A Survey on Chirp Spread Spectrum-based Waveform Design for IoT

Ali Waqar Azim, Ahmad Bazzi, Raed Shubair, Marwa Chafii

Abstract—Long Range (LoRa) is one of the most promising and widespread chirp spread spectrum (CSS)-based physical (PHY) layer technique for low-power wide-area networks (LPWANs). Using different spreading factors, LoRa can attain different spectral/energy efficiencies, and can target multitude of Internet-of-Thing (IoT) applications. However, one of the limiting factors for LoRa is the low bit rate. Little to no effort has been made in order to improve the achievable rate of LoRa, until recently, when a number of CSS-based PHY layer LoRa alternative are proposed for LPWANs. In this survey, for the first time, we present a comprehensive waveform design of these CSS-based schemes that have been proposed between 2019 to 2022. In total, fifteen alternatives to LoRa are compared. Other survey articles related to LoRa mostly tackle different issues, such as LoRa networking, LoRa deployment in massive IoT networks, and LoRa architectures, etc. This survey, on the other hand, comprehensively elucidates the waveform design of LoRa alternatives. The CSS schemes studied in this survey are divided into single chirp, multiple chirp, and index modulation based on the multiplexing pattern of the chirps. Complete transceiver architecture of these CSS schemes is studied, and performance is evaluated in terms of energy efficiency (EE), spectral efficiency (SE), bit-error rate (BER) performance in additive white Gaussian noise, BER in the presence of phase and frequency offsets. It has been observed that the EE, SE and robustness against the offsets is primarily linked to transmit symbol frame structure. The public versions of the MATLAB codes for the CSS schemes studied in this survey shall be provided to promote reproducible research.

Index Terms—Waveform design, LoRa, chirp spread spectrum, IoT.

I. INTRODUCTION

Low-Power Wide-Area Networks (LPWANs) is a powerful emerging Internet-of-Thing (IoT) technology for seamless connectivity of millions of smart devices to access core network via the internet. LPWANs allow ubiquitous connectivity between millions of sensors to collect data and cooperate with each other, thus allowing a plethora of smart IoT application development leading to a smart way of living, where everything is connected and controlled via the internet [1]. Therefore, it is expected that IoT devices shall be omnipresent in the near future.

One of the most important aspect of IoT connectivity is the development of waveform design. Since pervasive connectivity of millions of devices is expected, low-power consumption resulting from the efficient waveform design would be of immense importance. This is because the nodes in the IoT network will be battery powered terminals, where regular maintenance of each and every node would be a strenuous and unmanageable task. A general target for the battery life based on the frequency of transmission is somewhere between 5 to 10 years. Various LPWAN standards have been proposed, using either licensed and/or unlicensed frequency bands, such as industrial, scientific and medical (ISM) band. The 3GPP standard that uses the licensed band is Narrowband-IoT (NB-IoT) [2]. The other, license free standard, is the LoRaWAN. Other notable mentions are Sigfox, and Ingenu [3]. This survey shall restrict the attention to LoRaWAN because of its widespread deployment relative to other counterparts [4].

LoRaWAN uses Long Range (LoRa) as MODEM (MODulation DEModulation). LoRa is touted as one of the promising physical (PHY) layer techniques to attain the benefits of low-power consumption and widespread connectivity over long distances [5], [6], [7], [8], [9]. Since LoRa is a proprietary techniques developed by Semtech, the details of its design are largely unknown. However, it is well known that it is a derivative of chirp spread spectrum (CSS) technique. In the recent past, some attempts have been made to theoretically elucidate the properties of LoRa modulation, such as [7], [10], [11]. In [7], Vangelista has described LoRa as frequency-shift chirp modulation, in which there is a frequency-shift keying (FSK) component that is the un-chirped symbol, and a spreading part, that spreads the bandwidth of the un-chirp symbol in the entire bandwidth. The frequency spreading makes LoRa resilient to Doppler shift. Moreover, it is also inherently robust against doubly dispersive channels, and offers high sensitivity in detecting potentially weak signals consuming extremely low power [11], [12]. Another beneficial property of LoRa is that it possesses a constant envelope that allows the use of low-cost devices, such as power-efficient non-linear amplifiers [13], [14], [15].

LoRa operates in the ISM band, and uses three different bandwidths, that are 125 kHz, 250 kHz, and 500 kHz. The bit rate is characterized by the parameter referred to as the spreading factor (SF), that ranges between 6 and 12. Using different SFS, and code rates, a wide range of throughputs can be attained using LoRa, that may be used to target a multitude of applications ranging from smart cities to agriculture, etc. [13]. Nonetheless, the achievable bit rate/spectral efficiency (SE) is low, which may be a limiting factor for LoRa. LoRa MODEMs can also have FSK transceiver to double the throughput but adding these separate transceiver can be costly and may impact the power consumption due to the requirement of additional computational resources.

In order to overcome this limitation of LoRa, several PHY layer variants of CSS are proposed in the literature as direct competitors to LoRa. The main objectives of these schemes is to either improve the bit-rate/SE and/or to improve the energy

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efficiency (EE). These improvements are essentially linked to the waveform design of these schemes. It may be noticed that the waveform design and properties of CSS alternatives to LoRa vary significantly, e.g., some schemes introduce a phase-shift (PS) to the chirp \[16\], while other designs use different chirp rates \[17, 18\], and index modulation (IM) paradigm. \[19, 20\], etc. A limited number of schemes have been proposed in literature in recent past, however, to the best of our knowledge, a survey article that consolidate the waveform design of these CSS PHY layer alternatives to LoRa is not available in the literature. Our survey comprehensively provides the waveform design of a number of CSS alternatives to LoRa that have been proposed in the literature from 2019 to present. Table I provides a list of acronyms that are used in this survey.

A. LoRa Surveys

A considerable number of survey articles with keyword 'LoRa' have been published in the recent years. A non-exhaustive list of these surveys and their succinct description is provided in Table II in a chronological order. From Table II, it can be observed that primarily the survey articles published on LoRa are either based on the performance analysis, network analysis/deployment, and architecture etc. It is worth noting that there is currently no survey in the literature that evaluates the performance of CSS alternatives to LoRa. Only \[21\] presents different CSS-based PHY layer LoRa alternatives in a very brief manner, without any emphasis on the waveform design.

B. Motivation and Contributions

Unlike the surveys listed in Table II, which focus on providing a general overview of LoRa technology where the networking, deployment, architecture and applications are the main subject, our work focuses on surveying the CSS-based waveform design that has been proposed in the literature as a possible alternative to LoRa. The contributions of this survey are summarized as follows:

- First, we provide coherent and non-coherent detection principles from a waveform design perspective along with the system model. To the best of our knowledge, these principles are never explicitly provided in the literature.
- Next, we classify the PHY layer CSS waveform designs into different categories based on number of chirps that are multiplexed. This categorization results in three different groups, that are, single chirp, multiple chirp and multiple chirp with IM. The CSS-based schemes that are discussed in the survey are listed in Table III. In total, fifteen CSS-based PHY layer alternatives to LoRa published between 2019 and 2022 are investigated.
- Then, we provide the waveform design of each scheme in a comprehensive manner, where time-domain symbol structure, coherent detection and non-coherent detection mechanisms (where applicable) are also provided. Moreover, for the first time, we also present the coherent detection mechanism of different schemes, such as ICS-LoRa, DO-CSS, DM-CSS, and TDM-CSS, which are not available in the existing literature. The block diagram of the transceivers of these schemes are also provided for clarity.
- Afterwards, we investigate the performance of the CSS schemes in terms of EE, SE, bit-error rate (BER) per-
TABLE II: LoRa related surveys available in the literature.

| Ref. | Description                                                                 | Year |
|------|------------------------------------------------------------------------------|------|
| [22] | Provides an overview of LPWANs, and discusses its advantages for applications related to smart cities. | 2016 |
| [14] | Provides an overview of LoRa, and evaluates the PHY and data link layer performance by field tests and simulations. | 2016 |
| [23] | Compare the strength and limitations of different LoRa/LoRaWAN test beds.     | 2017 |
| [24] | Compares NB-IoT and LoRa considering battery lifetime, capacity, cost, quality-of-service (QoS), latency, reliability, and range. | 2017 |
| [9]  | Surveys the design and techniques employed in LPWANs offering wide-area coverage. | 2017 |
| [25] | Evaluates the performance of LoRa considering IoT requirements                | 2017 |
| [26] | Surveys different LPWAN solutions in context of machine-to-machine communications | 2017 |
| [27] | Compares research and industrial states of NB-IoT and LoRa in terms of energy efficiency, capacity, QoS, reliability and coverage range. | 2017 |
| [28] | Consolidates LoRa and LoRaWAN research from 2015 to September 2018 considering PHY layer aspects, network layer aspects, possible improvements, and extensions to the standard. | 2018 |
| [29] | Identify LoRa applications, research trends and practical applications along with recommendations on how to leverage LoRa to full extent in IoT deployment. | 2018 |
| [13] | Surveys technical challenges in LoRa networks, and their possible solutions.  | 2019 |
| [30] | Explores LoRa application fields and integration of Edge computation capability in IoT. | 2019 |
| [31] | Investigates NB-IoT and LoRa, and provides an overview of challenges preventing LPWANs from moving from theory to wide-spread practice. | 2019 |
| [32] | Provides a review of the multihop proposals for LoRaWAN.                      | 2019 |
| [33] | Surveys adaptive data rate (ADR) algorithms for LoRaWAN.                     | 2019 |
| [34] | Reviews LoRaWAN architecture, applications, and security concerns, and list possible countermeasures. | 2019 |
| [35] | Reviews LoRa and LoRaWAN, and discusses applying Ultra-Dense Network concept on LPWAN. | 2019 |
| [36] | Reviews the methods for image transfer via LoRa infrastructure.             | 2020 |
| [37] | Reviews use cases requiring confirmed traffic, and investigates LoRaWAN confirmed traffic. | 2020 |
| [38] | Surveys the available tools for simulating LoRa networks in NS-3 network simulator. | 2020 |
| [39] | Presents an overview on LoRa networks considering PHY layer characteristics, deployment and hardware features, end device transmission settings, and MAC protocols. | 2021 |
| [40] | Provides experimental performance evaluation of ADR enhancements in a mobile node scenario. | 2021 |
| [41] | Studies the real-time deployments of unmanned aerial vehicle (UAV)-based LoRa network. | 2021 |
| [42] | Compares simulation tools for investigating and analyzing LoRa/LoRaWAN network performance. | 2021 |
| [43] | Provides a tutorial on the LoRa standard, and surveys existing solutions, hot topics and future insights for building energy efficient IoT infrastructures and IoT devices. | 2022 |
| [44] | Categorizes and compares LoRa networking techniques.                       | 2022 |
| [45] | Surveys LoRa from a systematic perspective: LoRa analysis, communication, security, and its enabled applications. | 2022 |
| [46] | Surveys the scalability challenges/solutions to assist. LoRaWAN deployment in massive IoT networks | 2022 |

In order to promote reproducible research, we will share the public versions of the MATLAB codes for all the schemes presented in this survey.

C. Paper Organization

The rest of the paper is organized as follows. We start Section II by providing the system model, used in the paper. Moreover, Section II also provides the coherent and non-coherent CSS detection principles, and some definitions of basic parameters that shall be used throughout the article. The transceiver design of the CSS schemes is provided in Section III. Section IV compares the performances of the CSS schemes in terms of EE, SE, BER performance in AWGN channel, and BER performance considering phase and frequency offsets. In Section V we elucidate the merits and limitations of each of the investigated CSS schemes, and also list a few better alternatives to LoRa. Finally, in Section VI we render
conclusions.

II. PRELIMINARIES

A. Basic Definitions

In this survey, we frequently use the terms, bandwidth, symbol duration, sample rate, bit rate, SE, EE, therefore, in this section, we have provided brief definitions of these parameters for the clarity of the readers.

1) Bandwidth: Bandwidth, $B$ is the frequency in Hertz (Hz) occupied by a transmit CSS symbol. The bandwidth, $B$, is divided into $M$ frequencies, where the separation between two adjacent frequencies is $\Delta f$ Hz, therefore, $B = M \Delta f$ Hz.

2) Symbol Duration: Symbol duration, $T_s$ is the time in which one CSS symbol, occupying bandwidth, $B$, can be transmitted. $T_s$ is linked to $\Delta f$ as $\Delta f = 1/T_s$. We may interchangeably use the term symbol period for the same parameter in the article.

3) Sample Duration: Sample duration, $T_e$ is the rate at which an analog symbol is sampled. In this work, we consider that $T_e = T_s/M$.

4) Bit Rate: Bit rate, $R$ in bits/s is the number of bits that can be transmitted in a symbol period, $T_s$. Assuming that $\lambda$ bits can be transmitted per $T_s$, the bit rate is given as $R = \lambda/T_s$.

5) Spectral Efficiency: SE is the achievable bit rate per occupied bandwidth, and is mathematically given as $R/B$. Alternatively, SE can also be defined as the number of transmitted bits per symbol of duration $T_s$. In the sequel, we shall denote SE by $\eta$.

6) Energy Efficiency: EE is the required signal-to-noise ratio SNR for correct bit detection at a given BER.

B. System Model

Without loss of generality and following the system model proposed by Vangelita [7], we consider a generalized CSS modulation, whereby, the chirped symbol is obtained by the multiplication of two components: (i) an un-chirped symbol, and (ii) a spreading/chirp symbol that spreads the information in the bandwidth, $B$. The un-chirped symbol can be a pure sinusoid when only one frequency-shift (FS) $k$ is activated, or it can be a combination of multiple sinusoids, in case multiple FSs are used. In this generalized system model, we consider that only one FS is used, resulting in a FSK symbol. It may be noticed that when the FSK symbol is chirped, the FS has an injective mapping to a unique cyclic time-shift of the chirped signal. A total number of $M$ FSs available, and therefore, the FS $k$ is $k \in [0, M-1]$, and the occupied bandwidth is $B = M \Delta f$. The number of FSs in the CSS modulation are defined as $M = 2^\lambda$, where $\lambda$ is the SF. It is recalled that in LoRa, SF is equal to the number of bits transmitted per symbol. In the discrete-time, a CSS symbol like LoRa consisting of $M$ samples is denoted by $s(n) = f(n)c_u(n)$, for $n = [0, M-1]$, where $f(k) = \exp \left\{ j \frac{2\pi}{M} kn \right\}$ is the un-chirped symbol, and $c_u(n)$ is the up-chirp symbol given as $c_u(n) = \exp \left\{ j \frac{2\pi}{M^2} n^2 \right\}$ with $j^2 = -1$. Here, it is considered that only up-chirp, $c_u(n)$ is used for spreading the un-chirped symbol, however, it is highlighted that different CSS schemes may employ spreading symbols, other than the up-chirp. The sample rate $T_e$ that corresponds to a sampling frequency of $F_c = M/T_e = B$. This implies that each discrete-time chirp symbol is be completely represented by $M$ samples taken at rate $F_c$.

For the sake of simplicity of the system model, hereby, it is considered that the information bearing element in the CSS symbol, $s(n)$ is only the FS, $k$, however, it shall become evident in the sequel when different CSS schemes are presented, that in different CSS modulations, the information can be transmitted via different components of the transmit symbol structure, such as the PS, etc. It is also considered that all the possible symbols of CSS modulation are equiprobable. For example, in case of LoRa, the discrete orthonormal basis are equal to $M$, therefore, the probability of transmission of each symbol is $1/M$.

In LPWANs, CSS symbols have a narrow bandwidth of 500 kHz or less, therefore, it is safe to assume a flat fading channel having a constant attenuation over the entire bandwidth, $B$. The discrete-time baseband received symbol is:

$$y(n) = s(n) + w(n), \quad (1)$$

for $n = [0, M-1]$, where $w(n)$ are the AWGN samples having single-sided noise power spectral density of $N_0$, and noise variance of $\sigma^2_w = N_0 B$. It may be noticed that here we consider that the channel attenuation is unity. Moreover, channel is not shown in eq. (1) due to full channel state information (CSI) knowledge.

C. Coherent Detection Principles

Consider that the information bearing element of the CSS symbols is the FS, $k$, therefore, in the simplest form, coherent detection involves the estimation of $k$. However, in other CSS variants, the estimation problem may be extended to also determine other information bearing parameters. Nonetheless, the coherent detection process is capable of estimating all these parameters jointly by maximizing the likelihood function considering all the possible orthonormal basis.

The coherent detection process dictates to maximize the probability of receiving $y = [y(0), y(1), \ldots, y(M-1)]^T$ given $s = [s(0), s(1), \ldots, s(M-1)]^T$, i.e., $\text{prob} \left\{ y | s \right\}$. Considering that all the possible symbols are equiprobable, the likelihood function, $\text{prob} \left\{ y | s \right\}$ follows a Gaussian probability density function as:

$$\text{prob} \left\{ y | s \right\} = \left( \frac{1}{2\pi\sigma^2_w} \right)^M \exp \left\{ -\frac{\| y - s \|^2}{2\sigma^2_w} \right\}, \quad (2)$$

where $\| \cdot \|^2$ evaluates Euclidean norm, $\mathbb{R}\{ \cdot \}$ determines the real component of a complex argument, $\langle \cdot, \cdot \rangle$ evaluates the inner product, and $C_{st} = (1/2\pi\sigma^2_w)^M \exp \left\{ -\| y \|^2 / 2\sigma^2_w \right\}$. Thus, the coherent detection problem is simplified as:

$$\hat{k}_{coh} = \arg \max_k \text{prob} \left\{ y | s \right\} = \arg \max_k \mathbb{R}\{ \langle y, s \rangle \}. \quad (3)$$

Eq. (3) implies that using coherent detection, the inner product of $y$ and all the possible transmit symbol have to
be determined. This evaluation is computationally intense. However, considering that the transmit symbol is \( s(n) = f(n)c_d(n) \), \( \Re \{ \langle y, s \rangle \} \) can be further simplified as:

\[
\Re \{ \langle y, s \rangle \} = \Re \left\{ \sum_{n=0}^{M-1} y(n)s(n) \right\} = \Re \left\{ \sum_{n=0}^{M-1} y(n)f(n)c_d(n) \right\} = \Re \left\{ \sum_{n=0}^{M-1} r(n)f(n) \right\} = \Re \left\{ \sum_{n=0}^{M-1} r(n) \exp \left\{ -j \frac{2\pi}{M} nk \right\} \right\} = \Re \{ R(k) \},
\]

where \( (\cdot) \) is the conjugate operator, \( r(n) = y(n)c_d(n) \), and \( R(k) \) is the DFT of \( r(n) \) evaluated at \( k \)th index. Here, \( c_d(n) = \exp \left\{ -j \frac{2\pi}{\lambda} n^2 \right\} \) is the down-chirp which is the conjugate of \( c_d(n) \). Now, the detection problem \( 4 \) in its simplified form becomes:

\[
\hat{k}_{\text{coh}} = \arg \max_k \Re \{ R(k) \}.
\]

In practice, the received signal is first correlated with \( c_d(n) \) resulting in \( r(n) \). Then, the DFT of \( r(n) \) resulting in \( R(k) \) is evaluated, from which, an estimate on the transmitted FS, i.e., \( \hat{k} \) is obtained by taking the real argument of \( R(k) \). It can be observed that for coherent detection, only the real component of \( R(k) \) is considered, thereby, neglecting all the noise on the imaginary components.

It is also accentuated that eq. \( 5 \) is only valid for simple CSS schemes like LoRa, whereas, for other CSS approaches, the inner product \( \langle y, s \rangle \) may be completely different depending on the signal structure of the transmit symbol. Therefore, in the sequel, when we consider the coherent detection for a CSS scheme, we will derive \( \langle y, s \rangle \) according to the transmit symbol structure.

\section{Non-Coherent Detection Principles}

When the CSI is unavailable, in that case, non-coherent detection mechanism can be used. The non-coherent detection mechanism is more practical in a sense that the computational complexity is considerably less relative to the coherent detection that is ideal for low-power consumption and low-cost components in LPWANs. Using non-coherent detection, the transmit FS is identified as:

\[
\hat{k}_{\text{non-coh}} = \arg \max_k \left| \sum_{n=0}^{M-1} y(n)s(n) \right| = \arg \max_k \left| R(k) \right|,
\]

where \( | \cdot | \) is the absolute operator, and the second equality hold owing to eq. \( 4 \).

For the estimation of the FS using non-coherent detection, first the DFT of the de-chirped received symbol, \( r(n) \) is evaluated, that results in \( R(k) \). Then, the maximum argument of absolute of \( R(k) \) is evaluated which essentially gives the frequency index which has the maximum peak. It can be observed that unlike coherent detection, non-coherent detection also takes into account the impact of noise on both, the real, as well as imaginary components. Thus, it is foreseen that the performance of coherent detection would be better than the non-coherent detection in AWGN. Furthermore, it is highlighted that with the coherent detector, any phase rotation of the channel can be reverted, whereas, with non-coherent detection the random channel phase rotation cannot be reverted.

\section{CSS-based PHY Layer Waveform Design}

In the sequel, for simplicity and clarity, we only consider the PHY layer waveform design of un-coded CSS-based variants. The performance with appropriate coding is interesting and is left as a topic for future study. Table \( \text{III} \) lists the CSS schemes that are studied in this survey. These CSS schemes are classified into three categories: (i) single chirp, (ii) multiple chirp, and (iii) multiple chirp with IM. In single chirp schemes, only one FS is used for an un-chirped symbol which corresponds to the use of a certain cyclic time-shift for the chirped symbol. In multiple chirp schemes, multiple FS are used for the un-chirped symbol that corresponds to the multiplexing of multiple cyclic time-shifts for the chirped symbols. In multiple chirp with IM, multiple FSs are used for the un-chirped symbol, where the FSs are determined based on a certain activation pattern. Moreover, note that in the sequel, single chirp schemes are referred to as LoRa, whereas, multiple chirp schemes are referred to as CSS.

\subsection{LoRa}

LoRa is a proprietary single chirp derivative of CSS modulation, developed by the Semtech corporation, capable of trading off sensitivity with data-rates for fixed channel bandwidths. Even though the details of LoRa modulation were never published by the Semtech corporation, Vangelista in \( 17 \) has provided a comprehensive theoretical explanation of LoRa modulation with an optimal low-complexity detection process based on discrete Fourier transform (DFT). The scalable parameter of the LoRa modulation is the spreading factor, \( \lambda \), where \( \lambda = [6,12] \). \( \lambda \) is in fact equal to the number of bits transmitted by a LoRa symbol of duration \( T_s \). Based on \( \lambda \), a total of \( M \) distinct orthogonal symbols are possible for LoRa, which are defined using \( M \) distinct cyclic time-shifts of the chirp. These cyclic time-shifts corresponds to frequency-shifts (FSs) of the complex conjugate of the chirp signal, i.e., down-chirp signal, thus, LoRa can be regarded as a chirped frequency-shift keying (FSK) modulation. It is highlighted that LoRa is a physical layer modulation which is agnostic to higher-layer implementations allowing it to coexist along with other network architectures.

\subsubsection{Transmission}

The transmitter architecture of LoRa is presented in Fig. \( 1 \) Consider that there are \( M \) frequencies available in the bandwidth, \( B \). The activated FS \( k \) is determined after binary-to-decimal (bi2de) conversion of the \( \lambda \) bits.
TABLE III: Different CSS schemes discussed in this survey.

| Ref. | Name              | Time     | Nature           | Design                                                                 |
|------|-------------------|----------|------------------|------------------------------------------------------------------------|
| [12], [7] | LoRa/FSCM        | 2015/2017| Single Chirp     | Use single chirp symbol                                               |
| [47]  | ICS-LoRa          | 2019     | Single Chirp     | Use single chirp symbol and its interleaved version                   |
| [48]  | E-LoRa            | 2019     | Single Chirp     | Use either in-phase or quadrature chirped symbol                      |
| [16]  | PSK-LoRa          | 2019     | Single Chirp     | Use single chirp symbol with a PS                                     |
| [49]  | DO-CSS            | 2019     | Multiple Chirp   | Multiplex even and odd chirped symbols                                |
| [17]  | SSK-LoRa          | 2020     | Single Chirp     | Use single chirp symbol that is either up-chirped or down-chirped     |
| [50]  | IQ-CSS            | 2020     | Multiple Chirp   | Multiplex in-phase and quadrature chirped symbols                    |
| [18]  | DCRK-CSS          | 2021     | Single Chirp     | Use single chirp symbol that can be chirped at different discrete chirp rates |
| [19]  | FSCSS-IM          | 2021     | Multiple Chirp with IM | Multiplex multiple chirps using IM precept                           |
| [20]  | IQ-CIM            | 2021     | Multiple Chirp with IM | Use in-phase and quadrature chirped symbols with IM                  |
| [51]  | SSK-ICS-LoRa      | 2022     | Single Chirp     | Use up-chirped or down-chirped symbols and their interleaved versions |
| [52]  | ePSK-LoRa         | 2022     | Multiple Chirp   | Use multiple chirp symbols each having a different PS for redundancy  |
| [53]  | GCSS              | 2022     | Multiple Chirp   | Multiplex multiple group-based chirped symbols                       |
| [54]  | TDM-CSS           | 2022     | Multiple Chirp   | Multiplex both up-chirped and down-chirped symbols simultaneously     |
| [55]  | IQ-TDM-CSS        | 2022     | Multiple Chirp   | Multiplex in-phase and quadrature up-chirped and down-chirped symbols simultaneously |
| [56]  | DM-CSS            | 2022     | Multiple Chirp   | Multiplex even and odd up-chirped or down-chirped symbols with a binary PS |

Using $k$, a discrete-time un-chirped symbol, $f(n)$ is obtained as:

$$f(n) = \exp\left\{ 2\pi \frac{k n}{M} \right\},$$

for $n = [0, M - 1]$. $f(n)$ is then chirped using an up-chirp symbol, $c_u(n)$, resulting in $s_{\text{LoRa}}(n)$. The discrete-time chirped symbol, $s_{\text{LoRa}}(n)$ is given as:

$$s_{\text{LoRa}}(n) = f(n)c_u(n) = \exp\left\{ j \frac{\pi}{M} n(2k + n) \right\}.$$  \hfill (8)

Note that we are ignoring the normalization factors for notational simplicity. The symbol energy of a LoRa symbol is $E_s = \frac{1}{M} \sum_{n=0}^{M-1} |s_{\text{LoRa}}(n)|^2 = 1$.

2) Detection: The received LoRa symbol can be detected by both coherent and non-coherent detection mechanisms.

The coherent detector for LoRa is provided in Fig. 2. For the coherent detection, the inner product of the received symbol vector, $y(n)$ and the transmit symbol vector, $s_{\text{LoRa}} = [s_{\text{LoRa}}(0), s_{\text{LoRa}}(1), \ldots, s_{\text{LoRa}}(M-1)]^T$, i.e., $\langle y, s_{\text{LoRa}} \rangle$ is given as:

$$\langle y, s_{\text{LoRa}} \rangle = \sum_{n=0}^{M-1} y(n)s_{\text{LoRa}}^*(n) = R(k),$$  \hfill (9)

where eq. (9) is attained following the footsteps of eq. (4).

Then using the coherent detection principle presented in eq. (5), the FS of the received symbol, $k$ can be identified as:

$$\hat{k} = \arg \max_k |\Re\{R(k)\}|.$$  \hfill (10)

Thus, for the coherent detection, the received symbol samples, $y(n)$ are first de-chirped using a down-chirp, $c_d(n)$ yielding $r(n) = y(n)c_d(n)$. It is highlighted in the that $r(n)$ corresponds to received $f(n)$ (cf. eq. (7)). Subsequently, $r(n)$ is fed to a DFT evaluator, that results in $R(k)$ for $k = [0, M - 1]$. Note that the detection problem for LoRa is exactly the same as provided in eq. (9).

The non-coherent detector for LoRa is illustrated in Fig.
Following the evaluation of \( R(k) \), the FS, \( \hat{k} \) is ascertained employing the non-coherent detection principle (cf. eq. (6)) as:

\[
\hat{k} = \arg \max_k |R(k)|.
\] (11)

After decimal-to-binary (de2bi) conversion of \( \hat{k} \), the transmitted bit sequence comprising of \( \lambda \) bits is determined.

### B. Interleaved Chirp-Spreading (ICS)-LoRa

One of the first PHY layer inspired approaches to proposed to increase the bit rate relative to LoRa is ICS-LoRa [47]. The authors in [56] investigate the coexistence of LoRa and ICS-LoRa, whereby, it is foreseen that LoRa network can accommodate both LoRa as well as ICS-LoRa. Moreover, in [57], ICS-LoRa is employed as a parallel logical network along with LoRa to enhance the LoRa capacity. ICS-LoRa appends an additional dimension to increasing the bit rate of LoRa by using interleaved versions of the LoRa symbols along with the up-chirp symbols used in LoRa. This, however, only increases the number of bits transmitted per symbol of duration \( T_s \) by 1. To be more precise, the number of bits transmitted per ICS-LoRa symbol are \( \lambda + 1 \) as the interleaved versions of the LoRa symbols can also be transmitted. Besides the fact that ICS-LoRa symbols have a constant envelop, which is optimal for low-cost LPWAN devices, the interleaved LoRa symbols lack orthogonality with the LoRa symbols. Consequently, high correlation peaks appear around three DFT indexes [17], which may contribute to BER degradation.

1) Transmission: The transmitter configuration of ICS-LoRa is presented in Fig. 4. Similar to LoRa, we first push \( \lambda \) of bits towards the bi2de converter, followed by a FS, which yields \( f(n) \) as defined in (7). Then, \( f(n) \) is up-chirped by \( c_a(n) \) resulting in \( s_{\text{LoRa}}(n) \) (as in eq. (8)). Subsequently, \( \lambda_1 = 1 \) bit determines \( \beta_1 \), that dictates whether the transmitter will send the conventional LoRa symbol, \( s_{\text{LoRa}}(n) \) or its interleaved version, i.e., \( \Pi [s_{\text{LoRa}}(n)] \), where \( \Pi [\cdot] \) denotes the interleave operation. \( \beta_1 \) is obtained after bi2de conversion of \( \lambda_1 \), where \( \beta_1 = [0, 2^{\lambda_1} - 1] = \{0, 1\} \). The discrete-time ICS-LoRa transmit symbol is given as:

\[
s_{\text{ICS}}(n) = \begin{cases} s_{\text{LoRa}}(n) & \beta_1 = 0 \\ \Pi [s_{\text{LoRa}}(n)] & \beta_1 = 1 \end{cases}.
\] (12)

Considering \( s_{\text{LoRa}}(n) \) of length \( M \), the interleave (depicted in Fig. 4) performs the following operation:

\[
\Pi [s_{\text{LoRa}}(n)] = \begin{cases} s_{\text{LoRa}}(n) & n = [0, M/4] \\ s_{\text{LoRa}}(n + M/4) & n = [M/4, M/2] \\ s_{\text{LoRa}}(n - M/4) & n = [M/2, 3M/4] \\ s_{\text{LoRa}}(n) & n = [3M/4, M] \end{cases}.
\] (13)

In other words, the interleave operation only shuffles the 2nd with the 3rd quartered portions of \( s_{\text{LoRa}}(n) \).

The symbol energy of a LoRa symbol is \( E_s = 1/M \sum_{n=0}^{M-1} |s_{\text{ICS}}(n)|^2 = 1 \).

2) Detection: In [47], only a non-coherent detector was proposed for ICS-LoRa, however, coherent detection of ICS-LoRa symbols is also possible. Hereby, we present both the coherent detection and non-coherent detection for ICS-LoRa.

The coherent detection architecture for ICS-LoRa is illustrated in Fig. 5. According to eq. (12), it is straightforward to see that LoRa is a particular instance of ICS-LoRa. Therefore, the inner product of the received symbol vector \( y \) and the transmit symbol vector, \( s_{\text{ICS}} = [s_{\text{ICS}}(0), s_{\text{ICS}}(1), \ldots, s_{\text{ICS}}(M-1)]^T \), i.e., \( (y, s_{\text{ICS}}) \) have to be calculated for two different cases, that are, (i) when conventional LoRa symbol is transmitted, i.e., \( \beta_1 = 0 \), and (ii) the interleaved LoRa symbol is transmitted for \( \beta_1 = 1 \).
For case (i), the inner product term $\langle y, s_{\text{ICS}} \rangle$ yields:

$$
\langle y, s_{\text{ICS}} \rangle = \sum_{n=0}^{M-1} y(n) s_{\text{ICS}}
= \sum_{n=0}^{M-1} y(n) \int(n) c_d(n)
= \sum_{n=0}^{M-1} r_1(n) \int(n)
= R_1(k),
$$

(14)

where $r_1(n) = y(n)c_d(n)$. Conversely, when the interleaved LoRa is transmitted, then $\langle y, s_{\text{ICS}} \rangle$ yields:

$$
\langle y, s_{\text{ICS}} \rangle = \sum_{n=0}^{M-1} y(n) s_{\text{ICS}}
= \Pi \{ y(n) \} \int(n) c_d(n)
= \sum_{n=0}^{M-1} r_2(n) \int(n)
= R_2(k),
$$

(15)

where $r_2(n) = \Pi \{ y(n) \} c_d(n)$.

It can be observed from eqs. (14) and (15) that at the receiver, $y(n)$ and an interleaved version of $y(n)$, i.e., $\Pi \{ y(n) \}$ are de-chirped using $c_d(n)$ resulting in $r_1(n) = y(n)c_d(n)$ and $r_2(n) = \Pi \{ y(n) \} c_d(n)$. Subsequently, DFT is evaluated for $r_1(n)$ and $r_2(n)$ that results in $R_1(k)$ and $R_2(k)$, respectively. The next step is to determine whether a LoRa symbol was transmitted, or an interleaved LoRa symbol was transmitted. Thus, two parameters, $\kappa_1 = \max \{ R_1(k) \}$ and $\kappa_2 = \max \{ R_2(k) \}$ are evaluated in the interleave decision block. These parameters help determine $\hat{\beta}_1$ as:

$$
\hat{\beta}_1 = \begin{cases} 
0 & \kappa_1 > \kappa_2 \\
1 & \kappa_1 < \kappa_2
\end{cases},
$$

(16)

which in turn identify the bits transmitting the interleave information, $\hat{\lambda}$ of the transmit symbol after de2bi conversion.

Then, using the same parameters, i.e., $\kappa_1$ and $\kappa_2$, it is also determined that whether $R_1(k)$ or $R_2(k)$ should be used to identify $\hat{k}$ as:

$$
R(k) = \begin{cases} 
R_1(k) & \kappa_1 > \kappa_2 \\
R_2(k) & \kappa_1 < \kappa_2
\end{cases}.
$$

(17)

Finally, the FS $\hat{k}$ using the coherent detection is evaluated as:

$$
\hat{k} = \arg \max_k \{ R(k) \},
$$

(18)

that is used to determine $\hat{\lambda}$ after de2bi conversion.

The non-coherent detector configuration for ICS-LoRa is presented in Fig.[5] It is highlighted that there are two main differences between the coherent detection and the non-coherent detection mechanisms. The first difference is in the interleave decision block, where instead of using $\kappa_1$ and $\kappa_2$, $\kappa_1 = \max \{ |R_1(k)| \}$, and $\kappa_2 = \max \{ |R_2(k)| \}$

used to determine $\hat{\lambda}$ and $R(k)$. The other difference relative to coherent detection is that the FS is identified using the following criterion:

$$
\hat{k} = \arg \max_k \{ |R(k)| \}.
$$

(19)

$\hat{k}$ is then used to determine $\hat{\lambda}$.

C. Extended (E)-LoRa

E-LoRa is another PHY layer approach which aims at increasing the number of transmit bits per symbol relative to LoRa.

In the classical LoRa, the un-chirped symbol, $f(n)$ depicted in (7) confines the information only to the in-phase components. Contrary to that, E-LoRa exploits both in-phase and quadrature axis. Consequently, an additional bit can be incorporated to the E-LoRa symbol relative to LoRa modulation, thus, the total number of bits transmitted per E-LoRa symbol of duration $T_s$ are $\lambda + 1$.

1) Transmission: The transmitter architecture of E-LoRa is depicted in Fig.[7] Like classical LoRa modulation, $\lambda$ bits determine the FS, $\hat{k}$ using which $f(n)$ is obtained as given in eq. (7). Subsequently, $\lambda_{IQ}$ is used to evaluate $\hat{\beta}_{IQ}$ that determines whether the information is confined to the in-phase or the quadrature component of the un-chirped symbol, $\tilde{f}(n)$ as:

$$
\tilde{f}(n) = \begin{cases} 
f(n) & \beta_{IQ} = 0 \\
\beta_{IQ} & \beta_{IQ} = 1
\end{cases}.
$$

(20)

Subsequently, $\tilde{f}(n)$ is chirped using an up-chirp, $c_u(n)$. 

![Fig. 6: Non-coherent detector architecture for ICS-LoRa.](image_url)

![Fig. 7: E-LoRa transmitter architecture.](image_url)
resulting in \( s_{\text{E-LoRa}}(n) = \tilde{f}(n)c_a(n) \), that is mathematically represented as:

\[
s_{\text{E-LoRa}}(n) = \begin{cases} 
\exp \left\{ j \frac{\pi}{M}n(2k + n) \right\} & \beta_{\text{IQ}} = 0 \\
\exp \left\{ j \frac{\pi}{M}n(2k + n) \right\} & \beta_{\text{IQ}} = 1 
\end{cases} \tag{21}
\]

The symbol energy of E-LoRa is \( E_s = \frac{1}{M} \sum_{n=0}^{M-1} |s_{\text{E-LoRa}}(n)|^2 = 1. \)

2) Detection: It may be observed from the transmit symbol structure that the information can be transmitted in the phase of the un-chirped symbol, thus, for E-LoRa, only coherent detection is viable, otherwise the information in the phase cannot be retrieved. The coherent detector architecture of E-LoRa is presented in Fig. 8.\(^7\) Note that in E-LoRa, either the in-phase or the quadrature un-chirped symbols are used, that would result in different inner products between the received symbol vector, \( y(n) \), and the transmit symbol vector, \( s_{\text{E-LoRa}} = [s_{\text{E-LoRa}}(0), s_{\text{E-LoRa}}(1), \cdots, s_{\text{E-LoRa}}(M-1)]^T \), i.e., \( \langle y, s_{\text{E-LoRa}} \rangle \). Therefore, we evaluate \( \langle y, s_{\text{E-LoRa}} \rangle \) considering both cases.

For the first case, when \( \beta_{\text{IQ}} = 0 \), i.e., the in-phase un-chirped symbol is used, then \( \langle y, s_{\text{E-LoRa}} \rangle \) is evaluated as:

\[
\langle y, s_{\text{E-LoRa}} \rangle = \sum_{n=0}^{M-1} y(n)s_{\text{E-LoRa}}(n) = R(k) \tag{22}
\]

On the contrary, when the quadrature component of the un-chirped symbol is used, then \( \langle y, s_{\text{E-LoRa}} \rangle \) yields:

\[
\langle y, s_{\text{E-LoRa}} \rangle = -j \sum_{n=0}^{M-1} r(n)\tilde{f}(n) = -jR(k) \tag{23}
\]

When the real argument of eqs. (22) and (23) is evaluated, we attain the following:

\[
\Re \{ \langle y, s_{\text{E-LoRa}} \rangle \} = \begin{cases} 
\Re \{ R(k) \} & \beta_{\text{IQ}} = 0 \\
\Im \{ R(k) \} & \beta_{\text{IQ}} = 1 
\end{cases} \tag{24}
\]

Thus, for the coherent detection, the received symbol \( y(n) \) is first de-chirped using \( c_d(n) \) that results in \( r(n) \). Afterwards, the DFT of \( r(n) \) is computed that results in \( R(k) \). Moreover, from eq. (20), we can observe that information is either carried by the in-phase component or the quadrature component of the un-chirped symbol. Moreover, eq. (25) dictates that the in-phase and the quadrature components of \( R(k) \) should be separated. This separation is performed by the I/Q separator block, that yields \( \Re \{ R(k) \} \) and \( \Im \{ R(k) \} \). Subsequently, \( \kappa_1 = \max \Re \{ R(k) \} \) and \( \kappa_2 = \max \Im \{ R(k) \} \) are evaluated in the phase evaluator, that determines \( \beta_{\text{IQ}} \), according to the following criterion:

\[
\beta_{\text{IQ}} = \begin{cases} 
0 & \kappa_1 > \kappa_2 \\
1 & \kappa_1 < \kappa_2 
\end{cases} \tag{25}
\]

using which \( \hat{\lambda}_{\text{IQ}} \) is determined after de2bi conversion. Thanks to the phase evaluator block, \( \beta_{\text{IQ}} \) will help identify whether a \( \pi/2 \) rotation was introduced by the transmitter.

Moreover, based on the values of \( \kappa_1 \) and \( \kappa_2 \), \( \hat{R}(k) \) which is required for further evaluation as:

\[
\hat{R}(k) = \begin{cases} 
\Re \{ R(k) \} & \kappa_1 > \kappa_2 \\
\Im \{ R(k) \} & \kappa_1 < \kappa_2 
\end{cases} \tag{26}
\]

Subsequently, the FS, \( \hat{k} \) is identified as:

\[
\hat{k} = \arg \max_k \hat{R}(k). \tag{27}
\]

Then \( \hat{\lambda} \) is determined after de2bi conversion of \( \hat{k} \).

D. Phase-Shift Keying (PSK)-LoRa

All the schemes that are discussed up to this point are capable of transmitting one additional bit relative to LoRa. Unlike these schemes, PSK-LoRa was the first scheme that was proposed that was capable of transmitting more than one additional bit compared to LoRa. Unlike LoRa, where the information bearing element is the FS of the un-chirped symbol, in PSK-LoRa, additional information is also transmitted in the PS of the un-chirped signal by using phase-shift keying (PSK) alphabets \(^{15} \). PSK-LoRa is still a single chip modulation because only one FS is activated. The additional number of bits that can be transmitted via PSK-LoRa depends on the cardinality of the PSK alphabets, \( M_\varphi \). Thus, in addition to \( \lambda \) bits transmitted in the FS, \( \log_2(M_\varphi) \) bits are transmitted in the PS resulting in a total of \( \lambda + \log_2(M_\varphi) \) bits per symbol of duration \( T_s \).

1) Transmission: The transmitter configuration of PSK-LoRa is depicted in Fig. 9. In PSK-LoRa, \( \lambda \) bits determine the FS, \( k \), using which the un-chirped symbol, \( f(n) \) is obtained.
Additionally, $\lambda_{PS} = \log_2(M_{\lambda})$ bits are used to ascertain, $p = [0, 2^{\lambda_{PS}} - 1]$, which results in a PS of

$$\varphi = \exp \left\{ \frac{j2\pi}{2\lambda_{PS}} p \right\}. \quad (28)$$

The resulting discrete time symbol chirped PSK-LoRa symbol is given as:

$$s_{PSK}(n) = f(n) \varphi c_u(n)$$

$$= \exp \left\{ j\pi \left( \frac{2}{M} kn + \frac{2p}{2\lambda_{PS}} + \frac{n^2}{M} \right) \right\}. \quad (29)$$

It is straightforward to see that PSK alphabets have a constant envelope, therefore, the symbol energy of PSK-LoRa is $E_s = \frac{1}{M} \sum_{n=0}^{M-1} |s_{PSK}(n)|^2 = 1$.

2) Detection: Like E-LoRa, PSK-LoRa encodes information in the phase of the symbol, therefore, it is not possible to establish a non-coherent detector, due to the loss of PS information. Thus, coherent detection and a so-called semi-coherent detection are possible. In coherent detection, both the FS and the PS are determined using maximum likelihood (ML) criterion, whereas, in semi-coherent detection, the FS is determined non-coherently, while the PS is estimated based on ML criterion.

The coherent detection configuration for PSK-LoRa is illustrated in Fig. 10. In order to understand the coherent detector design for PSK-LoRa, consider the inner product of the received symbol vector, $\mathbf{y}$ and the transmit symbol. For clarity of exposition, the transmit symbol for PSK-LoRa in vectorial form is given as $s_{PSK} = [s_{PSK}(0), s_{PSK}(1), \cdots, s_{PSK}(M-1)]^T$. Then, the inner product of $\mathbf{y}$ and $s_{PSK}$ is evaluated as:

$$\langle \mathbf{y}, s_{PSK} \rangle = \sum_{n=0}^{N-1} y(n)s_{PSK}(n)$$

$$= \overline{\mathbf{y}}^T \sum_{n=0}^{N-1} y(n)\overline{f(n)c_d(n)}$$

$$= \overline{\mathbf{y}}^T \mathbf{R}(k). \quad (30)$$

The FS, $\hat{k}$, and the PS, $\hat{p}$ using the coherent detection are jointly evaluated as:

$$\hat{k}, \hat{p} = \arg \max_{k,p} \Re \{ \overline{\mathbf{y}}^T \mathbf{R}(k) \} \quad (31)$$

For coherent detection, the received symbol, $y(n)$ is first de-chirped using $c_d(n)$, resulting in $r(n)$. Then using DFT of $r(n)$, $R(k)$ is ascertained. Subsequently, using $R(k)$ and multiplying it with the conjugate of all the possible PSs, the detection problem in eq. (31) is evaluated that yields both $\hat{k}$ and $\hat{p}$. Using de2bi conversion of $\hat{k}$ and $\hat{p}$, the bit sequences of lengths, $\hat{\lambda}$ and $\lambda_{PS}$ are attained.

The semi-coherent detector for PSK-LoRa is presented in Fig. 11. In semi-coherent detection, after evaluating $R(k)$, the FS is estimated using the non-coherent detection criterion as in eq. (6), that results in:

$$\hat{k} = \arg \max_k |R(k)| \quad (32)$$

which is used to determine the bit sequence consisting of $\hat{\lambda}$ bits.

Note that the PS cannot be determined using eq. (32). However, as the activated FS also carries the PS information, therefore, once $\hat{k}$ is known, then, the PS can be evaluated using $\hat{k}$ as:

$$\hat{p} = \arg \max_k \{|R(k)|\} \quad (33)$$

where, $\arg \{ \cdot \}$ is the angle operator, which is also the ML estimate of the PS component. $\hat{p}$ is then used to determine the bit sequence having $\lambda_{PS}$ bits.

E. Dual-Orthogonal Chirp Spread Spectrum (DO-CSS)

DO-CSS was the first multiple chirp modulation scheme that was proposed by Vangelista and Cattapan for LPWANs [49]. In DO-CSS, rather than activating a single FS as in LoRa, one FS corresponding to even frequency, and another FS corresponding to an odd frequency are activated. Since, each FS transmits $\lambda - 1$ bits, therefore, a total of $2\lambda - 2$ bits can be transmitted per symbol of duration $T_s$. Simultaneously activating two FS corresponds to multiplexing two chirp symbols with different cyclic time shifts.

1) Transmission: The transmitter configuration of DO-CSS is illustrated in Fig. 12. Firstly, the $M$ available frequencies
in bandwidth $B$ are divided into $M/2$ even and $M/2$ odd frequencies. Subsequently, using $\lambda_o = \log_2(M/2) = \lambda - 1$ bits, and $\lambda_v = \log_2(M/2) = \lambda - 1$ bits, even and odd FSs, i.e. $k_v$ and $k_o$ are determined respectively. Then using $k_v$ and $k_o$, the un-chirped symbols corresponding to $k_v$ and $k_o$ are determined as:

$$f_1(n) = \exp \left\{ \frac{2\pi}{M} k_v n \right\},$$

and

$$f_2(n) = \exp \left\{ \frac{2\pi}{M} k_o n \right\},$$

respectively. Subsequently, $f_1(n)$ and $f_2(n)$ are added resulting in $f_3(n)$, that multiplexes both $k_v$ and $k_o$ as follows:

$$f_3(n) = f_1(n) + f_2(n) = \exp \left\{ \frac{2\pi}{M} k_v n \right\} + \exp \left\{ \frac{2\pi}{M} k_o n \right\}.$$

Then $f_3(n)$ is multiplied with an up-chirp, $c_d(n)$ resulting in discrete time chirped DO-CSS symbol $s_{DO}(n)$, which is mathematically given as:

$$s_{DO}(n) = f_3(n) c_d(n) = \exp \left\{ \frac{\pi}{M} n (2k_v + n) \right\} + \exp \left\{ \frac{\pi}{M} n (2k_o + n) \right\}.$$

The symbol energy of DO-CSS symbol is $E_s = \frac{1}{M} \sum_{n=0}^{M-1} |s_{DO}(n)|^2 = 2$.

2) Detection: In [49], the authors only proposed the non-coherent detection mechanism for DO-CSS, however, in this work, in addition to the non-coherent detection, we also present the coherent detection for DO-CSS.

Let us first consider the inner product of the received symbol, $y$ and the DO-CSS transmit symbol, $s_{DO} = [s_{DO}(0), s_{DO}(1), \ldots, s_{DO}(M-1)]^T$, i.e., $\langle y, s_{DO} \rangle$, that is given as:

$$\langle y, s_{DO} \rangle = \sum_{n=0}^{M-1} y(n) \bar{s}_{DO}(n)$$

$$= \sum_{n=0}^{M-1} y(n) \bar{f}_3(n)c_d(n)$$

$$= \sum_{n=0}^{M-1} y(n) \{ \bar{f}_1(n) + \bar{f}_2(n) \}$$

$$= R(k_v) + R(k_o).$$

Thus, employing eq. (38), the even and odd activated FS can be evaluated as:

$$\hat{k}_v, \hat{k}_o = \text{arg max}_{k_v, k_o} \{ R(k_v) + R(k_o) \},$$

where $R(k_v)$ and $R(k_o)$ is the DFT output at even and odd frequency indexes. Eq. (39) jointly evaluates both $\hat{k}_v$ and $\hat{k}_o$. Nevertheless, the coherent detection problem in eq. (39) can be implemented in such a way that both $\hat{k}_v$ and $\hat{k}_o$ can be disjointly identified as:

$$\hat{k}_v = \text{arg max}_{k_v} \{ R(k_v) \},$$

and

$$\hat{k}_o = \text{arg max}_{k_o} \{ R(k_o) \},$$

respectively. The coherent detector architecture presented in Fig. 13 shows dis-joint identification of $\hat{k}_v$ and $\hat{k}_o$. In the coherent detection, $r(n)$ is attained after the de-chirping of the received symbol $y(n)$ using a down-chirp, $c_d(n)$. Afterwards the DFT of $r(n)$, i.e. $R(k)$, is evaluated. The even/odd separator isolates the even and odd frequency indexes of $R(k)$ resulting in $R(k_v)$ and $R(k_o)$. Subsequently, using the decision criteria in eqs. (40) and (41), the even and odd activated FSs, $\hat{k}_v$ and $\hat{k}_o$ are respectively identified. Subsequently, the bit sequences, consisting of $\hat{\lambda}_v$ and $\hat{\lambda}_o$ bits are determined after de2bi conversion of $\hat{k}_v$ and $\hat{k}_o$, respectively.

The configuration of non-coherent detector for DO-CSS is provided in Fig. 14. Relative to the coherent detector, the only difference is that after $R(k_v)$ and $R(k_o)$ are obtained, the decision on activated even and odd FSs is made according to:

$$\hat{k}_v = \text{arg max}_{k_v} |R(k_v)|$$

and

$$\hat{k}_o = \text{arg max}_{k_o} |R(k_o)|,$$

Afterwards, using $\hat{k}_v$ and $\hat{k}_o$, the bit sequences having respective lengths of $\hat{\lambda}_v$ and $\hat{\lambda}_o$ are ascertained.

F. Slope-Shift Keying (SSK)-LoRa

In LoRa, only up-chirps are used for spreading the un-chirped symbol, whereas, in SSK-LoRa, in addition to using up-chirps, down-chirps are also used for spreading. Thus, by incorporating the possibility of using either the up-chirp or the down-chirp, in addition to $\lambda$, one additional bit can be transmitted per symbol.
\[ \lambda_s = 1 \text{ bit} \]

![Diagram of SSK-LoRa transmitter architecture.](image)

1) Transmission: The transmitter architecture of SSK-LoRa is presented in Fig. 15. Like LoRa, \( \lambda \) bits are used to determine \( k \) and corresponding un-chirped symbol, \( f(n) \). Moreover, \( \lambda_s = 1 \) bit determines \( \beta_s = [0,2^{\lambda_s} - 1] = \{0,1\} \) which identifies the chirp rate, \( \gamma_s \) as:

\[
\gamma_s = \begin{cases} 
1 & \beta_s = 0 \\
-1 & \beta_s = 1 
\end{cases}
\]

(44)

Then, based on \( \gamma_s \), the spreading symbol \( c_s(n) \) is determined as:

\[
c_s(n) = \begin{cases} 
c_u(n) & \gamma_s = 1 \\
c_d(n) & \gamma_s = -1 
\end{cases}
\]

(45)

\( f(n) \) is then multiplied with the spreading symbol, \( c_s(n) \) resulting in the discrete-time chirped SSK-LoRa symbol, \( s_{\text{SSK}}(n) \), that is given as:

\[
s_{\text{SSK}}(n) = f(n)c_s(n)
\]

Then, the inner product is evaluated as:

\[
\langle y, s_{\text{SSK}} \rangle = \sum_{n=0}^{M-1} y(n)s_{\text{SSK}}(n) = \sum_{n=0}^{M-1} y(n)f(n)c_d(n)
\]

(47)

where \( r_1(n) = y(n)c_d(n) \).

On the other hand, when \( s_{\text{SSK}} = f(n)c_d(n) \), the inner product yields:

\[
\langle y, s_{\text{SSK}} \rangle = \sum_{n=0}^{M-1} y(n)s_{\text{SSK}}(n) = \sum_{n=0}^{M-1} r_1(n)f(n)
\]

(48)

Thus, for coherent detection, once the \( R_1(k) \) and \( R_2(k) \) are evaluated, then, the next step is to determine whether an up-chirp was used at the transmitter or a down-chirp. In order to determine that, we evaluate the following two parameters. \( \kappa_{1}^{\text{coh}} = \max \Re \{R_1(k)\} \) and \( \kappa_{2}^{\text{coh}} = \max \Re \{R_2(k)\} \). Using \( \kappa_{1}^{\text{coh}} \) and \( \kappa_{2}^{\text{coh}} \), the chirp rate estimation block computes \( \hat{\beta}_s \) as:

\[
\hat{\beta}_s = \begin{cases} 
0 & \kappa_{1}^{\text{coh}} > \kappa_{2}^{\text{coh}} \\
1 & \kappa_{1}^{\text{coh}} < \kappa_{2}^{\text{coh}}
\end{cases}
\]

(50)

After \( \text{de2bi} \) conversion of \( \hat{\beta}_s \), \( \hat{\lambda}_s \) is determined.

Once the spreading symbol in known, \( \kappa_{1}^{\text{coh}} \) and \( \kappa_{2}^{\text{coh}} \) are also used to determine whether the transmitted FS belongs to \( R_1(k) \) or \( R_2(k) \) as:

\[
R(k) = \begin{cases} 
R_1(k) & \kappa_{1}^{\text{coh}} > \kappa_{2}^{\text{coh}} \\
R_2(k) & \kappa_{1}^{\text{coh}} < \kappa_{2}^{\text{coh}}
\end{cases}
\]

(51)

Subsequently, using the coherent detection criterion, the FS is identified as:

\[
\hat{k} = \arg \max_k \Re \{R(k)\}
\]

(52)

Once \( \hat{k} \) is estimated, the bit sequence having \( \hat{\lambda} \) bits can be attained after \( \text{de2bi} \) conversion of \( \hat{k} \).

The non-coherent detector configuration for SSK-LoRa is presented in Fig. 17. Compared to coherent detection, there are two primary differences: (i) instead of using \( \kappa_{1}^{\text{coh}} \) and \( \kappa_{2}^{\text{coh}} \), \( \kappa_{1}^{\text{non-coh}} \) and \( \kappa_{2}^{\text{non-coh}} \) are evaluated as \( \kappa_{1}^{\text{non-coh}} = \)
max $|R_1(k)|$ and $\kappa_2^{\text{non-coh}} = \max |R_2(k)|$ are used to determine $\beta_k$ and $R(k)$; and (ii) the detection in the transmit FS $\hat{k}$ is done according to:

$$\hat{k} = \arg \max_k |R(k)|.$$  

(53)

The bit sequence comprising of $\hat{\lambda}$ bits is determined after de2bi conversion of $\hat{k}$.

G. In-Phase and Quadrature (IQ)-CSS

IQ-CSS is another multiple chirp modulation that encodes information bits on both in-phase and quadrature components of the chirp signal [50]. IQCSS makes use of the orthogonality between the sine and cosine waves to concurrently transmit two data symbols. Unlike E-LoRa, where the information is encoded either in the in-phase or the quadrature components, in IQ-CSS, the information is simultaneously encoded in both the in-phase and the quadrature components. In IQ-CSS, one FS corresponding generates the in-phase component of the un-chirped symbol, whereas, another FS is used to generate the quadrature component of the un-chirped symbol. Since both the in-phase and the quadrature components are capable of transmitting $\lambda$ bits, therefore, the total number of bits transmitted per IQ-CSS symbol is twice relative to conventional LoRa, i.e., $2\lambda$.

1) Transmission: The transmitter configuration of IQ-CSS is depicted in Fig. 18 In IQ-CSS, we transmit a total of $2\lambda$ bits, half of which are used to determine $k_i$ and the other half are used for $k_q$. Using $k_i$ and $k_q$, the in-phase, and the quadrature un-chirped symbols, $f_1(n)$ and $f_2(n)$ are generated, that are given as:

$$f_1(n) = \exp \left\{ j \frac{2\pi}{M} k_i n \right\},$$  

(54)

and

$$f_2(n) = \exp \left\{ j \frac{2\pi}{M} k_q n \right\},$$  

(55)

respectively. Subsequently, $f_1(n)$ is multiplied with $c_u(n)$ resulting in $s_1(n)$. On the other hand, $f_2(n)$ is also multiplied with $c_d(n)$ after a phase rotation of $\pi/2$ yielding $s_2(n)$, the chirped in-phase and the quadrature symbols are given as:

$$s_1(n) = f_1(n)c_u = \exp \left\{ j \frac{\pi}{M} n (2k_i + n) \right\},$$  

(56)

and

$$s_2(n) = jf_2(n)c_d = j \exp \left\{ j \frac{\pi}{M} n (2k_q + n) \right\}.$$  

(57)

Lastly, both $s_1(n)$ and $s_2(n)$ are added to attain the discrete-time IQ-CSS symbol as:

$$s_{IQ}(n) = s_1(n) + s_2(n) = \exp \left\{ j \frac{\pi}{M} n (2k_i + n) \right\} + j \exp \left\{ j \frac{\pi}{M} n (2k_q + n) \right\}.$$  

(58)

IQ-CSS symbol energy is equal to $E_{\text{IQ}} = 1/N \sum_{n=0}^{N-1} |s_{IQ}(n)|^2 = 2$.

2) Detection: In [50], the authors have proposed a coherent detector for IQ-CSS, that is illustrated in Fig. 19. The inner product of the received symbol, $y$ and the IQ-CSS transmit symbol $s_{IQ} = [s_{IQ}(0), s_{IQ}(1), \ldots, s_{IQ}(M-1)]^T$, $\langle y, s_{IQ} \rangle$ yields:

$$\langle y, s_{IQ} \rangle = \sum_{n=0}^{M-1} y(n)\bar{s}_{IQ}(n)$$

$$= \sum_{n=0}^{M-1} r(n)f_1(n) - j \sum_{n=0}^{M-1} r(n)f_2(n)$$

$$= R(k_i) - jR(k_q),$$

(59)

where $r(n) = y(n)c_d(n)$. It may be noticed that when the coherent detection (cf. [5]) is applied, the detection problem in eq. (59) can be disjointed, which leads to the following detection of the activated in-phase and quadrature FSs:

$$\hat{k}_i = \arg \max_k \Re \{R(k)\},$$  

(60)

and

$$\hat{k}_q = \arg \max_k \Im \{R(k)\},$$  

(61)

where $\Re \{R(k_i)\} = \Re \{R(k)\}$ and $\Im \{R(k)\} = \Re \{-jR(k_q)\}$. The coherent detector illustrated in Fig.
[19] shows that the received symbol, \( y(n) \) is first de-chirped using a down-chirp, \( c_d(n) \) resulting in \( r(n) \). Subsequently, using the output of the DFT of \( r(n) \), i.e., \( R(k) \), the in-phase and quadrature components are separated using the I/Q separator block. Subsequently, the in-phase and the quadrature FS are identified using eqs. (60) and (61), respectively.

Recently, in [20], the authors have also proposed a non-coherent detector for IQ-CSS, that is shown in Fig. 20. In the non-coherent detector, after evaluating \( R(k) \), non-coherent detection principle (cf. eq. (6)), indexes corresponding to the two largest peaks, i.e., \( k_1 \) and \( k_2 \) of \( R(k) \) are extracted as:

\[
k_1, k_2 = \arg \max_k |R(k)|,
\]

where \(|R(k_1)| > |R(k_2)|\). Subsequently, \( k_1 \) and \( k_2 \) are used to ascertain the FS of the in-phase and the quadrature components, \( \hat{\lambda}_i \) and \( \hat{\lambda}_q \). There can be two possible scenarios: (i) either \( \hat{k}_i \) and \( \hat{k}_q \) are the same, or (ii) \( \hat{k}_i \) and \( \hat{k}_q \) are different. If \( \hat{k}_i \) and \( \hat{k}_q \) are same, this implies that \( R(k_1) \gg R(k_2) \). On the other hand, for different \( \hat{k}_i \) and \( \hat{k}_q \), we are also faced with a problem of matching \( k_1 \) and \( k_2 \) with \( \hat{k}_i \) and \( \hat{k}_q \).

The following criterion will decide whether \( k_1 \) and \( k_2 \) are the same, namely:

\[
1 < \Gamma \leq \frac{|R(k_1)|}{|R(k_2)|},
\]

where \( \Gamma > 1 \) is a predefined threshold. This means that the largest peak is significantly greater than the second largest one, which implies that both FSs align as \( \hat{k}_i = \hat{k}_q = k_1 \).

Now, on the other hand, when \( \hat{k}_i \) and \( \hat{k}_q \) are distinguishable as:

\[
1 \leq \frac{|R(k_1)|}{|R(k_2)|} < \Gamma,
\]

suggesting that both peaks have relatively similar amplitudes. In this scenario, the index detection block also calculates, the phase difference between \( R(k_1) \) and \( R(k_2) \) as:

\[
\theta = \text{phase}\{R(k_1)R(k_2)\},
\]

Then using \( \theta \), the decision on \( \hat{k}_i \) and \( \hat{k}_q \) is made in the index detection block as:

\[
\hat{k}_i = \begin{cases} k_1 & 0 \leq \theta < \pi \\ k_2 & -\pi \leq \theta < 0 \end{cases},
\]

and

\[
\hat{k}_q = \begin{cases} k_2 & 0 \leq \theta < \pi \\ k_1 & -\pi \leq \theta < 0 \end{cases},
\]

respectively.

**H. Discrete Chirp Rate Keying (DCRK)-LoRa**

In SSK-LoRa, there is a possibility of using two chirp rates, i.e., up-chirp or a down-chirp, that adds one additional bit per symbol relative to LoRa. However, it is also possible to extend the framework to use multiple discrete chirp rates to encode higher number of bits compared to SSK-LoRa. The scheme which uses different discrete chirp rates is referred to as DCRK-LoRa [18]. It can also be gathered that DCRK-LoRa is a generalized version of SSK-LoRa. Moreover, it is also accentuated that chirps with different rates are not orthogonal, but encounter a low correlation resulting in lesser intrinsic interference leading to negligible performance loss. This intrinsic interference also reduces with an increase in \( \lambda \). A total of \( M_c \) different chirp rates can be used, therefore, \( \lambda + \log_2(M_c) \) bits are transmitted for DCRK-LoRa symbol of duration, \( T_s \).

1) Transmission: The transmitter configuration of DCRK-LoRa is given in Fig. 21. In DCRK-LoRa, the FS \( k \) is determined after bi2de conversion of \( \lambda \) bits. Using \( k \), the un-chirped symbol, \( f(n) \) is obtained (cf. eq. [5]). In addition to \( \lambda \) bits, \( \lambda_e = \log_2(M_c) \) bits determine the chirp rate. After bi2de conversion of \( \lambda_e, \beta_e = [0, 2^{\lambda_e} - 1] \) of cardinality \( M_c \) is obtained that determines the discrete chirp rate, \( \gamma_c \).

Considering, \( \lambda_e, \gamma_c \) assumes \( M_c \) non-zero integer values using \( \beta_e \) as:

\[
\gamma_c = \begin{cases} \frac{-M_c}{2} & \beta_e = 0 \\ \cdot & \cdot \\ \cdot & \cdot \\ \cdot & \cdot \\ \cdot & \cdot \\ 1 & \beta_e = \frac{2^{\lambda_e}-1}{2} + 1 \\ \cdot & \cdot \\ \cdot & \cdot \\ \cdot & \cdot \\ \cdot & \cdot \end{cases},
\]

where \( c \) is the indexing variable whose values are \( c = [0, M_c - 1] \). Once the chirp rate \( \gamma_c \) is determined, the spreading symbol \( c_c(n) \) is attained as:

\[
c_c(n) = \exp\left\{ j \frac{\pi}{M} \gamma_c n^2 \right\}.
\]

Afterwards, \( f(n) \) is chirped using \( c_c(n) \), resulting in
DCRK-LoRa discrete-time symbol, $s_{DCRK}(n)$ as:

$$s_{DCRK}(n) = f(n) c_c(n) = \exp \left\{ j \frac{\pi}{M} n (2k + \gamma_c n) \right\}.$$  \quad (70)

The symbol energy of DCRK-LoRa symbol is equal to $E_s = 1/M \sum_{n=0}^{M-1} |s_{DCRK}(n)|^2 = 1$.

2) Detection: The coherent detector for DCRK-LoRa is illustrated in Fig. 22. The coherent detector was proposed in \cite{20}. The detection in DCRK-LoRa involves the identification of (i) the FS of the un-chirped symbol, $\hat{k}$; and (ii) the chirp rate, $\hat{\gamma}_c$. In order to develop the coherent detector for DCRK-CSS, consider the inner product of $y$ and the DCRK-CSS transmit symbol $s_{DCRK} = [s_{DCRK}(0), s_{DCRK}(1), \ldots, s_{DCRK}(M - 1)]^T$, $(y, s_{DCRK})$, that yields:

$$\{ y, s_{DCRK} \} = \sum_{n=0}^{M-1} y(n) \overline{f(n)} c_c(n).$$  \quad (71)

where $r_c(n) = y(n) \overline{f(n)}$. Eq. (71) implies that the DFT corresponding to each chirp rate used at the transmitter has to be evaluated. In order to determine the chirp rate, firstly, the parameter, $\hat{\gamma}_c = \max_k \Re \{ R_c(k) \}$ is evaluated. Subsequently, the parameter corresponding to the chirp rate, $c_c(n)$, i.e., $\hat{\beta}_c$ is evaluated as:

$$\hat{\beta}_c = \arg \max_c \Re \{ R_c(k) \},$$  \quad (72)

using which the bit information in the chirp rate, i.e., $\hat{\lambda}_c$ is attained. Then using the DFT output corresponding to $\hat{\beta}_c$, i.e. $R_c(k)$, is used to determine the FS of the un-chirped symbol as:

$$\hat{k} = \arg \max_k \Re \{ R_c(k) \}.$$  \quad (73)

The bit sequence of length $\hat{\lambda}$ is determined after bi2de conversion of $\hat{k}$.

In \cite{18}, the non-coherent detector for DCRK-CSS was proposed. Like the non-coherent detection of SSK-LoRa, in FSCSS-IM transmitter architecture.

1. FSCSS with Index Modulation (FSCSS-IM)

FSCSS-IM was proposed by Hanif and Nguyen in \cite{19}. It is the first approach which amalgamates IM with CSS. In FSCSS-IM, multiple FSs are simultaneously activated for the un-chirped symbol. The FSs are activated based on a predetermined pattern that is referred to as frequency activation pattern (FAP) in this work. It is highlighted that in FSCSS-IM the information bits are transmitted in the FAP rather than activated FS of un-chirped symbol (resp. cyclic-time shift of chirped symbol) as in LoRa. It may be noted that the chirp IM approach presented in \cite{20} is same as FSCSS-IM. FSCSS-IM is capable of transmitting higher number of bits per symbol because the number of possible FAPs are enormous. It is accentuated that generally two to three FSs in the FAP are activated because activating a higher number of FS may reduce the EE and considerably increase the complexity of the system. CSS approach in which all the chirps are multiplexed is referred to as orthogonal chirp-division multiplexing that is investigated in \cite{58}. On the other hand, the amalgamation of IM with OCDM results in OCDM-IM that has been elucidated in \cite{59}.

1) Transmission: The transmitter architecture of FSCSS-IM is illustrated in Fig. 24. It is highlighted that a total of $M$ possible FSs can be activated. The set of these available FSs is denoted by $\Omega = \{0, 1, \ldots, M - 1\}$. Consider that $c$ FSs are to be activated. Then, bi2de conversion of $\lambda_{IM} = \log_2(M_c)$ bits results in $\lambda_{IM} = \{0, 2^{\lambda_{IM}} - 1\}$, which determines the set of FSs to be activated, i.e., FAP, $\sigma = \{\sigma_1, \sigma_2, \ldots, \sigma_c\} \in \Omega$. Using
different combinatorial algorithms (see [60] for details), \( z_{1M} \) is bijectively mapped to a unique \( \sigma \), where the set \( \sigma \) contains of all the FSs to be activated, i.e., \( \sigma = \{ k_1, k_2, \cdots , k_c \} \). Then, the un-chirped symbol is obtained as:

\[
 f^{1M}(n) = \sum_{k \in \sigma} \exp \left\{ \frac{2\pi}{M} km \right\} . \tag{75}
\]

It is highlighted that \( f^{1M}(n) \) comprises of multiple activated FSs. Subsequently, \( f^{1M}(n) \) is chirped using an up-chirp symbol as:

\[
 s^{1M}_{FS}(n) = f^{1M}(n)c_d(n) = \sum_{k \in \sigma} \exp \left\{ \frac{2\pi}{M} n(k + n) \right\} . \tag{76}
\]

The symbol energy of FSCSS-IM symbol is equal to \( E_{a} = \frac{1}{M} \sum_{n=0}^{M-1} |s^{1M}_{FS}(n)|^2 = \varsigma \).

2) Detection: The coherent detector for FSCSS-IM is presented in Fig. 25. The inner product of \( y \) and the FSCSS-IM transmit symbol \( s^{1M}_{FS} = [s^{1M}_{FS}(0), s^{1M}_{FS}(1), \cdots , s^{1M}_{FS}(M-1)]^T \), i.e., \( \langle y, s^{1M}_{FS} \rangle \) results in:

\[
 \langle y, s^{1M}_{FS} \rangle = \sum_{n=0}^{M-1} y(n)s^{1M}_{FS}(n) = \sum_{n=0}^{M-1} y(n)f^{1M}(n)c_d(n) = \sum_{n=0}^{M-1} r(n)f^{1M}(n) = \sum_{k \in \sigma} R(k) . \tag{77}
\]

Thus, using eq. (77) and the coherent detection principle (cf. eq. [4]), the \( \varsigma \) FSs, i.e., FAP is determined by considering the indexes of the \( \varsigma \) highest peaks in the real components of \( R(k) \) as:

\[
 \hat{\sigma} = \arg \max_{k \in \sigma} \Re \left\{ \sum_{k \in \sigma} R(k) \right\} . \tag{78}
\]

The coherent detection configuration shows that after de-chirping the received symbol \( y(n) \) using a down-chirp, \( c_d(n) \), \( r(n) \) is obtained. The latter is then fed to the DFT, which yields \( R(k) \). Subsequently, \( \varsigma \) highest peaks in the real part of \( R(k) \) are determined using the decision criterion as in eq. (65) which results in \( \hat{\sigma} \), using which the transmitted bits, \( \lambda_{1M} \) are determined after \( \text{de2bi} \) conversion of output of index demapping algorithm, i.e., \( \hat{z}_{1M} \).

The non-coherent detector for FSCSS-IM is illustrated in Fig. 26. The transmitter architecture of IQ-CIM is depicted in Fig. 27. After \( \text{bi2de} \) conversion of \( \lambda_{1M} \) is IQ-CIM, where the authors have combined IQ-CSS with IM. Unlike FSCSS-IM, the IQ-CSS applies IM for both in-phase and quadrature un-chirped symbols, resulting in a more flexible modulation compared to FSCSS, at the cost of increased computational complexity. In principle, IQ-CIM consists of two FSCSS-IM branches; one for in-phase components and another for quadrature components. Thus, two distinct FAPs are needed for the in-phase and the quadrature un-chirped symbols. As the information is transmitted in FAP, and IQ-CIM requires two FAPs, the number of bits that can be transmitted by IQ-CIM are double relative to FSCSS-IM provided that the number of activated FSs used in the FAPs are the same.

1) Transmission: The transmitter architecture of IQ-CIM is depicted in Fig. 27. After \( \text{bi2de} \) conversion of \( \lambda_{1M} = \lfloor \log_2 \left( \varsigma_{i} \right) \rfloor \) and \( \lambda_{1M} = \lfloor \log_2 \left( \varsigma_{q} \right) \rfloor \) bits, two integers, \( z_i \) and \( z_q \), are determined using the decision criterion as in eq. (79).
and \( z_q \) are obtained, respectively, where \( \varsigma_i \) and \( \varsigma_q \) are the number of activated FSs for the in-phase and the quadrature un-chirped symbol. When \( z_i \) and \( z_q \) are input to the the index selector block, two distinct FAPs for in-phase and quadrature components, \( \sigma_i \in \Omega \) and \( \sigma_q \in \Omega \) are attained, where \( \Omega = \{0, 1, \ldots, M - 1\} \) is the set of all possible FSs. We assume that \( \varsigma_i = \varsigma_q = \varsigma \leq M/2 \) FSs are activated for both the in-phase and the quadrature components, then \( \sigma_i = \{k_{i1}, k_{i2}, \ldots, k_{i\varsigma}\} \) and \( \sigma_q = \{k_{q1}, k_{q2}, \ldots, k_{q\varsigma}\} \). Then, by activating all the FSs in \( \sigma_i \) and \( \sigma_q \), the un-chirped in-phase and quadrature symbols are obtained as:

\[
f_1(n) = \sum_{k_i \in \sigma_i} \exp \left\{ j \frac{2\pi}{M} k_i n \right\},
\]

and

\[
f_2(n) = \sum_{k_q \in \sigma_q} \exp \left\{ j \frac{2\pi}{M} k_q n \right\},
\]

respectively. Then, \( f_1(n) \) and \( f_2(n) \) are chirped using an up-chirp symbol to attain the chirped in-phase and quadrature symbols, given as:

\[
s_1(n) = f_1(n) c_{u}(n) = \sum_{k_i \in \sigma_i} \exp \left\{ j \frac{2\pi}{M} n (k_i + n) \right\},
\]

and

\[
s_2(n) = f_2(n) c_{u}(n) = \sum_{k_q \in \sigma_q} \exp \left\{ j \frac{2\pi}{M} n (k_q + n) \right\},
\]

respectively. Subsequently, the chirped in-phase and quadrature chirped symbols are combined to attain IQ-CIM symbol as:

\[
s_{\text{IQ}}^\text{IM}(n) = s_1(n) + j s_2(n) = \sum_{k_i \in \sigma_i} \exp \left\{ j \frac{2\pi}{M} n (k_i + n) \right\} + \sum_{k_q \in \sigma_q} \exp \left\{ j \frac{2\pi}{M} n (k_q + n) \right\}.
\]

Considering that \( \varsigma_i = \varsigma_q = \varsigma \) FSs are activated for both the in-phase and the quadrature components, then, the symbol energy of IQ-CIM symbol is \( E_n = 1/M \sum_{n=0}^{M-1} |s_{\text{IQ}}^\text{IM}(n)|^2 = 2\varsigma \).

2) Detection: It is observed that in IQ-CIM, the information is transmitted using IM in both the in-phase and the quadrature components, therefore, only coherent detection is applicable. Moreover, it is highlighted that the non-coherent detector in IQ-CSS is not directly applicable on IQ-CIM. The coherent detector configuration for IQ-CIM is presented in Fig. 28. The inner product of \( y \) and the IQ-CIM transmit symbol \( s_{\text{IQ}}^\text{IM} = [s_{\text{IQ}}^\text{IM}(0), s_{\text{IQ}}^\text{IM}(1), \ldots, s_{\text{IQ}}^\text{IM}(M - 1)]^T \), i.e., \( \langle y, s_{\text{IQ}}^\text{IM} \rangle \) results in:

\[
\langle y, s_{\text{IQ}}^\text{IM} \rangle = \sum_{n=0}^{M-1} y(n) s_{\text{IQ}}^\text{IM}(n) = \frac{1}{M-1} \left( \sum_{n=0}^{M-1} y(n) f_1(n) c_u(n) - j \sum_{n=0}^{M-1} y(n) f_2(n) c_u(n) \right)
\]

\[
= \frac{1}{M-1} \left( \sum_{n=0}^{M-1} r(n) f_1(n) - j \sum_{n=0}^{M-1} r(n) f_2(n) \right)
\]

\[
= \frac{1}{M-1} \left( \sum_{k_i \in \sigma_i} R(k_i) - j \sum_{k_q \in \sigma_q} R(k_q) \right). \tag{85}
\]

Using eq. (85), the activated in-phase and quadrature FAPs, \( \sigma_i \) and \( \sigma_q \) using coherent detection can be evaluated as:

\[
\hat{\sigma}_i = \arg \max_{k_i \in \sigma_i} \Re \left\{ \sum_{k_i \in \sigma_i} R(k_i) \right\}, \tag{86}
\]

and

\[
\hat{\sigma}_q = \arg \max_{k_q \in \sigma_q} \Im \left\{ \sum_{k_q \in \sigma_q} R(k_q) \right\}, \tag{87}
\]

respectively, where \( \Re \left\{ \sum_{k_i \in \sigma_i} R(k_i) \right\} = \Re \left\{ \sum_{k_q \in \sigma_q} R(k_q) \right\} \) and \( \Im \left\{ \sum_{k_i \in \sigma_i} R(k_i) \right\} = \Im \left\{ \sum_{k_q \in \sigma_q} R(k_q) \right\} \). \( \hat{\sigma}_i \) and \( \hat{\sigma}_q \) are then used for index demapping that yields \( \hat{z}_i \) and \( \hat{z}_q \). Finally, both indices are precessed to determine the transmitted bit sequences having \( \hat{\lambda}_{\text{IM}}^i \) and \( \hat{\lambda}_{\text{IM}}^q \) bits after de2bi conversion.

K. SSK-ICS-LoRa

SSK-ICS-LoRa combines SSK-LoRa and ICS-LoRa. It is recalled that in SSK-LoRa uses both the up-chirp symbols and down-chirp symbol to increase the number of transmit bits by 1 relative to LoRa. On the other hand, by using the up-chirp symbols and their interleaved version, ICS-LoRa increases the number of transmitted bits by 1 relative to LoRa. By combining both SSK-LoRa and ICS-LoRa, SSK-ICS-LoRa uses up-chirp symbols, down-chirp symbols, and interleaved
versions of both to increase the number of transmitted bits per symbol by 2 relative to LoRa.

1) Transmission: The transmitter architecture of SSK-ICS-LoRa is provided in Fig. 29. Like classical LoRa, \( \lambda \) bits determine the FS, \( k \), using which the un-chirped symbol, \( f(n) \). Additionally, \( \lambda_n = 1 \) bit after bi2de results in \( \beta_n = \{0, 1\} \) which identifies, the chirp rate, \( \gamma_n = \{-1, 1\} \) (cf. eq. 44). It is highlighted that only two different chirp rate are possible as in SSK-LoRa. Once \( \gamma_n \) is determined, the chirp symbol, \( c_n(n) \) is obtained (cf. eq. 45). The chirped symbol, \( s_{SSK}(n) \) is obtained by multiplying the \( f(n) \) and \( c_n(n) \) (cf. eq. 46). It may be noticed that \( s_{SSK}(n) \) is same the SSK-LoRa symbol. Once \( s_{SSK}(n) \) is determined, \( \lambda_I \) identifies whether it has to be interleaved or not. After bi2de conversion of \( \lambda_I = 1 \) bit, another parameter, i.e., \( \beta_I \) is ascertained, that identifies whether the transmitter will transmit the conventional SSK-LoRa symbol, \( s_{SSK}(n) \) or its interleaved version, i.e., \( \Pi[s_{SSK}(n)] \). The discrete-time SSK-ICS-LoRa transmit symbol is given as:

\[
s_{SSK-ICS}(n) = \begin{cases} s_{SSK}(n) & \beta_I = 0 \\ \Pi[s_{SSK}(n)] & \beta_I = 1 \end{cases} \quad (88)
\]

The symbol energy of SSK-ICS-LoRa symbol is \( E_n = 1/M \sum_{n=0}^{M-1} |s_{SSK-ICS}(n)|^2 = 1 \).

2) Detection: The coherent detector for SSK-ICS-LoRa is presented in Fig. 30. It is observed that in SSK-ICS-LoRa, there are four different possibilities for the transmit symbol: (i) up-chirp symbol is used for spreading with no interleaving, i.e., \( \{\beta_s, \beta_I\} = \{0, 0\} \); (ii) down-chirp symbol is used for spreading with no interleaving, i.e., \( \{\beta_s, \beta_I\} = \{1, 0\} \); (iii) up-chirp symbol is used for spreading with interleaving, i.e., \( \{\beta_s, \beta_I\} = \{0, 1\} \); and (iv) down-chirp symbol is used for spreading with interleaving, i.e., \( \{\beta_s, \beta_I\} = \{1, 1\} \). In all these cases the inner product, \( \langle y, s_{SSK-ICS} \rangle \) of the received symbol, \( y \) and the SSK-ICS-LoRa transmit symbol, \( s_{SSK-ICS} = \left[ s_{SSK-ICS}(0), s_{SSK-ICS}(1), \ldots, s_{SSK-ICS}(M-1) \right] \) are different. For the first case when , i.e., \( \{\beta_s, \beta_I\} = \{0, 0\} \),

\[
\langle y, s_{SSK-ICS} \rangle \text{ evaluates to:}
\]

\[
\langle y, s_{SSK-ICS} \rangle = \sum_{n=0}^{M-1} y(n)\overline{s_{SSK-ICS}(n)} = \sum_{n=0}^{M-1} y(n)f(n)c_d(n) \quad (89)
\]

where \( R_1(k) \) is the DFT of \( r_1(n) = y(n)c_d(n) \). When \( \{\beta_s, \beta_I\} = \{1, 0\} \), \( \langle y, s_{SSK-ICS} \rangle \) evaluates to:

\[
\langle y, s_{SSK-ICS} \rangle = \sum_{n=0}^{M-1} y(n)\overline{s_{SSK-ICS}(n)} = \sum_{n=0}^{M-1} r_1(n)f(n)c_u(n) \quad (90)
\]

where \( R_2(k) \) is the DFT of \( r_2(n) = y(n)c_u(n) \). Considering up-chirp symbol for spreading with interleaving, i.e., \( \{\beta_s, \beta_I\} = \{0, 1\} \), then \( \langle y, s_{SSK-ICS} \rangle \) results in:

\[
\langle y, s_{SSK-ICS} \rangle = \sum_{n=0}^{M-1} y(n)\overline{s_{SSK-ICS}(n)} = \sum_{n=0}^{M-1} \Pi[y(n)]f(n)c_d(n) \quad (91)
\]

where \( R_3(k) \) is the DFT of \( r_3(n) = y(n)c_d(n) \). Considering down-chirp symbol for spreading with interleaving, i.e., \( \{\beta_s, \beta_I\} = \{1, 1\} \), then \( \langle y, s_{SSK-ICS} \rangle \) results in:

\[
\langle y, s_{SSK-ICS} \rangle = \sum_{n=0}^{M-1} y(n)\overline{s_{SSK-ICS}(n)} = \sum_{n=0}^{M-1} r_3(n)f(n) = R_3(k)
\]
where $R_4(k)$ is the DFT of $r_4(n) = \Pi[y(n)]c_4(n)$. Lastly, considering $\{\beta_s, \beta_l\} = \{1, 1\}$, the inner product yields:

$$\langle y, s_{SSK-ICS} \rangle = \sum_{n=0}^{M-1} y(n)\overline{s_{SSK-ICS}(n)} = \sum_{n=0}^{M-1} \Pi[y(n)]\overline{T(n)}c_u(n)$$

$$= \sum_{n=0}^{M-1} r_4(n)\overline{T(n)}$$

$$= R_4(k),$$

where $R_4(k)$ is the DFT of $r_4(n) = \Pi[y(n)]c_u(n)$. The next step is to use eqs. (89), (90), (91) and (92) to determine: (i) whether up-chirp or down-chirp is used for spreading at the transmitter, and (ii) the chirped symbol was interleaved or not. Therefore, the following parameters are evaluated:

$$\kappa_1^\text{coh} = \max \Re \{R_1(k)\}, \quad \kappa_2^\text{coh} = \max \Re \{R_2(k)\}, \quad \kappa_3^\text{coh} = \max \Re \{R_3(k)\}, \quad \kappa_4^\text{coh} = \max \Re \{R_4(k)\}.$$  

Then using these parameters, $\hat{\beta}_s$ and $\hat{\beta}_l$ are evaluated as:

$$\hat{\beta}_s = \begin{cases} 0 & \max\{\kappa_1^\text{coh}, \kappa_3^\text{coh}\} > \max\{\kappa_2^\text{coh}, \kappa_4^\text{coh}\} \\ 1 & \max\{\kappa_2^\text{coh}, \kappa_4^\text{coh}\} > \max\{\kappa_1^\text{coh}, \kappa_3^\text{coh}\} \end{cases}$$

(93)

and

$$\hat{\beta}_l = \begin{cases} 0 & \max\{\kappa_1^\text{coh}, \kappa_2^\text{coh}\} > \max\{\kappa_3^\text{coh}, \kappa_4^\text{coh}\} \\ 1 & \max\{\kappa_3^\text{coh}, \kappa_4^\text{coh}\} > \max\{\kappa_1^\text{coh}, \kappa_2^\text{coh}\} \end{cases}$$

(94)

respectively. Subsequently, $\hat{\lambda}$ and $\hat{\lambda}$ are used to determine $\hat{\lambda}$ and $\hat{\lambda}_1$. Once, $\hat{\beta}_s$ and $\hat{\beta}_l$ are evaluated, then the FS, $\hat{k}$ is identified coherently as:

$$\hat{k} = \arg\max_k \Re \{R(k)\},$$

(95)

where

$$R(k) = \begin{cases} R_1(k) & \{\hat{\beta}_s, \hat{\beta}_l\} = \{0, 0\} \\ R_2(k) & \{\hat{\beta}_s, \hat{\beta}_l\} = \{1, 0\} \\ R_3(k) & \{\hat{\beta}_s, \hat{\beta}_l\} = \{0, 1\} \\ R_4(k) & \{\hat{\beta}_s, \hat{\beta}_l\} = \{1, 1\} \end{cases}$$

(96)

The transmit bit sequence consisting of $\lambda$ bits is determined after de2bi conversion of $\hat{k}$. The non-coherent detector for SSK-ICS-LoRa is illustrated in Fig. 31. There are a couple of difference in the coherent and the non-coherent detector after the evaluation of $R_1(k), R_2(k), R_3(k)$, and $R_4(k)$: (i) the interleave and the chirp rate decision on $\hat{\beta}_s$ and $\hat{\beta}_l$ is made based on parameters calculated as $\kappa_1^\text{non-coh} = \max \Re \{R_1(k)\}, \kappa_2^\text{non-coh} = \max \Re \{R_2(k)\}, \kappa_3^\text{non-coh} = \max \Re \{R_3(k)\},$ and $\kappa_4^\text{non-coh} = \max \Re \{R_4(k)\}$, and (ii) the FS $\hat{k}$ is non-coherently determined as:

$$\hat{k} = \arg\max_k |R(k)|,$$

(97)

where $R(k)$ is same as in eq. (96).

L. Enhanced PSK-CSS (ePSK-CSS)

ePSK-CSS extends PSK-LoRa by incorporating a simple yet effective redundancy to improve the detection of FSs at the receiver [52]. In ePSK-CSS, the available bandwidth is divided into sub-bands. The so-called fundamental frequency (FF) for the first sub-band and its harmonics in each subsequent sub-bands are activated, and a different PS is introduced to each of them resulting in an un-chirped symbol. ePSK-CSS is a multiple chirp modulation as more than one FS are simultaneously activated. The symbol is chirped resulting in ePSK-CSS symbol. The SE of ePSK-CSS depends on the number of sub-bands and the PSs that are introduced in each sub-band. Based on different combinations of number of sub-bands and the number of PSs, a number of different variants of ePSK-CSS can be obtained. The optimal variant has two sub-bands and quaternary PSs are possible for each sub-band. This optimal variant of ePSK-CSS is capable of transmitting 3 additional bits relative to LoRa by using quaternary PS and two sub-bands.

1) Transmission: The transmitter architecture of ePSK-CSS is illustrated in Fig. 32. Consider that $M$ frequencies are available in $B$, which is divided into $N_b$ sub-bands, where, each sub-band contains $\alpha = M/N_b$ distinct frequencies. The so-called FF, $k \in [0, \alpha - 1]$ defined for the first sub-band encodes $\lambda_{FS} = \log_2(\alpha)$ bits. Using the remaining $N_b - 1$ sub-bands, redundancy is introduced by activating the harmonics of the FF, $k$ as $k + \alpha, k + 2\alpha, \cdots, k + (N_b - 1)\alpha$. Therefore, for the detection of $k$, i.e., the FF, there are essentially $N_b$
copies. The un-chirped symbol representing the activation of the FF and its harmonics is given as:
\[
f(n) = \exp \left\{ \frac{2 \pi}{M} n \left( k + l \left\lfloor \frac{M}{N_b} \right\rfloor \right) \right\},
\]
where \( l = [1, N_b] \). Moreover, a PS is introduced for the FF and \( l \)th harmonics, which is given as:
\[
\varphi_{l,p} = \exp \left\{ j \frac{2 \pi}{M} p_l \right\},
\]
where \( p_l \in [0, 2^{b_{\text{PS}}}-1] \) with \( b_{\text{PS}} \) being the number of bits encoded in the PSs. Then, the discrete time ePSK-CSS symbol obtained by multiplying \( f(n) \), \( \varphi_{l,p} \), and the up-chirp, \( c_u \) that is given as:
\[
s_{e\text{PSK}}(n) = \sum_{l=0}^{N_b-1} \sum_{l=0}^{N_b-1} f(n) \varphi_{l,p} c_u(n)
= \sum_{l=0}^{N_b-1} \exp \left\{ j \pi \left( \frac{2kn}{M} + \frac{2ln}{N_b} + \frac{2p_l}{2^{b_{\text{PS}}}} + n^2 \right) \right\}.
\]
Note that the information carrying elements are \( k \) and \( p_l \). Moreover, the symbol energy of ePSK-CSS is \( E_s = \left( \sum_{n=0}^{N_b-1} s_{e\text{PSK}}(n)^2 \right)^2 = N_b \).

2) Detection: The coherent detector for ePSK-CSS is presented in Fig. 33. The inner product of the received symbol, \( y \) and the ePSK-CSS transmit symbol, \( s_{e\text{PSK}} = [s_{e\text{PSK}}(0), s_{e\text{PSK}}(1), \cdots, s_{e\text{PSK}}(M-1)]^T \) is evaluated as:
\[
\langle y, s_{e\text{PSK}} \rangle = \sum_{n=0}^{M-1} \sum_{l=0}^{N_b-1} y(n) s_{e\text{PSK}}(n)
= \sum_{n=0}^{M-1} \sum_{l=0}^{N_b-1} y(n) \overline{f(n)} \varphi_{l,p} c_u(n)
= \sum_{n=0}^{M-1} \sum_{l=0}^{N_b-1} r(n) \overline{f(n)} \varphi_{l,p}
= \varphi_{l,p} \sum_{l=0}^{N_b-1} R \left( k + l \left\lfloor \frac{M}{N_b} \right\rfloor \right),
\]
where \( r(n) = y(n) c_u(n) \), and \( R(k + l(M/N_b)) \) is the DFT of \( r(n) \) evaluated at \( k + l(M/N_b) \) index. Thus, using coherent detection, the activated FF and the PS for the \( N_b \) sub-bands can be jointly estimated as:
\[
\hat{k}, \hat{p}_l = \arg \max_{k,p_l} \Re \left\{ \varphi_{l,p} \sum_{l=0}^{N_b-1} R \left( k + l \left\lfloor \frac{M}{N_b} \right\rfloor \right) \right\}.
\]

As seen from Fig. 33, the received signal is first correlated with the down-chirp, then DFT is evaluated, before multiplying it with conjugate of all the possible PSs. Finally the sum over \( N_b \) is evaluated to determine \( k \) and \( p_l \) for all the sub-bands.

The semi-coherent detector for ePSK-CSS is provided in Fig. 34. In the semi-coherent detector for ePSK-CSS, the FF is evaluated non-coherently, whereas the PSs are determined coherently. The FF from the received ePSK-CSS symbol is identified as:
\[
\hat{k} = \arg \max_{k} \left| \sum_{l=0}^{N_b-1} R \left( k + l \left\lfloor \frac{M}{N_b} \right\rfloor \right) \right|.
\]

Note that \( \hat{k} \) only estimates the FF not the PSs on the sub-bands. So, once the FF is obtained, the PSs for each sub-band can be determined by decoding \( R(k + l(M/N_b)) \) into the phase discriminator as:
\[
\hat{p}_l = \text{angle} \left\{ R \left( k + l \left\lfloor \frac{M}{N_b} \right\rfloor \right) \right\},
\]
for \( l = 0, 1, \cdots, N_b-1 \), where \( \text{angle}\{\cdot\} \) represents the phase discriminator. The function of the phase discriminator is to identify the PS on \( R(k + l(M/N_b)) \). It may be noticed that unlike ML detection, where the FS and PSs are detected jointly, in the semi-coherent detection, the identification of FS and PSs is disjoint. Fig. 34 implies that after the evaluation of \( \hat{k} \), firstly the FF is identified using eq. (103) that computes the average of \( R(k + l(M/N_b)) \) over \( N_b \) sub-blocks to attain redundancy benefits and evaluates the energy of the averaged symbol. Subsequently, once the FF is known, the PSs on the different sub-bands is determined using eq. (104).

M. Group-based CSS (GCSS)

GCSS has been proposed in [33] and is somewhat similar to ePSK-CSS in a sense that the available FSs are divided
into different groups (sub-bands for ePSK-CSS). Once the FSSs are divided into different groups, one FS is activated from each group. Thus, there is a flexibility of trade-off between EE and SE using the configurable parameter, the group number. Basically, the group number represents the number of groups in which the available FSSs are divided. For a high group number value, the SE will be higher and EE would be diminished as a large number of FSSs would be activated. On the contrary, for lower group number values, the EE would be higher and SE would be less.

1) Transmission: The transmitter architecture of GCSS is depicted in Fig. 35. The $M$ available frequencies available in the bandwidth, $B$ are divided into $G$ groups, where each group contains $M/G$ frequencies. The FSSs available in the $g$th group have indexes $k_g = \{[(g-1)(M/G), g(M/G) - 1]\}$, where $g \in [1, G]$. Moreover, $\lambda_g = \log_2(M/G)$ bits are used to determine the activated FS for the $g$th group. For the first group, $\lambda_1$ is used to determine the activated FSSs, $k_1$ from $k_1 \in [0, M/G - 1]$ after bi2de conversion. Similarly, $\lambda_G$ determines the activated FS from the $G$th group from $k_G = \{(G-1)(M/G), M-1\}$. Once all the FSSs from the $G$ groups are determined, their corresponding un-chirped symbols $f_1(n), f_2(n), \ldots, f_G(n)$. The un-chirped symbol for the $g$th group is obtained as:

$$f_g(n) = \exp \left\{ j \frac{2\pi}{M} k_g n \right\}.$$  \hspace{1cm} (105)

Subsequently, the un-chirped symbols from all the $G$ groups are added together resulting in:

$$\hat{f}(n) = \sum_{i=1}^{G} f_i(n) = \sum_{i=1}^{G} \exp \left\{ j \frac{2\pi}{M} k_i n \right\}. \hspace{1cm} (106)$$

$\hat{f}(n)$ is then chirped using an up-chirp, $c_u(n)$ yielding a discrete-time GCSS symbol, $s_{GCSS}(n)$, that is given as:

$$s_{GCSS}(n) = \hat{f}(n) c_u(n) = \sum_{i=1}^{G} \exp \left\{ j \frac{\pi}{M} n (2k_i + n) \right\}. \hspace{1cm} (107)$$

The symbol energy of GCSS is $E_s = 1/M \sum_{n=0}^{M-1} |s_{GCSS}(n)|^2 = G$.

2) Detection: The authors in [14] only proposed a coherent detector for GCSS that is shown in Fig. 36. In the non-coherent detector, the FSSs from all the groups are determined non-coherently as:

$$\hat{k}_i = \arg \max_{k \in k_i} |R(k)|, \hspace{1cm} (108)$$

where $k_i = \{(i-1)(M/G), \ldots, (iM/G) - 1\}$ for $i \in [1, G]$. Eq. (108) implies that G peaks having the highest amplitudes in the index range for each group identified by $k_i$ needs to be evaluated.

N. Time-Domain Multiplex-CSS (TDM-CSS)

The main idea of TDM-CSS is to simultaneously incorporate both up-chirps and down-chirps yielding a time-domain multiplex LoRa symbol [14]. Unlike SSK-LoRa, where chirp is used to encode information, TDM-CSS uses chirping to multiplex LoRa waveforms. Moreover, it is important to highlight that in SSK-LoRa only one FS is used (single chirp), whereas in TDM-CSS two distinct FSSs are used: one with an up-chirp and the other with a down-chirp. One un-chirped symbol is multiplied with an up-chirp and the other is multiplied with a down-chirp. Then, the two chirped symbols are added together resulting in TDM-CSS symbol. It is evident that the number of bits transmitted in the TDM-CSS symbol is twice relative to LoRa and is the same as that of IQ-CSS.

1) Transmission: The transmitter configuration of TDM-CSS is presented in Fig. 37. There are two bit sequences consisting of $\lambda \in \log_2(M)$ bits each. These sequences are used to determine two FSSs, that are $k_1$ and $k_2$. Once the FSSs are attained, two un-chirped symbol corresponding to these FSSs are attained as:

$$f_1(n) = \exp \left\{ j \frac{2\pi}{M} k_1 n \right\}, \hspace{1cm} (109)$$
and
\[ f_2(n) = \exp \left\{ j \frac{2\pi}{M} k_2 n \right\}, \]
respectively. Subsequently, \( f_1(n) \) is multiplied with an up-chirp, \( c_u(n) \), \( f_2(n) \) is multiplied with a down-chirp, \( c_d(n) \) resulting in two distinct chirped symbols, \( s_1(n) \) and \( s_2(n) \) which in discrete-time are given as:
\[ s_1(n) = f_1(n)c_u(n) = \exp \left\{ j \frac{\pi}{M} n (2k_1 + n) \right\}, \quad (111) \]
and
\[ s_2(n) = f_2(n)c_d(n) = \exp \left\{ j \frac{\pi}{M} n (2k_2 - n) \right\}. \quad (112) \]

Lastly, the two chirped symbols, \( s_1(n) \) and \( s_2(n) \) are added producing the TDM-CSS symbol, \( s_{TDM}(n) \) that is given as:
\[
\begin{align*}
    s_{TDM}(n) &= s_1(n) + s_2(n) \\
    &= \exp \left\{ j \frac{\pi}{M} n (2k_2 - n) \right\} + \exp \left\{ j \frac{\pi}{M} n (2k_1 + n) \right\}.
\end{align*}
\quad (113)
\]

The symbol energy of TDM-CSS symbol is
\[
E_s = \frac{1}{M} \sum_{n=0}^{M-1} |s_{TDM}(n)|^2 = 2.
\]

2) Detection: In [54], the authors have proposed the non-coherent detection for TDM-CSS, however, the coherent detection for TDM-CSS is also possible. Hereby, we shall provide both the coherent and the non-coherent detection mechanisms for TDM-CSS.

The coherent detector architecture for TDM-CSS is shown in Fig. 38. The inner product of the received symbol, \( y(n) \) and the transmit symbol, \( s_{TDM} = [s_{TDM}(0), s_{TDM}(1), \ldots, s_{TDM}(M-1)]^T \), i.e.,
\[
\langle y, s_{TDM} \rangle \text{ results in:}
\]
\[
\begin{align*}
    \langle y, s_{TDM} \rangle &= \sum_{n=0}^{M-1} y(n)s_{TDM}(n) \\
    &= \sum_{n=0}^{M-1} y(n) \{ \mathcal{T}_1(n)c_d(n) + \mathcal{T}_2(n)c_u(n) \} \\
    &= \sum_{n=0}^{M-1} r_1(n)\mathcal{F}_1(n) + \sum_{n=0}^{M-1} r_2(n)\mathcal{F}_2(n) \\
    &= R_1(k_1) + R_2(k_2).
\end{align*}
\quad (114)
\]
Since \( k_1, k_2 \in k \), therefore, \( R_1(k_1) \) and \( R_2(k_2) \) can be written as \( R_1(k) \) and \( R_2(k) \). Subsequently, the FS in the up-chirped stream and the down-chirped stream is identified coherently in a disjoint manner as:
\[
\hat{k}_1 = \arg\max_{k \in k} \mathcal{R} \{ R_1(k) \}, \quad (115)
\]
and
\[
\hat{k}_2 = \arg\max_{k \in k} \mathcal{R} \{ R_2(k) \}, \quad (116)
\]
respectively. It may be noticed from Fig. [38] that firstly the received symbol, \( y(n) \) is multiplied with both the down-chirp and the up-chirp, resulting in \( r_1(n) \) and \( r_2(n) \), respectively. Then DFT is evaluated for both \( r_1(n) \) and \( r_2(n) \) that yields \( R_1(k_1) \) and \( R_2(k_2) \), respectively. Subsequently, the FSs in the up-chirped stream and the down-chirped stream are identified using eqs. (115) and (116), respectively.

The non-coherent detector configuration for TDM-CSS is presented in Fig. 39. In the non-coherent detection, both the FSs in the up-chirped stream and the down-chirped stream are disjointly estimated as:
\[
\hat{k}_1 = \arg\max_{k_1 \in k} |R_1(k)|, \quad (117)
\]
and
\[
\hat{k}_2 = \arg\max_{k_2 \in k} |R_2(k)|, \quad (118)
\]
respectively.
In IQ-CSS, the in-phase and quadrature components of the un-chirped symbol are chirped using an up-chirp. In IQTDM-CSS, the up-chirped and down-chirped symbols are simultaneously used for both the in-phase components and the quadrature components. Thus, the SE of IQ-TDM-CSS is twice that of IQ-CSS and four times that of LoRa. Nonetheless, the overall complexity of IQ-TDM-CSS is much higher than LoRa and IQ-CSS.

1) Transmission: The transmitter architecture of IQ-TDM-CSS is illustrated in Fig. [40] It can be observed that four bit sequences, each having λ bits are used. These four bit sequences are used to determine four distinct FS, $k_i, k_q, \tilde{k}_i$ and $\tilde{k}_q$. It is highlighted that $k_i$ and $k_q$ are the FSs to determine the in-phase un-chirped symbols, $f_1(n)$ and $f_3(n)$ after bi2de conversion, whereas, $k_q$ and $\tilde{k}_q$ are used to ascertain the un-chirped quadrature symbols, $f_2(n)$ and $f_4(n)$. These in-phase and the quadrature un-chirped symbols are given as:

$$f_1(n) = \exp \left\{ \frac{2\pi}{M} k_i n \right\},$$  

$$f_2(n) = j \exp \left\{ \frac{2\pi}{M} k_q n \right\},$$  

$$f_3(n) = \exp \left\{ \frac{2\pi}{M} \tilde{k}_i n \right\},$$  

$$f_4(n) = j \exp \left\{ \frac{2\pi}{M} \tilde{k}_q n \right\},$$

respectively. It may be noticed that a phase rotation of $\pi/2$ is included in the quadrature un-chirped symbols. Subsequently, the in-phase un-chirped symbols, $f_1(n)$ and $f_3(n)$ are multiplied with an up-chirp and down-chirp symbols, yielding $s_1(n)$ and $s_3(n)$ that are given as:

$$s_1(n) = j \exp \left\{ \frac{\pi}{M} n (2k_i + n) \right\},$$

$$s_3(n) = \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_i - n) \right\}.$$ 

Similarly, the quadrature un-chirped symbols, $f_2(n)$ and $f_4(n)$ are also multiplied with up-chirp and down-chirp symbols, yielding $s_2(n)$ and $s_4(n)$ that are given as:

$$s_2(n) = j \exp \left\{ \frac{\pi}{M} n (2k_q + n) \right\},$$

and

$$s_4(n) = \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_q - n) \right\}.$$ 

Next, the in-phase up-chirped symbol and in-phase down-chirped symbols are added together to attain a consolidated in-phase chirped symbol, $s_i(n)$ that is given as:

$$s_i(n) = s_1(n) + s_3(n) = \exp \left\{ \frac{\pi}{M} n (2k_i + n) \right\} + \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_i - n) \right\}.$$ 

Similarly, the up-chirped and down-chirped quadrature symbols are also added together to attain quadrature chirped symbol, $s_q(n)$, that is given as:

$$s_q(n) = s_2(n) + s_4(n) = j \left[ \exp \left\{ \frac{\pi}{M} n (2k_q + n) \right\} \right] + \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_q - n) \right\}.$$ 

Lastly, the in-phase and the quadrature symbols, $s_i(n)$ and $s_q(n)$ are added together to attain the IQ-TDM-CSS symbol, $s_{\text{IQ-TDM}}(n)$ that is given as:

$$s_{\text{IQ-TDM}}(n) = s_i(n) + s_q(n) = \exp \left\{ \frac{\pi}{M} n (2k_i + n) \right\} + j \exp \left\{ \frac{\pi}{M} n (2k_q + n) \right\}$$

$$+ \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_i + n) \right\} + j \exp \left\{ \frac{\pi}{M} n (2\tilde{k}_q + n) \right\}.$$ 

The symbol energy of IQ-TDM-CSS symbol is $E_s = 1/M \sum_{n=0}^{M-1} |s_{\text{IQ-TDM}}(n)|^2 = 4$.

2) Detection: The coherent detector architecture for IQ-TDM-CSS is presented in Fig[41]. The inner product of the received symbol vector, $\mathbf{y}$ and the transmit symbol vector, $s_{\text{IQ-TDM}} = [s_{\text{IQ-TDM}}(0), s_{\text{IQ-TDM}}(1), \ldots, s_{\text{IQ-TDM}}(M-1)]^T$ is evaluated as:

$$\langle \mathbf{y}, s_{\text{IQ-TDM}} \rangle = \sum_{n=0}^{M-1} y(n) \overline{s_{\text{IQ-TDM}}}(n)$$

$$= \sum_{n=0}^{M-1} y(n) \{ \overline{f_1(n)c_1(n)} + \overline{f_2(n)c_4(n)} + \overline{f_3(n)c_3(n)} + \overline{f_4(n)c_2(n)} \}$$

$$+ \sum_{n=0}^{M-1} r_1(n) \overline{f_1(n)} + \sum_{n=0}^{M-1} r_2(n) \overline{f_2(n)}$$

$$+ \sum_{n=0}^{M-1} r_3(n) \overline{f_3(n)} + \sum_{n=0}^{M-1} r_4(n) \overline{f_4(n)}$$

$$= R_1(k_i) - jR_1(k_q) + R_2(\tilde{k}_i) - jR_2(\tilde{k}_q).$$

Since $k_i, k_q, \tilde{k}_i, \tilde{k}_q \in k$, where $k = [0, M-1]$, consequently, $R_1(k_i), R_1(k_q), R_2(\tilde{k}_i)$, and $R_2(\tilde{k}_q)$ can be simply
written as $R_1(k), R_1(k), R_2(k),$ and $R_2(k)$. According to the coherent detection criterion (cf. eq. 4) and as shown in Fig. 37, the received symbol, $y(n)$ is despreaded using both the down-chirp and the up-chirp yielding $r_1(n)$ and $r_2(n)$, respectively. Then, the DFT operation is performed on $r_1(n)$ and $r_2(n)$ that results in $R_1(k)$ and $R_2(k)$, respectively. Then, the FS in the in-phase and quadrature components of the up-chirp symbol and down-chirp symbol is evaluated after I/Q separation performed by the I/Q separator as:

$$\hat{k}_i = \arg \max_{k_i \in k} \Re \{ R_1(k) \},$$

$$\hat{k}_q = \arg \max_{k_q \in k} \Im \{ R_1(k) \},$$

$$\hat{k}_i = \arg \max_{k_i \in k} \Re \{ R_2(k) \},$$

and

$$\hat{k}_q = \arg \max_{k_q \in k} \Im \{ R_2(k) \},$$

respectively. Note that eqs. (132) and (134) follows from the fact that $\Re \{-jx\} = \Im \{ x \}$, where $x$ is an arbitrary complex number.

**P. Dual-Mode CSS (DM-CSS)**

DM-CSS has been proposed as an alternative modulation for LPWANs in [55]. DM-CSS is an enhanced variant of the DO-CSS. In DO-CSS, one even and one odd FS of the un-chirped symbol were simultaneously used. However, in DM-CSS, in addition to using the even and odd FSs of the un-chirped symbol, the symbol can be chirped by using either an up-chirp or a down-chirp as in SSK-LoRa. Moreover, each of the even and odd FS also incorporate binary PSs. Therefore, DM-CSS is capable of transmitting 3 more bits relative to DO-CSS.

1) Transmission: The transmitter configuration of DM-CSS is provided in Fig. 42. Consider that $M$ frequencies are available in bandwidth $B$. Among these $M$ frequencies, there are $M/2$ even and $M/2$ odd frequencies. $\lambda_e = \log_2(M/2)$ and $\lambda_o = \log_2(M/2)$ bits are used to determine the even and odd FS that is to be activated, i.e., $k_e$ and $k_o$. Moreover, the activated frequencies (even or odd) have a PS of either 0 or $\pi$ radians. This PS is attained using $\lambda_{PS} = 2$ bits for both even and odd FS. The PS included in the even activated FS is denoted by $\alpha_e$, while the PS for the odd FS is denoted as $\alpha_o$. The PS for even and odd FS is given as:

$$\alpha_e/\alpha_o = \begin{cases} 1 & \lambda_{PS} = 0 \\ -1 & \lambda_{PS} = 1. \end{cases}$$

Using $k_e$ and $k_o$, the un-chirped symbols corresponding to even and odd FSs are given by:

$$f_1(n) = \alpha_e \exp \left\{ \frac{2\pi}{M} k_e n \right\},$$

and

$$f_2(n) = \alpha_o \exp \left\{ \frac{2\pi}{M} k_o n \right\},$$

respectively. The consolidated un-chirped symbol, $f(n)$ is attained by adding $f_1(n)$ and $f_2(n)$ as:

$$f(n) = f_1(n) + f_2(n) = \alpha_e \exp \left\{ \frac{2\pi}{N} k_e n \right\} + \alpha_o \exp \left\{ \frac{2\pi}{N} k_o n \right\}.$$
On the contrary, if \( f(n) \) will be chirped with the down-chirp symbol when \( \gamma_s = -1 \), i.e., \( c_\alpha(n) = c_o(n) \), then, the chirped symbol \( \text{DM-CSS} \) symbol is given as:

\[
s_{\text{DM}}(n) = f(n)c_o(n) = \alpha_e \exp \left\{ \frac{2\pi}{N} n (k_e - n) \right\} + \alpha_o \exp \left\{ \frac{2\pi}{N} n (k_o - n) \right\}.
\] (140)

The symbol energy, \( E_s \), of \( s_{\text{DM}}(n) \) irrespective of whether \( s(n) \) is chirped with an up-chirp or a down-chirp is \( E_s = 1/N \sum_{n=0}^{N-1} |s_{\text{DM}}(n)|^2 = 2 \).

2) Detection: In [55], we only proposed the non-coherent detector for DM-CSS, however, a coherent detection for DM-CSS is also possible that is explained in the sequel.

The coherent detector architecture for DM-CSS is presented in Fig. 43. The inner product of the received symbol vector, \( y \) and the transmit symbol vector, \( s_{\text{DM}} = [s_{\text{DM}}(0), s_{\text{DM}}(1), \cdots, s_{\text{DM}}(M - 1)]^T \) will yield two different results as either an up-chirp symbol or a down-chirp symbol can be used for spreading the un-chirped symbol at the transmitter. The inner product, \( \langle y, s_{\text{DM}} \rangle \) when an up-chirp is used at the transmitter is given as:

\[
\langle y, s_{\text{DM}} \rangle = \sum_{n=0}^{M-1} y(n)\bar{s}_{\text{DM}}(n) = \sum_{n=0}^{M-1} y(n)\bar{r}_1(n)c_\alpha(n) = \sum_{n=0}^{M-1} r_1(n)\bar{f}(n)
\] (141)

On the other hand, when down-chirp is used at the transmitter, then, the inner product, \( \langle y, s_{\text{DM}} \rangle \) yields

\[
\langle y, s_{\text{DM}} \rangle = \sum_{n=0}^{M-1} y(n)\bar{s}_{\text{DM}}(n) = \sum_{n=0}^{M-1} y(n)\bar{r}_2(n)c_o(n) = \sum_{n=0}^{M-1} r_2(n)\bar{f}(n)
\] (142)

Moreover, it may be noticed that \( R_1(k_e) + R_1(k_o) = R_1(k) \) and \( R_2(k_e) + R_2(k_o) = R_2(k) \), where \( k = [0, M - 1] \). The first step is to determine whether an up-chirp or a down-chirp was used at the transmitter for spreading the un-chirped symbol. In order to do so, firstly, \( \kappa_{1\text{coh}} = \max\{\Re\{R_1(k)\}\} \) and \( \kappa_{2\text{coh}} = \max\{|R_2(k)|\} \) are ascertained, which leads to the estimation of the chirp rate as:

\[
\hat{\beta}_s = \begin{cases} 0 & \kappa_{1\text{coh}} > \kappa_{2\text{coh}} \\ 1 & \kappa_{1\text{coh}} < \kappa_{2\text{coh}} \\ \end{cases}
\] (143)

using \( \hat{\beta}_s, \hat{\lambda}_o \) is attained after de2bi conversion. \( \kappa_{1\text{coh}} \) and \( \kappa_{2\text{coh}} \) also help determine whether \( R_1(k) \) or \( R_2(k) \) contain the largest peak and should be used for further processing as:

\[
R(k) = \begin{cases} R_1(k) & \kappa_{1\text{coh}} > \kappa_{2\text{coh}} \\ R_2(k) & \kappa_{1\text{coh}} < \kappa_{2\text{coh}} \end{cases}
\] (144)

Subsequently, the even and the odd FSs and the PSs on the even and odd FSs are determined jointly in a coherent manner after the separation of even and odd frequency indexes using even/odd separator as:

\[
\hat{k}_e, \hat{\alpha}_e = \arg \max_{k_e, \alpha_e} \Re\{\alpha_e R(k_e)\},
\] (145)

and

\[
\hat{k}_o, \hat{\alpha}_o = \arg \max_{k_o, \alpha_o} \Re\{\alpha_o R(k_o)\},
\] (146)

respectively. Using \( \hat{k}_e, \hat{k}_o, \hat{\lambda}_e \) and \( \hat{\lambda}_o \) are ascertained using de2bi conversion.

The non-coherent detector configuration for DM-CSS is provided in Fig. 44. In the non-coherent detector, the chirp rate is identified using \( \kappa_{1\text{non-ch}} = \max\{R_1(k)\} \) and \( \kappa_{2\text{non-ch}} = \max\{|R_2(k)|\} \). Moreover, the activated even and odd FSs are
estimated as:
\[
\hat{k}_c = \arg \max_{k_o} |R(k_o)|, \quad (147)
\]
and
\[
\hat{k}_o = \arg \max_{k_o} |R(k_o)|, \quad (148)
\]
respectively. Subsequently, the PSs of the even and odd frequencies are respectively evaluated by determining the polarity \(R(\hat{k}_c)\) and \(R(\hat{k}_o)\) as:
\[
\hat{\alpha}_c = \text{polarity}\{R(\hat{k}_c)\}, \quad (149)
\]
and
\[
\hat{\alpha}_o = \text{polarity}\{R(\hat{k}_o)\}, \quad (150)
\]
where \(\text{polarity}\{\cdot\}\) is the ML criterion to determine the polarity of the input frequency index. If the output of polarity \(\{\cdot\}\) is 0, then \(\hat{\alpha}_c/\hat{\alpha}_o = 1\), conversely, \(\hat{\alpha}_c/\hat{\alpha}_o = -1\). It is accentuated that if the channel’s phase rotation is \(\geq \pi/2\), then, only coherent detection would be viable because the polarity \(\{\cdot\}\) function will no longer be applicable.

IV. PERFORMANCE ANALYSIS

In this section, we compare the performance of the discussed schemes in terms of achievable SE, required SNR per bit versus SE, BER performance and BER performance considering OP and FO.

A. Spectral Efficiency Analysis

We consider the same \(B\) and \(T_s\) for all the approaches. Also, each modulation is treated separately in the sequel.

- In LoRa, the information bearing element is the \(M\) possible FSs, therefore, the number of bits that can be transmitted per LoRa symbol is \(\lambda = \log_2(M)\). The total number of possible symbols for LoRa are \(M_{\text{LoRa}} = M = 2^\lambda\). As a result, the SE of the LoRa is \(\eta_{\text{LoRa}} = \lambda/M\) bits/s/Hz.

- In ICS-LoRa, \(\lambda\) bits are transmitted via the FS. Additionally, \(\lambda_1 = 1\) bit can be transmitted by considering that either the LoRa symbol or its interleaved version can be transmitted. Thus, in total \(\lambda + 1\) bits can be transmitted per ICS-LoRa symbol implying the total number of possible symbols for ICS-LoRa is \(M_{\text{ICS}} = 2^{\lambda+1} = 2M\). The resulting SE for ICS-LoRa is \(\eta_{\text{ICS}} = (\lambda + 1)/M\) bits/s/Hz.

- In E-LoRa, FS carries \(\lambda\) information bits. Additionally, \(\lambda_{1Q} = 1\) bit determines whether the in-phase or quadrature component of the FS is used. Therefore, the total number of bits transmitted in an E-LoRa symbol is \(\lambda + 1\), and the number of possible E-LoRa symbols are \(M_{E-\text{LoRa}} = 2^{\lambda+1} = 2M\). The SE for E-LoRa is \(\eta_{E-\text{LoRa}} = (\lambda + 1)/M\) bits/s/Hz.

- In PSK-LoRa, the information bearing elements are the FSs and PSs of the un-chirped symbols that transmit \(\lambda\) and \(\lambda_{PS} = \log_2(M_{c})\) bits, respectively, resulting in a total of \(\lambda + \lambda_{PS}\) bits per symbol. The SE and the number of possible symbols for PSK-LoRa is \(\eta_{PSK} = (\lambda + \lambda_{PS})/M\) bits/s/Hz and \(M_{PSK} = 2^{\lambda + \lambda_{PS}}\), respectively. If binary PSs, \(M_{c} = 4\), \(\lambda_{PS} = 2\). It has been demonstrated in [16] that the optimal number of PSs that result in the highest EE is \(M_{c} = 4\). Therefore, the optimal variant of PSK-LoRa, referred to as QPSK-LoRa, transmits \(\lambda + 2\) bits per symbol, having a SE of \(\eta_{PSK} = (\lambda + 2)/\lambda\) bits/s/Hz, and the total number of possible symbols is \(M_{PSK} = 2^{\lambda+2} = 4M\).

- In DO-CSS, the information is transmitted via the even and the odd FSs, i.e., \(k_c\) and \(k_o\). Since there are \(M/2\) even and \(M/2\) odd FSs, then, the number of bits that can be transmitted by either the even or the odd FS is \(\lambda = 1\). This results in 2\(\lambda - 2\) bits are transmitted per DO-CSS symbol and a total of \(M_{DO} = 2^{\lambda-2} = M^2 - 4\) possible symbols. Moreover, the SE of DO-CSS is \(\eta_{DO} = (2\lambda - 2)/M\) bits/s/Hz.

- In SSK-LoRa, the information bearing elements are the FS of the un-chirped symbol, and the slope-shift of the chirp symbol, i.e., whether the symbol is chirped using an up-chirp or a down-chirp. The FS is determined using \(\lambda\) bits, whereas, the slope-shift is determined using \(\lambda_o = 1\) bit. So, a total of \(\lambda + 1\) bits are transmitted per SSK-LoRa symbol. Thus, the SE of SSK-LoRa is \(\eta_{SSK} = (\lambda - 1)/M\) bits/s/Hz, and the number of possible symbols is \(M_{SSK} = 2^{\lambda+1} = 2M\).

- In IQ-CSS, a total of \(2\lambda\) bits are transmitted per IQ-CSS symbol, where the first half are embedded in the in-phase part, and the second half onto the quadrature part. Therefore, the total number of possible IQ-CSS symbols are \(M_{IQ} = 2^{2\lambda} = M^2\) and the SE is \(\eta_{IQ} = 2\lambda/M\) bits/s/Hz.

- In DCRK-LoRa, in addition to \(\lambda\) bits that are transmitted in the FS of the un-chirped symbol, \(\lambda_c = \log_2(M_c)\) bits can also be transmitted in the discrete chirp rate, where \(M_c\) is the total number of chirp rates. So, a total of \(\lambda + \lambda_c\) bits are transmitted per DCRK-CSS symbol yielding a SE of \(\eta_{DCRK} = (\lambda + \lambda_c)/M\) bits/s/Hz. Therefore, the total number of possible symbols in DCRK-LoRa is equal to \(M_{DCRK} = 2^{\lambda + \lambda_c}\).

- In FSCSS-IM, a total of \(\lambda_{IM} = \lfloor \log_2(M) \rfloor\) bits are transmitted per symbol. It may be noticed that the total number of distinct FAPs used for transmission in FSCSS-IM may significantly exceed \(M\). Consequently, the number of bits transmitted in FSCSS-IM is significantly higher than that of LoRa. For example, for \(M = 128\) and \(\varsigma = 2\), the number of bits that can be transmitted per symbol is 12 bits, whereas, for the same occupied bandwidth, LoRa is capable of transmitting only 7 bits. It is also highlighted that the maximum number of bits that can be transmitted per symbol would be maximum when \(\varsigma \approx M/2\), however, in this case the EE of would be considerably poor relative to LoRa because an increase in \(\varsigma\) diminishes the EE. Moreover, FSCSS-IM is extremely flexible in terms of achieving different spectral efficiencies by changing \(\varsigma\). The SE of FSCSS-IM is \(\eta_{IM} = \lambda_{IM}/M\) bits/s/Hz, and the number of possible symbols that can be transmitted is \(M_{FS} = 2^{\lambda_{IM}}\).

- In IQ-CIM, both the in-phase and the quadrature un-chirped symbols employ IM, therefore a total of \(2\lambda_{IM}\) bits
can be transmitted per symbol provided the same number of FSs are used for both the in-phase and the quadrature un-chirped symbols. This implies that the number of bits transmitted per IQ-CIM symbol is twice relative to that of FSCSS-IM symbol of the same duration. The total number of possible IQ-CIM symbols are $M_{IQ}^{FS} = 2^{2\lambda M}$. Moreover, the SE of IQ-CIM is $\eta_{IQ}^{FS} = 2\lambda M/\lambda$ bits/s/Hz implying that $\eta_{IQ} = 2\eta_{IQ}^{FS}$.

- SSK-ICS-LoRa combines ICS-LoRa and SSK-LoRa. In addition to the $\lambda$ bits transmitted in the FS of the un-chirped symbol, $\lambda_a = 1$ bit decides whether an up-chirp or a down-chirp shall be used at the transmitter, and $\lambda_1 = 1$ bit determines if the chirped symbol will be interleaved or not. Thus, a total of $\lambda + \lambda_a + \lambda_1 = \lambda + 2$ bits can be transmitted per SSK-ICS-LoRa symbol. Moreover, the number of possible symbols for SSK-ICS-LoRa is $M_{SSK-ICS} = 2^{\lambda+2} = 4\lambda$. As a result, the SE of SSK-ICS-LoRa is $\eta_{SSK-ICS} = (\lambda + 2)/\lambda$ bits/s/Hz.

- In ePSK-CSS, the FF is ascertained using $\lambda_{FS} = \log_2((M/N_b))$ bits. Moreover, the FF and its $N_b - 1$ harmonics each carry $\lambda_N$ bits in the FS. Thus, the total number of bits transmitted per ePSK-CSS symbol is $\lambda_{FS} + N_b\lambda_{APS}$. The resulting SE and total number of possible symbols for ePSK-CSS is $S_{ePSK} = (\lambda_{FS} + N_b\lambda_{APS})/\lambda$ bits/s/Hz, and $M_{ePSK} = 2^{\lambda_{FS} + N_b\lambda_{APS}}$. It had been demonstrated in [52] that in the optimal variant of ePSK-CSS, there are quaternary PSs and $N_b = 2$. Accordingly, the number of bits transmitted by the optimal variant of ePSK-CSS is $\lambda + 3$ and the total number of possible symbols is $M_{ePSK} = 8\lambda$.

- The number of bits that can be transmitted in GCSS depends on the number of groups, $G$. The higher the number of groups, the higher is the number of transmitted bits. This is because the number of bits transmitted by the $g$th group is $\lambda_g = \log_2((M/G))$. It may be noticed that the number of bits transmitted in each group is the same. Thus, when there are $G$ groups, the total number of bits transmitted per GCSS symbol is $G\log_2((M/G))$. This results in a SE of $\eta_{GCSS} = (G\log_2((M/G)))/G$ bits/s/Hz. Moreover, the number of possible symbols in GCSS having $G$ groups is $M_{GCSS} = 2^{G\log_2((M/G))}$.

- In TDM-CSS, $\lambda$ bits are to be transmitted in the up-chirped symbol and another $\lambda$ bits can be transmitted in the down-chirped symbol. Thus, a total of $2\lambda$ bits can be transmitted per TDM-CSS symbol. A total of $M_{TDM} = 2^{2\lambda} = M^2$ symbols are possible for TDM-CSS, and its SE is $\eta_{TDM} = 2\lambda M/\lambda$ bits/s/Hz, which is the same as that of IQ-CSS and twice as much when compared to LoRa.

- IQ-TDM-CSS consists of two in-phase symbols and two quadrature symbols, each of which transmits $\lambda$ bits. To be more precise, two TDM-CSS schemes are simultaneously transmitted, one on the in-phase and the other on the quadrature components. Thus, a total of $4\lambda$ bits can be transmitted per IQ-TDM-CSS symbol. Accordingly, the SE of IQ-TDM-CSS is $\eta_{IQTDM} = 4\lambda M/\lambda$ bits/s/Hz, which is the double of $\eta_{IQ-CSS}$. Moreover, the number of possible symbols for IQ-TDM-CSS is $M_{IQTDM} = M^4$.

- In DM-CSS symbol, $\lambda_e = \log_2((N/2))$ and $n_o = \log_2((N/2))$ bits are encoded in the activated even and odd FSs, i.e., $k_e$ and $k_o$. In addition, $2\lambda_{PS} = 2\log_2(2) = 2$ bits are encoded in the PSs of activated even and odd FSs, i.e., $\alpha_o$ and $\alpha_o$. Lastly, $\lambda_s = \log_2(2) = 1$ bit is encoded in the slope of the chirp. Therefore, the total number of bits that can be transmitted per DM-CSS symbol is $\lambda_e + \lambda_s + 2\lambda_{PS} + \lambda_e = 2\lambda + 1$. Thus, the SE of DM-CSS is $\eta_{DM} = (2\lambda + 1)/\lambda$ bits/s/Hz. In DM-CSS, a total of $M_{DM} = 2^{2\lambda + 1} = M^2 + 2$ distinct symbols is possible.

The spectral efficiencies of each modulation, their respective increase relative to LoRa and the total number of distinct symbols that each modulation can have are listed in Table [V]. It is clear that IQ-TDM-CSS enjoys the highest SE, whereas the lowest SE is obtained by LoRa. The number of possible symbols are enumerated for the purpose of lookup-table (LUT) implementations. This implies that higher SE requires more LUT memory.

As different schemes possess different spectral efficiencies (cf. Table [V]), therefore, for a fair comparison, the schemes having almost similar spectral efficiencies should be compared against one another. In order to do so, we have divided the schemes into six different groups based on attainable spectral efficiencies. These groups are presented in Table [V]. The first group has a SE range of $[\lambda/M, (\lambda + 1)/M]$ bits/s/Hz and includes LoRa, ICS-LoRa, E-LoRa, SSK-LoRa. The schemes in the second group consists of approaches that can provide SE of $[\lambda + 2]/M$ bits/s/Hz, that are, PSK-LoRa and SSK-LoRa. Note that hereby we consider PSK-LoRa having quaternary PSs that yields a SE of $[\lambda + 2]/M$ bits/s/Hz. The next group comprises of schemes achieve a SE of $[\lambda + 3]/M$ bits/s/Hz. In this group, we consider the optimal variant of ePSK-CSS, having $N_b = 2$ and quaternary PSs, and DCRK-CSS having $M_e = 8$ different chirp rates. DO-CSS and GCSS are included in the fourth group that has a SE of $[2\lambda - 2]/M$ bits/s/Hz. The remaining schemes, that are, IQ-CSS, TDM-CSS and DM-CSS having SE $> 2\lambda$ are included in the fifth group, and the last group comprises of IM-based CSS schemes, that are, FSSC-IM and IQ-CIM. In the sequel, we retain this partitioning of the CSS schemes when different performance metrics shall be evaluated.

B. Spectral Efficiency vs Energy Efficiency Performance

In this section, we evaluate the EE performance of the schemes discussed in the previous sections. To be more precise, for a given SE, we ascertain the EE by evaluating $E_b/N_0 = (E_bT_c)/(\eta_{S})$ required to attain a BER of $10^{-3}$. Moreover, the SE is changed by varying the number of available FSs, $M = 2^\lambda$, where $\lambda = \lfloor[6, 12]\rfloor$. In the sequel, we present the EE performance versus required $E_b/N_0$ for all the groups in an AWGN channel. We also separate the performance of coherent and non-coherent detection mechanisms. Note that the missing modulations, such as non-coherent E-LoRa, are not depicted due to their inability to retrieve the entire transmit information. Other modulations, such as GCSS, simply do not exist in the literature. The performance of CSS schemes in each group are also compared with that of LoRa as it is the most widely adopted CSS scheme for LPWANs.
TABLE IV: SE characteristics of different CSS modulations.

| Modulation       | Spectral Efficiency | SE increase w.r.t LoRa | Number of Possible Symbols |
|------------------|---------------------|------------------------|---------------------------|
| LoRa             | $\frac{\lambda}{M}$ | $-$                     | $2^\lambda = M$           |
| E-LoRa           | $\frac{\lambda+1}{M}$ | $\frac{1}{M}$          | $2^{\lambda+1} = 2M$      |
| ICS-LoRa         | $\frac{\lambda+1}{M}$ | $\frac{1}{M}$          | $2^{\lambda+1} = 2M$      |
| PSK-LoRa         | $\frac{\lambda+\log_2(M_{\phi})}{M}$ | $\frac{\log_2(M_{\phi})}{M}$ | $2^{\lambda+\log_2(M_{\phi})}$ |
| DO-CSS           | $\frac{2\lambda-2}{M}$ | $\frac{\lambda-2}{M}$ | $2^{2\lambda-2} = M^2 - 4$ |
| SSK-LoRa         | $\frac{\lambda+1}{M}$ | $\frac{1}{M}$          | $2^{\lambda+1} = 2M$      |
| IQ-CSS           | $\frac{2\lambda}{M}$ | $\frac{\lambda}{M}$   | $2^{2\lambda} = M^2$      |
| DCRK-CSS         | $\frac{\lambda+\log_2(M_{\phi})}{M}$ | $\frac{\log_2(M_{\phi})}{M}$ | $2^{\lambda+\log_2(M_{\phi})}$ |
| FSCSS-IM         | $\frac{\log_2(M_{\epsilon})}{M}$ | $\frac{\lambda-2}{M}$ (2 active indices) | $2^{\lambda IM}$          |
| IQ-CIM           | $\frac{\log_2(M_{\epsilon})}{M}$ | $\frac{\lambda-2}{M}$ (2 active indices) | $2^{\lambda IM}$          |
| SSK-ICS-LoRa     | $\frac{\lambda+2}{M}$ | $\frac{2}{M}$          | $2^{\lambda+2} = 4M$      |
| ePSK-CSS         | $\frac{\log_2(M_0/\Lambda_{\phi})+N_b}{M} \log_2(M_{\phi})$ | $\frac{2}{M}$ (for $\{N_b, M_{\phi}\} = \{4, 2\}$) | $2^{\Lambda+2} + 2^{\log_2(M/\phi)}$ |
| GCSS             | $\frac{G \log_2(M/\phi)}{M}$ | $\frac{2\Lambda-2}{M}$ (for $G = 2$) | $2^G \log_2(M/\phi)$      |
| TDM-CSS          | $\frac{2\lambda}{M}$ | $\frac{\lambda}{M}$   | $2^{2\lambda} = M^2$      |
| IQ-TDM-CSS       | $\frac{4\lambda}{M}$ | $\frac{3\lambda}{M}$  | $2^{4\lambda} = M^4$      |
| DM-CSS           | $\frac{2\lambda+1}{M}$ | $\frac{\lambda+1}{M}$ | $2^{2\lambda+1} = M^2 + 2$ |

TABLE V: Partitioning of CSS schemes into groups based on achievable spectral efficiencies.

| Group | SE (bits/s/Hz) | CSS schemes               |
|-------|----------------|---------------------------|
| Group 1 | $\left[\frac{\lambda}{M}, \frac{\lambda+1}{M}\right]$ | LoRa, ICS-LoRa, E-LoRa, SSK-LoRa |
| Group 2 | $\frac{\lambda+2}{M}$ | PSK-LoRa, SSK-ICS-LoRa |
| Group 3 | $\frac{\lambda+3}{M}$ | ePSK-CSS, DCRK-CSS |
| Group 4 | $\frac{2\lambda-2}{M}$ | DO-CSS, GCSS |
| Group 5 | $\left[\frac{2\lambda}{M}, \frac{2\lambda+1}{M}\right]$ | IQ-CSS, TDM-CSS, IQ-TD-CSS, DM-CSS |
| Group 6 | IM Approaches | FSCSS-IM, IQ-CIM |

1) Group 1: Fig. 45 compares the performance of LoRa, ICS-LoRa, E-LoRa, and SSK-LoRa using coherent detection. It can be observed that the EE performance versus $E_b/N_0$ for E-LoRa and SSK-LoRa is similar, whereas, the performance of ICS-LoRa is marginally deteriorated. This is because the different interleaved chirp symbols present a relatively higher cross-correlation compared to the chirp symbols in E-LoRa and SSK-LoRa (loss of about 0.2 dB relative to E-LoRa and SSK-LoRa when $\lambda = 12$, i.e., when SE $\simeq 0.0029$ bits/s/Hz and loss of about 0.35 dB relative to E-LoRa when $\lambda = 6$, i.e., when SE $\simeq 0.0375$ bits/s/Hz). However, all the schemes perform better than the coherently detected LoRa implying the required energy per bit for correct detection is reduced for the schemes in group 1.

On the other hand, it can be observed from Fig. 46 that depicts the EE performance versus $E_b/N_0$ using a non-coherent detection mechanism, that the performance of SSK-LoRa is marginally better than ICS-LoRa. This is again due to lower cross-correlation between the chirped symbols for SSK-LoRa that increases the minimum Euclidean distance between the distinct chirp symbols. Nonetheless, the performance of both ICS-LoRa and SSK-LoRa is better than that of non-coherently detected LoRa. Note that coherent E-LoRa is not depicted since the information will be lost in the PS of the
un-chirped symbol.

2) Group 2: Fig. 47 compares the performance of PSK-LoRa with quaternary PSs, i.e., QPSK-LoRa and SSK-ICS-LoRa using coherent detection with the coherent detection performance of LoRa. Firstly, it can be observed that the performance of the latter approach is relatively deteriorated compared to QPSK-LoRa and SSK-ICS-LoRa. Secondly, QPSK-LoRa performs better than SSK-ICS-LoRa. This is because the PS does not impact the orthogonality of symbols, whereas, the interleaved versions of either up-chirped or the down-chirped symbols causes a high cross-correlation that impacts the orthogonality.

Fig. 48 depicts the performance of coherent LoRa and SSK-ICS-LoRa. The non-coherent detection for QPSK-LoRa is not possible because of the PS of the un-chirped symbol. Therefore, we depict the performance of semi-coherent detection for QPSK-LoRa, in which the FS is non-coherently detected but the PS is coherently detected using ML criterion. This is necessary because in the presence of channel phase rotation, it would be impossible to determine the PS non-coherently. The performance of SSK-ICS-LoRa is better than non-coherent LoRa, but is worse than the semi-coherent QPSK-LoRa. However, the comparison between non-coherent SSK-ICS-LoRa and semi-coherent QPSK-LoRa is unfair because of the significant difference in terms of computational complexity involved in the detection process.

3) Group 3: It is highlighted that ePSK-CSS(4, 2) transmits \(\lambda + 3\) bits per symbol, whereas, DCRK-CSS is a very flexible approach as it is capable of transmitting more than \(\lambda + 3\) bits per symbol by increasing the number of discrete chirp rates, \(M_c\), at the cost of increased transceiver complexity.

Fig. 49 illustrates the coherent detection performance of DCRK-CSS and ePSK-CSS(2, 4). Here, we consider \(M_c = 8\) that yields a SE of \((\lambda + 3)/M\) bits/s/Hz. For coherent detection, the performance of ePSK-CSS(2, 4) is marginally better than DCRK-CSS for \(\lambda = [6, 8]\), whereas, for \(\lambda = [10, 12]\), the performance of DCRK-CSS is better than that of ePSK-CSS(2, 4). Both DCRK-CSS and ePSK-CSS(2, 4) perform better than coherent LoRa.

Fig. 50 shows the performance of non-coherent LoRa and DCRK-CSS, and semi-coherent ePSK-CSS(2, 4). The performance of non-coherent DCRK-CSS is better than non-coherent LoRa, whereas, the performance of semi-coherent
ePSK-CSS(2, 4) is better than both DCRK-CSS and LoRa. It is accentuated that semi-coherent detection of ePSK-CSS(2, 4) requires channel equalization, that may reduce the practicality of ePSK-CSS(2, 4).

4) Group 4: Group 4 comprises of DO-CSS and GCSS. We consider GCSS with G = 2 because its the most energy efficient variant. Note that both DO-CSS and GCSS with G = 2 achieve a SE of \(\frac{(20 - 2)}{M}\) bits/s/Hz.

Fig. 51 illustrates the EE performance versus required \(E_b/N_0\) using coherent detection. Since the coherent detector for GCSS does not exist, therefore, only the performance of DO-CSS is shown and compared with coherently detected LoRa. It can be noticed that, while the DO-CSS has a higher overall SE, its performance is essentially identical to LoRa for a fixed SE.

Fig. 52 shows the performance of non-coherently detected DO-CSS, GCSS and LoRa. It can be observed that all the schemes exhibit similar performances.

5) Group 5: Fig. 53 illustrates the performance of coherent detection of IQ-CSS, TDM-CSS, IQ-TDM-CSS, DM-CSS and LoRa. Firstly, it can be observed that in addition to be able to achieve higher spectral efficiencies, the performance of coherently IQ-CSS, TDM-CSS, IQ-TDM-CSS and DM-CSS is better than coherently detected LoRa. The performance of DM-CSS is the best among all the remaining approaches in the group. IQ-TDM-CSS performs better than IQ-CSS, TDM-CSS and LoRa for \(\lambda > 8\). However, the performance of IQ-TDM-CSS for \(\lambda = \{6, 7\}\) is very poor relative to other approaches, therefore, the advantage of IQ-TDM-CSS that it can have a higher achievable spectral efficiencies is less beneficial. On the other hand, even though TDM-CSS can attain higher achievable SE, the EE performance is worse than LoRa. IQ-CSS performs better than LoRa, and TDM-CSS but is worst than IQ-TDM-CSS for \(\lambda > 8\) and DM-CSS.

Fig. 54 shows the performance of the non-coherent detection for IQ-CSS with \(\beta = 2.4\), TDM-CSS and DM-CSS. Note the non-coherent detector for IQ-TDM-CSS does not exists. It is clear that the performance of DM-CSS outperforms the other techniques. Roughly speaking, there is a gain (at \(\lambda = 6\)) of 1.2 dB relative to non-coherent IQ-CSS for \(\beta = 2.4\), and 2.1 dB relative to non-coherent TDM-CSS. Also, there is a gain of about 0.8 dB compared to other approaches at \(\lambda = 12\). For \(\lambda < 7\), TDM-CSS performs even worse than LoRa.

Fig. 55: SE versus required SNR per bit for group 6 schemes considering coherent detection and target BER of \(10^{-3}\) in AWGN channel.
6) Group 6: Fig. 55 shows the EE performance of coherent detection of IM approaches, namely FSCSS-IM and IQ-CIM. For both approaches, we consider 2 active indexes, i.e., \( \varsigma = \varsigma_i = \varsigma_q = 2 \). Even though these schemes provide substantial flexibility in terms of achievable spectral efficiencies (if higher number of indexes are activated), their EE performance for coherent detection is even worst than coherently detected LoRa. Moreover, IQ-CIM can achieve high SE relative to FSCSS-IM and is also more energy efficient relative to the latter approach.

Fig. 56 presents the performance of non-coherently detected FSCSS-IM and LoRa. It is highlighted that non-coherent detector for IQ-CIM does not exist as it uses active indexes for both the in-phase and the quadrature components of the unchirped symbols. It is shown that FSCSS-IM with two active indexes can attain higher SE but performs worst in terms of EE relative to non-coherently detected LoRa.

5.1.4. Bit-Error Rate Performance

In this section, we present the BER performance of the discussed schemes considering \( \lambda = 8 \) and AWGN channel. Moreover, the BER performance of (coherent and non-coherent) LoRa is used as benchmark. When applicable, we employ both coherent and non-coherent detection mechanisms. We again divide the scheme into different groups as given in Table IV and present the performance of each group separately in subsequent subsections.

1) Group 1: Group 1 includes, LoRa, ICS-LoRa, E-LoRa, and SSK-LoRa. The BER performance of the schemes is shown in Fig. 57. We can observe that the performance of coherent SSK-LoRa, and E-LoRa is better than coherent LoRa and ICS-LoRa. ICS-LoRa with coherent detection performs marginally better than coherently detected LoRa. Using non-coherent detection, both SSK-LoRa and ICS-LoRa perform better than non-coherently detected LoRa. Note that non-coherent detection of E-LoRa is not possible. It can be observed that the ICS-LoRa, E-LoRa, and SSK-LoRa perform better than LoRa because of reduction in required SNR/bit for correct bit detection.

2) Group 2: The BER performances of schemes in group 2 are provided in Fig. 58. In this section, we consider PSK-LoRa with quaternary PSs, i.e., QPSK-LoRa. It can be observed that coherently detection QPSK-LoRa outperforms coherently detected SSK-ICS-LoRa, and LoRa. Nevertheless, coherently and non-coherently detected SSK-ICS-LoRa outperforms coherently and non-coherently detected LoRa. Note that non-coherent detection for QPSK-LoRa is not feasible, however, it can perform semi-coherent detection. The performance of semi-coherent QPSK-LoRa is the same as coherent LoRa. It is accentuated that the semi-coherent detection of QPSK-LoRa requires synchronization and equalization at the receiver.

3) Group 3: Fig. 59 compares the BER performances of CSS schemes in group 3, which include DCRK-CSS and ePSK-CSS(2, 4) with that of LoRa. It can be observed from Fig. 59 that coherent detection performance of DCRK-CSS, and ePSK-CSS(2, 4) is similar, and both are better than coherently detected LoRa. Note that non-coherent detection is not possible for ePSK-CSS(2, 4), therefore, we use a semi-coherent detection mechanism. The non-coherent detection of DCRK-CSS is better than the non-coherently detected LoRa. On the other hand, semi-coherently detected ePSK-CSS(2, 4) performs slightly worse than coherently detected ePSK-CSS(2, 4). It is recalled that the complexity of DCRK-CSS receiver, coherent or non-coherent, is substantially higher.
than that of ePSK-CSS(2, 4).

4) Group 4: Group 4 includes DO-CSS and GCSS schemes. Coherent detection mechanism for GCSS has not been proposed anywhere in the literature, therefore, we will only illustrate the performance of its non-coherent receiver. It can be observed from Fig. 60 that coherently detected DO-CSS exhibits marginal degradation in BER performance when compared to coherently detected LoRa. It is recalled that the achievable SE for DO-CSS is $\lambda - 2/\lambda$ bits/Hz higher relative to LoRa. Additionally, both DO-CSS and GCSS exhibit similar performances in terms of non-coherent detection, which is somewhat worse than non-coherently detected LoRa.

5) Group 5: IQ-CSS, TDM-CSS, IQ-TDM-CSS, and DM-CSS are included in group 5. Fig. 61 compares the BER performances of coherent and non-coherent detection of these schemes. It is recalled that IQ-CSS, TDM-CSS, IQ-TDM-CSS, and DM-CSS offers a SE increase of $\lambda/\lambda M$, $3\lambda/\lambda M$, and $(\lambda + 1)/M$, respectively over LoRa. The BER performance of coherently detected IQ-CSS is better than that of coherently detected TDM-CSS, IQ-CSS and LoRa. On the other hand, the BER performance of non-coherently detected IQ-CSS with $\beta = 2.4$ is worse than that of non-coherently detected LoRa, TDM-CSS, and DM-CSS. The BER performance of coherently detected TDM-CSS is only better than coherently detected IQ-TDM-CSS, however, IQ-TDM-CSS offers $3\lambda/\lambda M$ bits/Hz higher SE than TDM-LoRa. The BER performance of coherently detected IQ-TDM-CSS is worse than the BER performance of all other schemes using coherent detection, however, its achievable SE is the highest. Lastly, it can be observed that the BER performance of DM-CSS both using coherent detection and non-coherent detection outperforms all the other schemes in the group.

6) Group 6: In group 6, both IM schemes are considered. For FSCSS-IM, $\lambda = 2$, and for IQ-CIM, $\lambda = \psi_i = 2$. The BER performance of both FSCSS-IM and IQ-CIM is depicted in Fig. 62. It can be observed that the BER performance of FSCSS-IM using both coherent and non-coherent detection is worse than LoRa, however, its SE is higher. Moreover, the BER performance of coherently detected IQ-CIM is almost similar to the BER performance of coherently detected FSCSS-IM. It may be observed that the achievable SE of IQ-CIM is double that of FSCSS-IM, and is $2\lambda - 4/\lambda M$ bits/Hz higher than LoRa, when $\lambda = \psi_i = \psi_q = 2$.

D. Bit-Error 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:

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

where $\psi$ is the PO. We evaluate the performance of each scheme considering $\psi = \pi/8$ radians, $\psi = \pi/4$ radians, and $\lambda = 8$. Moreover, we consider the same categorization of the schemes as listed in Table V.

1) Group 1: The BER performances of schemes in group 1 considering $\psi = \pi/8$ radians is found in Fig. 63. It can be observed that the BER performance using coherent detection for LoRa is negatively impacted in the presence of PO, whereas, the BER of non-coherently detected LoRa remains...
unchanged. In general, the BER performance using coherent detection is more impacted relative to when non-coherent detection is used. To be more precise, for $\psi = \pi/8$ radians, the BER performances of ICS-LoRa and SSK-LoRa using both coherent detection and non-coherent detection is almost the same as coherently detected LoRa. The BER performance of E-LoRa is mostly affected in the presence of PO, and is similar to that of non-coherently detected LoRa.

Fig. 64 shows the performance of schemes in group 1, when $\psi$ is increased from $\psi = \pi/8$ radians to $\psi = \pi/4$ radians. It can be observed from Fig. 64 that BER performance using coherent detection for all the schemes is severely impacted, i.e., more than 2 dB degradation in $E_b/N_0$. Because the information is sent in the phase of the unchirped symbols, coherently detected E-LoRa has very low BER performance. The BER performance for LoRa, ICS-LoRa, and SSK-LoRa with non-coherent detection remains agnostic, demonstrating the superiority of non-coherent detection over coherent detection.

2) Group 2: Fig. 65 shows the BER performances of QPSK-LoRa and SSK-ICS-LoRa in the presence of PO at $\psi = \pi/8$ radians. The BER performance of QPSK-LoRa using both the coherent detection and semi-coherent detection deteriorates by about 0.5 dB for $\psi = \pi/8$ radians relative to when $\psi = 0$ radians. This is because QPSK-LoRa transmits information in the PS, which may be negatively influenced by the PO. For SSK-ICS-LoRa, the BER performance using non-coherent detection remains unaffected, however, the BER performance using coherent detection suffers a degradation of about 1 dB when $\psi = \pi/8$ radians compared to when $\psi = 0$ radians. The BER performance considering non-coherent detection for SSK-ICS-LoRa is marginally better relative to when coherent detection was used for low $E_b/N_0$ values. The BER performance of coherently detected LoRa and coherently detected SSK-ICS-LoRa is the same implying that SSK-ICS-LoRa suffers a higher degradation in the presence of PO because of high cross-correlation between different symbols.

Fig. 66 shows the BER performance of group 2 schemes when PO is increased from $\psi = \pi/8$ radians to $\psi = \pi/4$ radians. It can be observed from the BER performance that when $\psi = \pi/4$ radians, both coherent and semi-coherent detection
mechanisms results an error floor of about $10^{-1}$ because the high PO impacts the PSs that have been incorporated in FS. The BER performance of coherently and non-coherently detected SSK-ICS-LoRa is similar to coherently and non-coherently detected LoRa, respectively. There is a degradation in BER performance in coherent detection, when compared to non-coherent detection for converging schemes.

3) Group 3: Fig. 67 illustrates the BER performance of DCRK-CSS and ePSK-CSS(2, 4) in the presence of PO of $\psi = \pi/8$ radians. It can be observed that the BER performance of coherently detected DCRK-CSS degrades by 0.5 dB when $\psi = \pi/8$ radians compared to when $\psi = 0$ radians. On the other hand, the BER performance of non-coherently detected DCRK-CSS remains unaffected when PO is increased from $\psi = 0$ radians to $\psi = \pi/8$ radians. Moreover, the BER performance of both coherently detected and non-coherently detected DCRK-CSS is the same when $\psi = \pi/8$ radians. On the other hand, the BER performance of ePSK-CSS(2, 4) using coherent and semi-coherent detection degrades by approximately 1 dB when PO of $\psi = \pi/8$ radians is introduced. The BER performance of DCRK-CSS using both coherent detection and non-coherent detection is superior to LoRa, and ePSK-CSS(2, 4) using either coherent or semi-coherent/non-coherent detection.

The BER performance of group 3 schemes for $\psi = \pi/4$ radians is shown in Fig. 68. It can be observed that the BER performance of non-coherently detected DCRK-CSS remains unchanged even when PO is increased from $\psi = \pi/8$ radians to $\psi = \pi/4$ radians. However, the BER performance of DCRK-CSS using coherent detection suffers a BER degradation of 2.5 dB when PO increases from $\psi = 0$ radians to $\psi = \pi/4$ radians. Unlike the BER performance of semi-coherently detected PSK-LoRa, semi-coherently detected ePSK-CSS(2, 4) performs relatively better but its performance degrades significantly. The same is true for coherently detected ePSK-CSS(2, 4), which experiences a 2.5 dB BER degradation as PO rises from $\psi = 0$ radians to $\psi = \pi/4$ radians. Coherently detected ePSK-CSS(2, 4) performs slightly better than DCRK-CSS for high $E_b/N_0$ values.

4) Group 4: The BER performance of coherently and non-coherently detected DO-CSS and non-coherently detected GCSS when $\psi = \pi/8$ radians is illustrated in Fig. 69. It can be observed that coherently detected DO-CSS suffers a BER penalty of approximately 0.5 dB when the PO increases from $\psi = 0$ radians to $\psi = \pi/8$ radians. The BER performance of non-coherently detected DO-CSS, however, does not exhibit any PO degradation. Moreover, the BER performance of coherently detected and non-coherently detected DO-CSS are almost identical. For non-coherently detected GCSS, the BER performance remains largely unaffected in the presence of PO. It may also be noticed from Fig. 69 that the BER performances of both coherently and non-coherently detected LoRa remains better than that of coherently and non-coherently detected DO-CSS, and non-coherently detected GCSS.

Fig. 70 shows a considerable degradation of about 3 dB when PO increases from $\psi = 0$ radians to $\psi = \pi/4$ radians for coherently detected DO-CSS. On the other hand, the BER performances of non-coherently detected DO-CSS and GCSS remains unchanged even when PO increases from $\psi = 0$
radians to $\psi = \pi/4$ radians.

5) Group 5: The BER performances in the presence of PO of $\psi = \pi/8$ radians for schemes in group 5 is shown in Fig. 71. It can be observed that the performances of coherently detected IQ-CSS, and TDM-CSS for $\psi = \pi/8$ radians degrades by about 1.5 dB and 0.5 dB, respectively relative to when there is no PO. On the contrary, the BER performance of non-coherently detected IQ-CSS and TDM-CSS remains agnostic to PO. The BER performance of coherently detected IQ-TDM-CSS degrades severely as can be seen from Fig. 72. The BER performance of coherently detected IQ-TDM-CSS suffers a penalty of about 1 dB and 1.5 dB when PO increases from $\psi = 0$ radians to $\psi = \pi/8$ radians. Among coherently detected approaches, the BER performance of DM-CSS is the best, whereas, among non-coherently detected approaches, apart from LoRa, the performance of TDM-CSS is the best.

The BER performances of group 5 schemes when PO is $\psi = \pi/4$ radians is shown in Fig. 72. It can be observed from Fig. 72 that the BER of coherently and non-coherently detected IQ-CSS, and coherently detected IQ-TDM-CSS achieve an error floor at BER of $10^{-1}$ for $\psi = \pi/4$, implying a severe degradation in the BER performance. For coherently detected TDM-CSS, the BER performance degrades by approximately 3.2 dB when $\psi = \pi/4$ radians compared to when there is no PO, whereas, in the same scenario, the BER performance of non-coherently detected TDM-CSS remains unchanged. Moreover, the BER performance of coherently and non-coherently detected DM-CSS suffers a penalty of 2.8 dB when PO increases from $\psi = 0$ radians to $\psi = \pi/4$ radians. This is due to encoding binary PSs in the signal structure. Overall, the BER performance of non-coherently detected TDM-CSS remains the best.

6) Group 6: The BER performances of IM approaches for $\psi = \pi/8$ radians are illustrated in Fig. 73. For FSCSS-IM and IQ-CIM, we use $\varsigma = c_2 = s_q = 2$. It can be observed that coherently detected FSCSS-IM suffers a loss of approximately
1 dB in BER when $\psi = \pi / 8$ radians relative to when there is no PO. The BER performance of non-coherently detected FSCSS-IM remains constant even when PO is increased from $\psi = 0$ radians to $\psi = \pi / 8$ radians. On the other hand, the BER performance of coherently detected IQ-CIM degrades severely when PO is increased.

From Fig. 74 it can be observed that coherently detected FSCSS-IM suffers by almost 3 dB when PO is increased from $\psi = 0$ radians to $\psi = \pi / 4$ radians. IQ-CIM achieve a very high error floor in the presence of a PO of $\psi = \pi / 4$ radians. Overall, FSCSS-IM performs relatively better than IQ-CIM.

**E. Bit-Error Performance considering Frequency Offset**

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

$$y(n) = \exp \left( j 2 \pi f n \frac{1}{M} \right) s(n) + w(n), \quad (152)$$

where $\Delta f$ is the FO in Hertz (Hz). To evaluate the BER performance, we consider FOs of $\Delta f = 0.1$ Hz, $\Delta f = 0.2$ Hz, $\lambda = 8$ and AWGN channel for all the CSS schemes.

1) Group 1: The BER performances of group 1 schemes for $\Delta f = 0.1$ Hz are presented in Fig. 75. It can be observed that all schemes in group 1 do not suffer any deterioration in the BER performance when non-coherently detected in the presence of $\Delta f = 0.1$ Hz. On the contrary, when non-coherent detection is employed, then, the BER performances of all the schemes show a loss of around 0.6 dB when $\Delta f = 0.1$ Hz relative when $\Delta f = 0$ Hz. The BER performances of SSK-LoRa and ICS-LoRa are the best in both cases when either coherent detection or non-coherent detection is applied.

Fig. 76 shows the BER performances of the schemes in group 1 when FO is set at $\Delta f = 0.2$ Hz. We observe that there is a BER degradation of approximately 3 dB when coherent detection is applied for LoRa, ICS-LoRa and SSK-LoRa when $\Delta f = 0.2$ Hz. Nonetheless, the BER performances of these schemes when non-coherent detection is applied, do not show any degradation for the same FO at $\Delta f = 0.2$ Hz. On the other hand, the performance of E-LoRa is severely affected as can be seen from Fig. 76. The BER performances of coherently and non-coherently detected SSK-LoRa is the best among other schemes using coherent and non-coherent detection, respectively.

2) Group 2: Fig. 77 compares the BER performances of group 2 schemes, that are, QPSK-LoRa and SSK-ICS-LoRa considering FO at $\Delta f = 0.1$ Hz. It can be observed from the BER performances that when coherent detection is applied for the above-mentioned schemes, there is a degradation of almost 0.6 dB when $\Delta f = 0.1$ Hz. On the other hand, the BER performance attained using semi-coherent detection for QPSK-
LoRa is not affected by the FO at $\Delta f = 0.1$ Hz, whereas, the BER performance of non-coherently detected SSK-ICS-LoRa culminates a degradation of 0.2 dB for the same FO. Overall, the performance of semi-coherently detected QPSP-LoRa is the best.

Fig. 78 shows the BER performances of QPSK-LoRa and SSK-ICS-LoRa when $\Delta f = 0.2$ Hz. It can be observed that even though the performance of coherently and semi-coherently detected QPSK-LoRa was better than SSK-ICS-LoRa when $\Delta f = 0.1$ Hz, an increase in FO by 0.1 Hz, i.e., $\Delta f = 0.2$ Hz, the BER performances of coherently and non-coherently detected QPSK-LoRa are severely affected. When FO grows from $\Delta f = 0$ Hz to $\Delta f = 0.2$ Hz, SSK-ICS-LoRa coherently detected performance suffers by roughly 2.5 dB. Coherent and non-coherent detection remain the best.

3) Group 3: The BER performances of group 3 schemes are shown in Fig. 79 when $\Delta f = 0.1$ Hz. We observe that both coherently detected DCRK-CSS and ePSK-CSS(2, 4) deteriorate by approximately 0.3 dB and 1 dB. In comparison to coherent and semi-coherently detected DCRK-CSS and ePSK-CSS(2, 4), coherent and non-coherently detected DCRK-CSS performs better in terms of BER.

The BER performances of DCRK-CSS and ePSK-CSS(2, 4) considering FO of $\Delta f = 0.2$ Hz are illustrated in Fig. 80 where a degradation of 2.5 dB and 2.7 dB for coherently detected DCRK-CSS and ePSK-CSS(2, 4), respectively, is observed. In terms of BER performance attained using non-coherent detection, we observe that DCRK-CSS suffers a loss of 0.6 dB, whereas, semi-coherently detected ePSK-CSS(2, 4) manifests a loss of around 2 dB when FO increases from $\Delta f = 0$ Hz to $\Delta f = 0.2$ Hz. The performance of non-coherently detected DCRK-CSS is the best.

4) Group 4: Fig. 81 depicts the BER performances of group 5 schemes when $\Delta f = 0.1$ Hz. It can be seen that when coherent detection is applied for DO-CSS, its BER performance suffers a loss of approximately 0.5 dB when $\Delta f = 0.1$ Hz compared to FO-free case. On the contrary,
BER performances using non-coherent detection for DO-CSS and GCSS shows a loss of 0.1 dB when FO increases from $\Delta f = 0$ Hz to $\Delta f = 0.1$ Hz.

Fig. 82 shows the BER performance of DO-CSS and GCSS when FO of $\Delta f = 0.2$ Hz is considered. When FO is increased from $\Delta f = 0$ Hz to $\Delta f = 0.2$ Hz, then, the BER performance of non-coherently detected DO-CSS and GCSS suffers a loss of 0.7 dB and 0.6 dB, respectively. Moreover, in the same scenario, the BER performance of coherently detected DO-CSS degrades by 2.7 dB. Among DO-CSS and GCSS, the performance of GCSS is better if not the same than DO-CSS.

5) Group 5: The BER performances of the CSS schemes in group 5 considering FO of $\Delta f = 0.1$ Hz are illustrated in Fig. 83. When FO increases from $\Delta f = 0$ Hz to $\Delta f = 0.1$ Hz, we can observe a degradation of 1 dB and 0.1 dB in the BER performances of coherently and non-coherently detected IQ-CSS, respectively. Considering the same change in FO, the BER deteriorates by 0.4 dB in both cases when coherent detection or non-coherent detection is applied to TDM-CSS. Coherently detected IQ-TDM-CSS suffers severe losses in the similar scenario. Lastly, DM-CSS suffers BER loss of 0.7 dB for both coherent and non-coherent detection. It can be observed that the BER performance of coherently and non-coherently detected DM-CSS remains the best relative to when coherent and non-coherent detection is used for the remainder of the schemes.

Fig. 84 shows the BER performance of group 5 schemes when FO is increased from $\Delta f = 0$ Hz to $\Delta f = 0.2$ Hz. Firstly, we observe that coherently detected IQ-TDM-CSS and IQ-TDM-CSS achieve an error floor around BER of $10^{-1}$. The BER performance of non-coherently detected IQTDM-CSS, IQ-TDM-CSS, and DM-CSS suffers a loss of 0.6 dB, 0.7 dB and 2 dB, respectively. Whereas, under similar conditions, the BER performances of coherently detected IQTDM-CSS and DM-CSS deteriorates by 2.7 and 3.4 dB, respectively. Overall, the performance of non-coherently detected IQ-TDM-CSS is best when FO of $\Delta f = 0.2$ Hz is considered.

6) Group 6: The BER performances of IM schemes, that are, FSCSS-IM and IQ-CIM are illustrated in Fig. 85 consid-
In this section, we provide the merits and limitations of each CSS approaches, the main features we consider are:

1) constant envelope property;
2) EE improvement with respect to LoRa;
3) SE improvement with respect to LoRa;
4) possibility of coherent detection;
5) possibility of non-coherent detection;
6) robustness against PO;
7) robustness against FO;

If any waveform exhibits a constant envelope, then, the peak-to-average power ratio (PAPR) of that approach would be very low, resulting in ease of system design and implementation. On the contrary, high PAPR could negatively impact the EE. This is due to the Input Back Off at which the power amplifier operates in. A high PAPR waveform will be more susceptible to clipping and hence increasing $E_b/N_0$ will not improve the BER. The EE improvement with respect to LoRa shows a direct improvement in terms of power consumption compared to LoRa. If a scheme is more energy efficient than LoRa, then the battery terminals used in LPWAN nodes would consume less power, resulting in a longer battery lifetime at the same BER accuracy. If a scheme is more spectral efficient than LoRa, then reducing the latency between the transmission of information and its reception. When coherent detection is used, the overall system complexity significantly grows, but non-coherent detection keeps the overall system complexity in check, making it more feasible. Moreover, coherent detection will also increase the cost of the system as additional processing is required for equalization, synchronization, etc. In low-cost LPWAN devices, it is expected that there would be some PO due to the phase-noise, and residual FO due to imperfect compensation of carrier frequency offsets. The schemes that are more robust to these offsets would offer a lower power consumption, allow low-cost implementation, and an overall better performance.

It is expected that in the presence of PO/FO, non-coherent detection will outperform coherent detection. This could be explained from the fact that PO/FO impairments could be seen

V. MERITS AND LIMITATIONS OF CSS SCHEMES

In this section, we provide the merits and limitations of each approach. Though there can be significant number of features

that can be used to ascertain the merits and limitations of the CSS approaches, the main features we consider are:

1) constant envelope property;
2) EE improvement with respect to LoRa;
3) SE improvement with respect to LoRa;
4) possibility of coherent detection;
5) possibility of non-coherent detection;
6) robustness against PO;
7) robustness against FO;

If any waveform exhibits a constant envelope, then, the peak-to-average power ratio (PAPR) of that approach would be very low, resulting in ease of system design and implementation. On the contrary, high PAPR could negatively impact the EE. This is due to the Input Back Off at which the power amplifier operates in. A high PAPR waveform will be more susceptible to clipping and hence increasing $E_b/N_0$ will not improve the BER. The EE improvement with respect to LoRa shows a direct improvement in terms of power consumption compared to LoRa. If a scheme is more energy efficient than LoRa, then the battery terminals used in LPWAN nodes would consume less power, resulting in a longer battery lifetime at the same BER accuracy. If a scheme is more spectral efficient than LoRa, then reducing the latency between the transmission of information and its reception. When coherent detection is used, the overall system complexity significantly grows, but non-coherent detection keeps the overall system complexity in check, making it more feasible. Moreover, coherent detection will also increase the cost of the system as additional processing is required for equalization, synchronization, etc. In low-cost LPWAN devices, it is expected that there would be some PO due to the phase-noise, and residual FO due to imperfect compensation of carrier frequency offsets. The schemes that are more robust to these offsets would offer a lower power consumption, allow low-cost implementation, and an overall better performance.

It is expected that in the presence of PO/FO, non-coherent detection will outperform coherent detection. This could be explained from the fact that PO/FO impairments could be seen

...
as part of the channel and non-coherent detection does not need to know the channel.

In Table VI we have summarized the performances of the studied schemes for the performance metrics listed above. It may be noticed that the constant envelope property is directly related to whether the scheme is a SC or MC. In general, SC schemes have a constant envelope, whereas, MC or IM schemes do not possess a constant envelope because multiple chirps are simultaneously multiplexed. On the other hand, SE improvement is higher for the approaches which either MC or IM. Thus, there is essentially a trade-off between ease of system design and/or possible improvement in SE. Moreover, generally, only coherent detection is possible for the schemes which carry information in the PS, e.g., E-LoRa, PSK-LoRa, etc. Moreover, the scheme which do not transmit information via the PS are in general more robust to the PO and FO.

In Table VI we have used √ and x to illustrate if a certain function is possible for a given approach or not. On the other hand, different symbol to represent different levels of performance, e.g., + means a slight improvement in performance, or a minute robustness against a given offset, ++ means a considerable improvement in performance, or high robustness against an offset, whereas, +++ is used for high performance improvement.

Hereby, we also give a brief idea about which scheme apart from LoRa is among the best in each group listed in Table VI.

- In group 1, SSK-LoRa is the most promising approach because it possesses the constant envelop property, provides higher SE and improved EE when compared to LoRa. In addition, SSK-LoRa enjoys both coherent and non-coherent detection, thus can attain absolute resilience against PO and FO. It may be noticed that ICS-LoRa also possesses the same properties, however, SSK-LoRa is better than ICS-LoRa because it is capable of producing symbols with lower cross-correlation relative to ICS-LoRa.

- In group 2, SSK-ICS-LoRa is the best scheme as it provides good performance against the different performance metrics. On the other hand, it can be observed that PSK-LoRa is not capable of non-coherent detection as well as it is not robust against the PO and the FO.

- DCRK-CSS performs quite well in all performance metrics; nevertheless, the complexity of the DCRK-CSS receiver is very high, which may be a limiting factor. On the other hand, ePSK-CSS cannot have a non-coherent receiver, and has a lower robustness against the PO and the FO compared to DCRK-CSS, however, its complexity is relatively less than DCRK-CSS. Overall, DCRK-CSS may be the best option because of its flexibility in achieving different spectral efficiencies by employing different chirp rates.

- In group 4, DO-CSS and GCSS with G = 2 perform equally well. However, for GCSS, the SE can be further improved when a higher G is used. However, by doing so, the EE, and robustness against the offsets could be affected. Therefore, GCSS can provide a better trade-off between SE and other parameters.

- In group 5, the performance of DM-CSS is optimal in an ideal non-linear channel; however, in the presence of PO and FO, TDM-CSS outperforms DM-CSS. It may be noticed that TDM-CSS transmits one bit less per symbol relative to DM-CSS. Moreover, the non-coherent detector for DM-CSS only works if the channel induced phase rotation is less than π/2. The robustness of IQ-CSS and IQ-TDM-CSS is weak in the presence of PO and FO. Nonetheless, IQ-TDM-CSS is capable of providing extremely high bit rate compared to other counterparts in the group. To conclude, the performances of TDM-CSS makes it the best approach in group 5.

- Among the IM approaches, while IQ-CIM can achieve greater SE, its constraints, such as the lack of a coherent receiver and limited resistance to PO and FO, make it a less appealing scheme than FSCSS-IM. It may be noticed that a coherent detection is possible for FSCSS-IM and it is sufficiently robust against the offsets when ς = 2. For higher values of ς, this robustness may be negatively affected.

VI. CONCLUSIONS

Although LoRa is one of the most promising and widespread CSS-based PHY layer scheme employed in LP-WANs that can target a multitude of applications, its most significant limiting factor is low transmission rates. A number of different CSS alternatives to LoRa have been proposed in the literature, which are capable of either transmitting a higher number of bits per symbol and/or more energy efficient. The current state of the art primarily examines LoRa for IoT system assessment, with little to no consideration devoted to LoRa alternatives for IoT deployment and system evaluation. In this survey, we present a comprehensive waveform design for some CSS alternatives to LoRa. It has been observed that different schemes provide different spectral efficiencies. Based on achievable spectral efficiencies, the CSS schemes are divided into different groups. Moreover, the number of chirps that are multiplexed in the transmit symbol is another factor used to categorize the schemes. It has been observed that the schemes that multiplex more than one chirp are capable of achieving higher SE, however, are less robust against the phase and frequency offsets. On the contrary, the schemes that employ only a single chirp have lower spectral efficiencies but are resilient against the offsets. Although the IM-based techniques may achieve a wide range of various spectrum efficiencies, they are often less effective than LoRa in terms of EE. Moreover, it has been observed that the schemes which use either PSs to transmit information, or use in-phase and quadrature components to transmit additional information compared to LoRa are generally less robust against the offsets. Furthermore, these schemes cannot have a non-coherent detection mechanism due to loss of phase information. We have also observed that although the coherent detection performs relatively better against the non-coherent detection, its performance degrades significantly in the presence of phase and frequency offsets. It is also highlighted that the non-coherent detection does not care about the knowledge of the
CSI, whereas, for coherent detection, it is mandatory. It is foreseen that the waveform design presented in this work would encourage future research.

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| Modulation   | Const. Env. | SE imp. | EE imp. | Coh. Detec. | Non-Coh. Detec. | Robustness to PO | Robustness to FO |
|--------------|-------------|---------|---------|-------------|----------------|-----------------|-----------------|
| LoRa         | ✓           | X       | ✓       | ✓           | ✓              | ✓               | ✓               |
| E-LoRa       | ✓           | +       | +       | ✓           | ✓              | X               | ✓               |
| ICS-LoRa     | ✓           | +       | ✓       | ✓           | ✓              | ✓               | ✓               |
| PSK-LoRa     | ✓           | +       | ✓       | ✓           | ✓              | ✓               | ✓               |
| DO-CSS       | X           | X       | ✓       | ✓           | ✓              | ✓               | ✓               |
| IQ-CSS       | X           | ✓       | ✓       | ✓           | ✓              | ✓               | ✓               |
| DCRK-CSS     | ✓           | +       | ✓       | ✓           | ✓              | ✓               | ✓               |
| FSCSS-IM     | X           | ++      | X       | ✓           | ✓              | ✓               | ✓               |
| IQ-CIM       | X           | ++      | X       | ✓           | ✓              | ✓               | ✓               |
| SSK-ICS-LoRa | ✓           | +       | ++      | ✓           | ✓              | X               | ✓               |
| ePSK-CSS     | X           | +       | ++      | ✓           | ✓              | ✓               | ✓               |
| GCSS         | X           | ++      | X       | ✓           | ✓              | ✓               | ✓               |
| TDM-CSS      | X           | ++      | +       | ✓           | ✓              | ✓               | ✓               |
| IQ-TDM-CSS   | X           | ++      | ++      | ✓           | ✓              | ✓               | ✓               |
| DM-CSS       | X           | ++      | ++      | ✓           | ✓              | ✓               | ✓               |
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