance in AWGN Channel

Figs. 6 and 7 illustrate the BER performance of the proposed DM-TDM-CSS scheme and compare it with other alternatives considering coherent and non-coherent detection, respectively. The BER performance is obtained considering an AWGN for \( \lambda = 8 \). For \( \lambda = 8 \), the spectral efficiencies of the considered schemes are LoRa: 0.0312 bits/s/Hz, IQ-CSS: 0.0625 bits/s/Hz, TDM-CSS: 0.0625 bits/s/Hz, IQ-TDM-CSS: 0.125 bits/s/Hz, DM-CSS: 0.0664 bits/s/Hz, and DM-TDM-CSS: 0.1093 bits/s/Hz. With the given spectral efficiencies, DM-TDM-CSS offers approximately 250% increase in SE over LoRa, 75% increase in SE over IQ-CSS and TDM-CSS, and 65% over DM-CSS. However, the SE of DM-TDM-CSS is approximately 14% less relative to IQ-TDM-CSS.

The BER performance coherent detection (cf. Fig. 6) illustrates that the BER of DM-TDM-CSS is marginally higher relative to other counterparts. To attain a BER of \( 10^{-3} \), the value of required \( E_b/N_0 \) for DM-TDM-CSS is approximately 1.2 dB, 1 dB, 0.5 dB, and 0.2 dB higher compared to DM-CSS, LoRa/IQ-CSS, TDM-CSS, and IQ-TDM-CSS, respectively. However, it is highlighted that apart from IQ-TDM-CSS, the proposed DM-TDM-CSS provides higher spectral efficiencies, as mentioned before. Thus, considering the significant gain in SE, the increase in the required \( E_b/N_0 \) is not substantial.

The added value of DM-TDM-CSS becomes evident when we analyze the BER performance considering non-coherent detection (cf. Fig. 7). IQ-TDM-CSS, the counterpart capable of achieving similar spectral efficiencies, does not allow non-coherent detection, which makes it less advantageous. On the other hand, DM-TDM-CSS allows non-coherent detection. We can observe that the DM-TDM-CSS requires approximately 1.8 dB, 1 dB, 0.7 dB higher \( E_b/N_0 \) to attain a BER of \( 10^{-3} \) relative to DM-CSS, LoRa, and IQ-CSS/TDM-CSS, respectively. Again, the tradeoff between the achievable SE and the power consumption is evident. Nonetheless, the gain provided by DM-TDM-CSS over other alternatives is considerably high relative to the increase in power consumption, making the scheme a viable alternative to the other counterparts.

D. BER Performance in Frequency-Selective Fading Channel

In this section, we consider a frequency-selective 2-tap fading channel having an impulse response of \( h(n) = \sqrt{1 - \rho^2} \delta(nT) + \sqrt{\rho^2} \delta(nT - T) \), where \( T \) is the sampling duration and \( 0 \leq \rho \leq 1 \). The results are also depicted in Fig. 2, where \( \rho = 0.2 \). The BER performance of the considered approaches considering coherent and non-coherent detection is illustrated in Figs. 7 and 8.

From the BER performance considering coherent detection (cf. Fig. 8), we can observe that, unlike the BER performance in the AWGN channel, the BER performance of DM-TDM-CSS is reasonably better than that of IQ-TDM-CSS. The reason is that DM-TDM-CSS does not transmit any information in
the in-phase and quadrature components like IQ-TDM-CSS, which makes the latter approach more susceptible to frequency selective fading. On the other hand, the performance of DM-TDM-CSS is almost similar to that of IQ-CSS. It is essential to highlight that for a similar performance, the SE of DM-TDM-CSS is 65% higher than IQ-CSS.

The BER performance in fading channel considering non-coherent detection, as depicted in Fig. 9, illustrates that the performance of IQ-CSS and IQ-TDM-CSS is severely affected due to the PO because these schemes incorporate additional information in the IQ components. On the other hand, the performance of DM-TDM-CSS remains essentially robust against the PO. It can also be seen that the BER performance of IQ-CSS and IQ-TDM-CSS was better than the proposed DM-TDM-CSS in the AWGN channel (cf. Fig. 6); however, in the presence of distortions, the performance degrades severely. It is also accentuated that apart from DM-TDM-CSS, no other scheme can transmit $4\lambda - 4$ bits per symbol and is also robust against the PO.

Fig. 11 portrays the BER performance of schemes that employ non-coherent detection and a PO of $\psi = \pi/4$ radians. It may be noticed that the non-coherently detected DM-CSS that was performing the best in the AWGN and fading channels is severely influenced by the PO. Moreover, the performance of non-coherently detected TDM-CSS and IQ-CSS remains better than DM-TDM-CSS; however, these schemes possess half of the SE of DM-TDM-CSS. Consequently, DM-TDM-CSS is the only approach that transmits $4\lambda - 4$ bits per symbol while also employing non-coherent detection.

E. BER Performance Considering Phase Offset

In this section, we analyze the performance of all the schemes considering PO, which is expected to exist in low-cost devices.

To this end, the received symbol corrupted by PO and AWGN is given as:

$$y(n) = \exp\{j\psi\}s(n) + w(n),$$

(40)

where $\psi$ is the PO. We evaluate the performance of the considered schemes for coherent and non-coherent detection and consider a PO of $\psi = \pi/4$ radians and $\lambda = 8$.

Fig. 10 depicts the BER performance of the considered schemes using coherent detection and a PO of $\psi = \pi/4$ radians. The performance of IQ-CSS and IQ-TDM-CSS is severely affected due to the PO because these schemes incorporate additional information in the IQ components. On the other hand, the performance of DM-TDM-CSS remains essentially robust against the PO. It can also be seen that the BER performance of IQ-CSS and IQ-TDM-CSS was better than the proposed DM-TDM-CSS in the AWGN channel (cf. Fig. 6); however, in the presence of distortions, the performance degrades severely. It is also accentuated that apart from DM-TDM-CSS, no other scheme can transmit $4\lambda - 4$ bits per symbol and is also robust against the PO.

F. BER Performance Considering Frequency Offset

In this section, we investigate the BER performance in the presence of FO. The carrier FO linearly accumulates phase rotations from one symbol to another. In this case, the received
symbol incorporating the impact of FO is

$$y(n) = \exp \left\{ \frac{2\pi \Delta f n}{M} \right\} s(n) + w(n),$$  \hspace{1cm} (41)

where $\Delta f$ is the FO in Hz.\(^1\) To evaluate the BER performance, we consider FO of $\Delta f = 0.2$ Hz, $\lambda = 8$, and AWGN channel for the considered schemes.

Fig. 12 shows the BER performance of coherently detected schemes considering FO of $\Delta f = 0.2$ Hz in the AWGN channel for $\lambda = 8$. Again, we can observe that the performance of the schemes which transmit information in the IQ components is considerably affected by the FO. It can also be observed that DM-TDM-CSS requires 1 dB higher $E_b/N_0$ to attain a BER of $10^{-3}$ relative to DM-CSS, and 0.9 dB higher $E_b/N_0$ compared to TDM-CSS and IQ-CSS. It should be noted that the results presented in the figure are based on the assumption of $\lambda = 8$, where DM-TDM-CSS demonstrates significantly higher spectral efficiencies compared to most other schemes. As a result, we can infer that despite the higher SE, the corresponding increase in BER is relatively small. Furthermore, we can also observe that in addition to being capable of achieving lower SE, the performance of DM-CSS and IQ-CSS is severely affected by the PO.

G. Advantages and Disadvantages of DM-TDM-CSS

From the results in the previous section, DM-TDM-CSS has some apparent advantages. Firstly, we observe that it can achieve a higher SE than other alternatives, apart from IQ-TDM-CSS, which transmits only four additional bits per symbol. Secondly, enhanced resilience against the PO and FO compared to other counterparts. It is noticeable that IQ-CSS, IQ-TDM-CSS, and non-coherently detected DM-CSS are very much affected by the PO. The performance of IQ-CSS and IQ-TDM-CSS also degrades due to FO. Thirdly, it manifests a robust performance in the frequency selective channel compared to other counterparts, such as IQ-CSS and IQ-TDM-CSS. Lastly, there is no other CSS approach capable of both coherent and non-coherent detection, and yields a SE of $(4\lambda-4)/M$ at the same time.

One major limitation of DM-TDM-CSS is its lack of a constant envelope, which could pose practical challenges due to the waveform’s high peak-to-average power ratio (PAPR). PAPR, $\rho$, is mathematically defined as:

$$\rho = \frac{\max \{|s(n)|^2\}}{\frac{1}{M} \sum_{n=0}^{M-1} |s(n)|^2}. \hspace{1cm} (42)$$

\(^1\) $\Delta f$ could also be seen as residual FO because IoT modems (for example, Bluetooth) normally implement a carrier frequency offset compensator before demodulation, thus, after compensation, $\Delta f$ will reflect the residual FO.
To graphically illustrate PAPR of a signal, the complementary cumulative distribution function (CCDF) is used, which measures the probability that $\rho$ exceeds a specified threshold, $\rho_0$, i.e., $Pr(\rho > \rho_0)$.

Fig. 14 depicts the CCDFs for PAPR of various schemes considered in this study, all evaluated for $\lambda = 8$. Notably, the two schemes with the highest SE, IQ-TDM-CSS, and DM-TDM-CSS, also exhibit the highest PAPR values. However, our previous findings indicate that DM-TDM-CSS outperforms other schemes in terms of robustness against various offsets. Additionally, we can observe that most CSS schemes, which outperform the classical LoRa in terms of SE, have high PAPR due to using PS or additional FS to achieve higher SE.

V. CONCLUSION

The proposed DM-TDM-CSS scheme presents several advantages over the existing alternatives in the literature, as demonstrated in this study. DM-TDM-CSS offers both coherent and non-coherent detection, as well as a higher maximum achievable spectral efficiency. Additionally, it is robust against both the phase and carrier frequency offset. The BER performance of the proposed scheme outperforms other schemes that offer similar spectral efficiencies in frequency-selective fading channels. Our mathematical analysis shows that DM-TDM-CSS symbols are not orthogonal, and interference caused by chirp symbols is also evaluated. We demonstrate that the simultaneous activation of two chirp symbols with different chirp rates results in interference. However, the high PAPR of the proposed approach is a limiting factor. Despite this, the advantages of DM-TDM-CSS highlighted in this study could encourage further research into the scheme, and potential solutions could be explored to overcome its limitations.

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