A Spectral Efficiency Enhancement for Chirp Spread Spectrum Downlink Communications

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***Abstract*—The objective of the paper is to propose a transmitter/receiver system that increases the spectral efficiency in the context of CSS-type downlink communications. For this purpose, we propose to transmit simultaneously on the same time/frequency resources signals intended for several receivers. By controlling the level of interference between the different users and the properties of quasi-orthogonality between up and down chirps, we show that each receiver can correctly demodulate the symbols intended for it.**

***Index Terms*—NOMA, Chirp Spread Spectrum, Successive Interference Cancellation, LoRa, IoT**

I. INTRODUCTION

The paper proposes to address the congestion problem of the radio interface of IoT networks. This problem is all the more true in industrial, scientific and medical (ISM) bands. Indeed, the exponential deployment of connected objects associated with the random access to the channel considerably increases the probability of packet collisions. This subject has moreover been the subject of recent papers such as [1], [2], [3] and [4]. Among the popular communication technologies used in ISM bands, there is the one patented by Semtech: LoRa, which is based on chirp spread spectrum (CSS) modulation. This communication technique is based on a parameter, commonly called spreading factor (SF). This parameter improves the receiver sensitivity and thus the range of communications at the cost of a reduction in spectral efficiency. Indeed, an increase of one in the SF leads to a spreading gain of 3dB. CSS is naturally robust to narrowband interference and LoRa signals with different SF but not those signals with the same SF. Indeed, when two or more signals of the same SF collide at the receiver, at best it correctly demodulates only one packet. More specifically, the CSS demodulation estimates the frequency that maximizes a periodogram to retrieve symbols. Thus, if a received packet with higher power than all the others, generally 6 dB in a synchronized context, the search fo the optimum frequency will lead to estimate correctly the symbol. The possible estimation errors that may occur are due to the proximity of frequencies associated with other users' symbols. This symbol-to-symbol interference is compensated by the association of a bit interleaver, a gray coding of the symbols and an error-correcting code with the ability to correct one bit. It is for these reasons that LoRa technology uses an (8,4) Hamming code.

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There are several methods to avoid packet losses. First of all, the level 2 protocol side (MAC layer) exploits time and frequency diversity by allowing the connected device to send back the information multiple times to receive an acknowledgment (ACK) from the gateway [5], [6]. This attempt occurs within the limit of the duty cycle to prevents the device to degrade the network congestion. On the physical (PHY) layer side, recent works proposed particularly to implement successive interference cancellation (SIC) algorithms [7] to deal with these collisions and retrieve all the signals. PHY layer proposals provide huge advantages regarding the spectral efficiency since the nodes do not need to send back the information multiple times which is not the case for the MAC layer solution. The [8] solution is also worth noticing since it proposes to divide the frequency band into two parts to allow the simultaneous transmission of two users but still at the expense of spectral efficiency. This idea of the simultaneous transmission to reduce the packet loss was also studied by [9], which provided some feasibility insights by using slope-opposite chirps. Based on this idea, in this paper, we propose to enhance the spectral efficiency of a CSS system by combining [9] with an extension of our previous work in [10] for an uplink scenario. With this combination, we can transmit on the same time-frequency resource at least four non- or quasi-orthogonal signals. The simulations results demonstrate the relevance of our approach in a various scenario. Essentially, the gateway uses SF 8 to 12 the BER curves of each node are superposed, showing that the four nodes are quasi-orthogonal. Considering the non-perfect synchronization, some of the nodes experiences BER error floor, i.e. increasing the SNR the BER does not decrease. The error floor reduces as the gateway selects high spreading factors.

The paper is organized as follows. In section II we briefly explain the CSS principle. The system model that we consider is exposed in section III, whereas the proposed receiver is detailed in section IV. Before concluding, we present different simulation results in sections VI and V, respectively.

II. BRIEF DESCRIPTION OF CSS

In the following, we recall the CSS PHY layer principle. For more information about the CSS modulation and demodulation principle, the reader are invited to read [11].

A. CSS transmission

Spread spectrum techniques [12] are used in communication systems to spread a signal in the frequency domain. This is beneficial for instance to improve resistance against noise and intentional interference. Particularly, the LoRa technology adopts the chirp spread spectrum (CSS) technique in which bit sequences are modulated by using frequency modulated sinusoidal signals known as chirps [13]. A chirp may have a frequency increase or decrease over time which causes a spread along the signal bandwidth B. That is, the chirp frequency varies between f_i and f_f, so that $B = |f_i - f_f|$, during the symbol period T. When the chirp has a frequency increase, i.e., $f_i < f_f$, it is referred to as an *up-chirp*; otherwise, it is referred as a *down-chirp*.

More specifically, let M be the modulation cardinality, i.e., the number of symbols that can be generated by the CSS modulation. The spreading factor (SF) of the CSS modulation has the following ratio with the bandwidth B:

$$B \cdot T = 2^{\text{SF}}, \tag{1}$$

where T is the symbol time. Besides, let SF – a.k.a. spreading factor – be the number of bits per symbol, each CSS symbol then browsing the time-frequency surface with a specific trajectory. This gives the notion of spread spectrum as a way to code the information into different formats, following a modulo T on the base chirp which allows us M chirp trajectories of the same duration. Hence, the CSS signal, considering a binary sequence, is formed by several trajectories, randomly selected from the chirp sequences [14].

B. Receiver principle

As an adaptive modulation scheme, the LoRa modulation allows the selection of the SF so that one can create, for different values of the SF, several orthogonal channels in the same band [15].

Therefore, within an orthogonal channel, the detection of LoRa signals is performed when the receiver is in listening mode. Synchronization and detection method of Lora-like signals can be found in [16], which implements the procedure described in the patent describing the LoRa [17].

The de-chirping operation is performed by the CSS demodulator once the signals are synchronized, as proposed in [18]. This operation usually multiplies the received signal by a reference (e.g., an up- or down-chirp signal modulated with symbol zero) and applies the FFT operation. The receiver then needs to estimate the most likely symbol by searching into the set of chirp sequences which one maximizes the absolute value of the FFT.

Nonetheless, since we are assuming to transmit a superposed signal composed of both up- and down-chirp, we need some extra processing in order to properly recover the information which is conveyed by the superposed signals.

In [9], the authors address the quasi-orthogonality property between up- and down-chirp LoRa signals. Specifically, according to [9] after a de-chirping operation for a chirp type, the residual contribution of the other type is negligible in the resulting signal.

Since we may use different SFs for up- and down-chirps, this will be the basis for the construction of the proposed receiver in the later sections.

III. LoRa communication scenario

We consider a communication system where several CSS signals are simultaneously transmitted on the same channel and within the same symbol time T. In practice, this happens when the gateway receives several superposed signals in uplink and is able to decode all of them. A dedicated algorithm, such as the one in [7], decodes the superposed LoRa signals. In turn, the gateway sends the acknowledgments for the corresponding nodes using the same time window. That is, each node will receive all the acknowledgements simultaneously and decode its signal by using a similar reception technique used by the gateway in the uplink. In what follows, the system model describes downlink signals.

In this work, we focus on the simultaneous transmission of four LoRa signals, each one intended for a given node. To this end, we leverage the idea proposed in [9] that exploits the quasi-orthogonality between up- and down-chirp LoRa signals. Then, for each chirp type, we apply a controlled time desynchronization between two signals. In other words, the gateway can simultaneously transmit two down-chirp LoRa signals desynchronized from each other, and two up-chirp LoRa signals also desynchronized from each other.

We first define the two chirp signal types. Let $s_i^u(t)$ be the complex envelope of an up-chirp signal transmitted by the gateway that conveys the payload intended to node i. It can be defined as

$$s_i^u(t) = \sum_{k=0}^{K_i-1} e^{j\phi_{k,i}^u(t-kT)}\,, \quad i \in \{1, 2\}\,, \tag{2}$$

where K_i is the number of transmitted chirps indexed by k, and $\phi_{k,i}^u(t)$ stands for the phase of the k-th complex signal intended to the i-th node. The phase $\phi_{k,i}^u(t)$ is given by

$$\phi_{k,i}^u(t) = 2\pi \int_{-\infty}^{t} f_{k,i}^u(t')\mathrm{d}t'\,, \tag{3}$$

where $f_{k,i}^u(t)$ is the k-th modulated up-chirp for node i. Let $m_i(k) \in \{0, \ldots, M-1\}$ denote the k-th transmitted symbol intended to the i-th node. Then, the up-chirp frequency $f_{k,i}^u(t) \in [0, B]$ of a LoRa signal can be expressed as

$$f_{k,i}^u(t) = \begin{cases} \frac{B}{T}t + \frac{m_i(k)}{T} & \text{if} \quad t \in \mathcal{T}_a\,, \\ \frac{B}{T}t - \left(1 - \frac{m_i(k)}{M}\right)B & \text{if} \quad t \in \mathcal{T}_b\,, \end{cases} \tag{4}$$

where $\mathcal{T}_a = [kT\,, T - \gamma_i(k))$ and $\mathcal{T}_b = [T - \gamma_i(k)\,, (k+1)T)$, with $\gamma_i(k) = \frac{m_i(k)}{B}$, are adjacent time intervals where the chirp signal is defined. Now, plugging (4) into (3), we obtain

$$\phi_{k,i}^{u}(t)=\begin{cases}2\pi\left(\frac{B}{2T}t^2+\frac{m_i(k)}{T}t\right) & \text{if} \quad t\in\mathcal{T}_a\,,\\ 2\pi\left[\frac{B}{2T}t^2+\left(\frac{m_i(k)}{T}-B\right)t\right] & \text{if} \quad t\in\mathcal{T}_b\,.\end{cases} \tag{5}$$

Similarly, let $s_j^d(t)$ be the complex envelope of a down-chirp signal transmitted by the gateway that conveys the payload intended to node i, which is defined as

$$s_i^d(t)=\sum_{k=0}^{K_i-1} e^{j\phi_{k,i}^{d}(t-kT)}\,, \quad i\in\{3,4\}\,. \tag{6}$$

The down-chirp phase $\phi_{k,i}^{d}(t-kT)$ – now assuming that the chirp frequency decreases linearly wrapping around in $[0,B]$ – can be derived as

$$\phi_{k,i}^{d}(t)=\begin{cases}2\pi\left[-\frac{B}{2T}t^2+\left(B-\frac{m_i(k)}{T}\right)t\right] & \text{if } t\in\mathcal{T}_a\,,\\ 2\pi\left[-\frac{B}{2T}t^2+\left(2B-\frac{m_i(k)}{T}\right)t\right] & \text{if } t\in\mathcal{T}_b\,.\end{cases} \tag{7}$$

We can now express the signal $s(t)$ in the downlink as follows:

$$\begin{aligned}s(t)&=\sqrt{P_1}s_1^u(t)+\sqrt{P_2}s_2^u(t-\delta t_{1,2})\\ &\quad+\sqrt{P_3}s_3^d(t)+\sqrt{P_4}s_4^d(t-\delta t_{3,4}),\end{aligned} \tag{8}$$

where $\delta t_{1,2}$ and $\delta t_{3,4}\in[0,T)$ corresponds to the time desynchronization between nodes 1 and 2, and nodes 3 and 4, respectively, and P_i is the transmit power allocated to the i-th node. Without loss of generality, we assume $\delta t_{1,2}=\delta t_{3,4}=\delta t$.

In order for a LoRa node to detect and synchronize on $s(t)$, all nodes share the same preamble. Thus, a preamble signal $s_{train}(t)$ of duration T_t similar for both nodes is transmitted prior to signal $s(t)$. Therefore, the signal $x(t)$ transmitted by the gateway can be expressed as:

$$x(t)=\sqrt{P}s_{train}(t)+s(t-T_t)\,, \tag{9}$$

with the transmit power $P=\sum_{i=1}^{4}P_i$.

To illustrate the proposed LoRa superposed signal transmission, Figure 1 shows the signal design in time with a common preamble of duration T_t followed by two up-chirps and two down-chirps comprising a superposed signal of duration T. Signals of a same chirp type are desynchronized by δt.

A. Received signals

We consider a communication through a non-frequency selective channel with complex-valued channel gain $h_i\in\mathbb{C}$. The discrete received signal at the ith node, sampled at $T_s=\frac{1}{uB}$ $(u\geq 1)$, is given by:

$$r_i(n)=h_i x(n-\Delta n_i)e^{j2\pi n T_s\Delta f_i+j\theta_i}+w_i(n)\,, \tag{10}$$

where $w_i(n)\sim\mathcal{N}_{\mathbb{C}}(0,\sigma_w^2)$ is the additive white Gaussian noise (AWGN). Δn_i and Δf_i stand for the time and frequency desynchronization respectively while θ_i represents the initial phase.

To detect the reception of LoRa-like signals, the receiver has to be in a listening mode. A step by step method to detect and synchronize LoRa-like signal is described in [16]. It should be noted that the method is based on the patent [19] written by the LoRa inventors.

Once synchronized[1], the CSS demodulator needs to perform a de-chirping operation such as explained in [18]. Then, it needs to estimate the most likely symbol by looking for the index $j\in\{0,\ldots,M-1\}$ which maximize a FFT absolute value. However, given the proposed communication strategy, some additional processing must be implemented so that the nodes can correctly demodulate all the superposed signals sent by the gateway. In the next section, we propose an original way to perform the demodulation of four interfering LoRa-like signals.

IV. Proposed receiver

Before the signal detection, each node has to be able to synchronize the LoRa frame in time and frequency domains, i.e. they compensate the channel delay and Doppler shift using synchronization algorithm. Nodes 1, 2, 3 and 4 share the same $s_{train}(t)$, so the synchronization is achieved similarly. Nevertheless the gateway employs a non-orthogonal transmission scheme, each node cancels the interference signal out for detection of its own data. We assume a dedicated bit to identify the corresponding signal of the node. In the following, we describe the decoding processing on the node level.

We assume that after the time synchronization processing, any desynchronization applied into signal $s_j(t)$ of node j is removed from node j's perspective. Thus, for nodes $j\in\{2,4\}$ that, according to Eq. (8), have their signals desynchronized by δt in the transmission, the post-synchronization received signal $y_j(n)$ can expressed as

$$\begin{aligned}y_j(n)&=\sqrt{P_1}h_j s_1^u(n+\delta n)+\sqrt{P_2}h_j s_2^u(n)\\ &\quad+\sqrt{P_3}h_j s_3^d(n+\delta n)+\sqrt{P_4}h_j s_4^d(n)+w_j(n),\end{aligned} \tag{11}$$

while for nodes $i\in\{1,3\}$,

$$\begin{aligned}y_i(n)&=\sqrt{P_1}h_i s_1^u(n)+\sqrt{P_2}h_i s_2^u(n-\delta n)\\ &\quad+\sqrt{P_3}h_i s_3^d(n)+\sqrt{P_4}h_i s_4^d(n-\delta n)+w_i(n),\end{aligned} \tag{12}$$

where $\delta n=\frac{\delta t}{T_s}\in\{0,\ldots,M-1\}$.

The CSS demodulation consists of multiplying the received signal by a sequence of down-chirp and up-chirp signals. Then, the symbol estimation is based on FFT processing.

Assuming the demodulation down-chirp portion, the signal obtained after the FFT and which corresponds to the processing of k-th chirp of node i is equal to:

$$Y_i^d(l,k)=\frac{1}{\sqrt{M}}\sum_{n=0}^{M-1}\underbrace{\left(y_i(n,k)e^{j2\pi\frac{B}{2T}(nT_s)^2}\right)}_{z_i^d(n,k)}e^{-j2\pi\frac{ln}{M}} \tag{13}$$

[1] The goal of this paper is not to propose a novel way to synchronize in time and frequency the CSS signals. To that end, for our simulation results we use solutions proposed in [16].

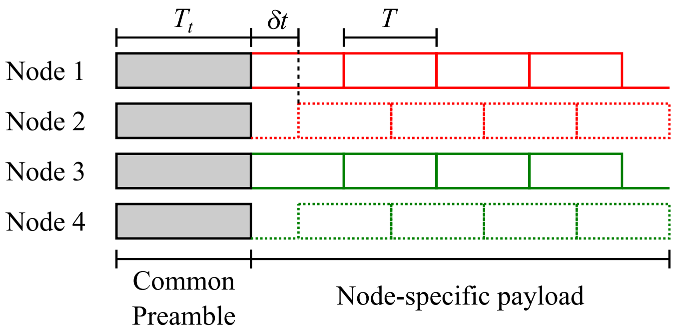

Fig. 1: LoRa-Like signal design comprising four simultaneous signal transmissions.

where $y_i(n,k) = y_i(n+kM)\ \forall n \in \{0,\ldots,M-1\}$. Analogously, $Y_i^u(l,k)$ can be obtained assuming an up-chirp demodulation with $z_i^u(n,k) = y_i(n,k)e^{-j2\pi\frac{B}{2T}(nT_s)^2}$. For example, for $i=1$ we derive $z_1^d(n,p)$ by considering Eq. (8), definitions (5) and (7), and after some calculations based on [20]:

$$\begin{aligned} z_1^d(n,p) \approx\ & h_1\sqrt{P_1}e^{j2\pi\frac{nm_1(p)}{M}} \\ &+ h_1\sqrt{P_2}e^{j2\pi\frac{n\bar{m}_2(p-1)}{M}}\mathbb{1}_{[0,\delta n-1]}(n) \\ &+ h_1\sqrt{P_2}e^{j2\pi\frac{n\bar{m}_2(p)}{M}}\mathbb{1}_{[\delta n,M-1]}(n) \\ &+ h_1\sqrt{P_4}e^{j2\pi\frac{n^2[1-M]+n\bar{m}_4(p-1)}{M}}\mathbb{1}_{[0,\delta n-1]}(n) \\ &+ h_1\sqrt{P_4}e^{j2\pi\frac{n^2+n\bar{m}_4(p)}{M}}\mathbb{1}_{[\delta n,M-1]}(n) + \tilde{w}_1(n) \end{aligned} \tag{14}$$

where $\mathbb{1}_{[0,x)}(t)$ stands for the indicator function, and

$$m_2(p) = (\delta n + \bar{m}_2(p)) \mod M \tag{15}$$
$$m_2(p-1) = (\delta n + \bar{m}_2(p-1)) \mod M \tag{16}$$
$$m_4(p) = (\delta n + \bar{m}_4(p)) \mod M \tag{17}$$
$$m_4(p-1) = (\delta n + \bar{m}_4(p-1)) \mod M. \tag{18}$$

Note that, the interference of node 3 is negligible since up- and down-chirp are quasi-orthogonal signals.

Finally, Eq. (14) is compactly expressed as

$$Y_1^d(l,k) = h_1\sqrt{P_1M}\phi(l-m_1(k)) + I_1(l,k) + \tilde{W}_1(l), \tag{19}$$

where $\phi(\cdot)$ is the Dirac function, $I_1(l,k)$ accounts for the interference at the output of the FFT and $\tilde{W}_1(l) \sim \mathcal{N}_{\mathbb{C}}(0,\sigma_w^2)$ is the FFT of the noise $\tilde{w}_1(n)$. Similarly, we can obtain to all nodes the received signal expressions using the up- and down-chirp signals at the de-chirping step.

A. Interference Cancellation

The strategy for canceling out the interference involves allocating different powers and applying a successive interference cancellation (SIC) algorithm. The power allocation between nodes that are asynchronous by half of the symbol period is essential to limit the effect of the inter-symbol interference (ISI). The node with more power will be first to be detected, and then canceling its contribution in the received signal expressed.

We assume $P_1 = P_3$, $P_2 = P_4$, and the power rate $PR = 10\log_{10}\left(\frac{P_1}{P_2}\right) \geq 0$. In this case, node 2 estimates the payload of node 1 and reconstructs its LoRa signal to remove it from the synchronized signal defined in Eq. (11). Similarly, node 4 estimates node 3 to remove its contribution from the synchronized signal. Additionally, we explore the fact that nodes in up-chirp leads to negligible interference to nodes in down-chirp and vice versa.

V. Simulation Results

We evaluate the proposed transmission scheme using a LoRa signal simulator that implements interleaving and channel coding blocks to encode the bits before the chirp generation. The use of such blocks increases the system robustness against interfering bursts and off-by-one errors [19]. Furthermore, we assume the signal model expressed in (10).

The interleaving block scrambles the data bits throughout the packet to turn the transmission more resilient to bursts of interference. In practical modems, it is combined with forward error correction (FEC) as shown in [21]. The interleaving block in our simulator implements the diagonal interleaver according to the patent [19].

FEC block controls data transmission over unreliable or noisy communication channels. The enconder used in LoRa is Hamming FEC with a variable codeword size ranging from 5 to 8 bits [19]. In practice, the data size per codeword is 4 bits, which leads to a coding rate equat to $\frac{4}{4+CR}$, where $CR \in \{1,\ldots 4\}$ is the number of redundancy bits.

We evaluate the transmission scheme by means of the bit error rate (BER) with the user signal-to-noise ratio (SNR), where the latter defines the SNR at the ith node as P_i/σ_w^2. In addition, when time and frequency synchronization are performed, Δf_i and Δn_i are uniformly distributed in $\pm\frac{B}{4}$ and $\pm 10T$, respectively. The system bandwidth $B = 125$ Hz, CR $= \{1,2,3,4\}$, PR $= 6$ dB, the number of bits at the LoRa modulator input is 300, and number of monte-carlo simulations is 2000. We summarize the simulation parameters in Table I.

TABLE I: Simulation parameter.

Parameter	Values
Bandwidth	125 Hz
CR	$\{1,2,3,4\}$
SF	$\{7,\ldots,12\}$
Number of bits	300
Power ratio	6 dB
$\delta t_{1,2}$	$T/2$
$\delta t_{3,4}$	$T/2$
Δf_i	$\mathcal{U}\left(-\frac{B}{4},\frac{B}{4}\right)$
Δn_i	$\mathcal{U}(-10T,10T)$
Monte-Carlo	2000

Figure 2 compares the BER performance among the proposed LoRa transmission scheme, the proposal in [9], and [20]. In [20], the LoRa receiver uses down-chirp mode and

TABLE II: Transmission scheme configuration.

Node	Chirp type	δt	Power
1	Up	0	P_{dBm}
2	Up	$T/2$	$P_{dBm} - 6\text{dB}$
3	Down	0	P_{dBm}
4	Down	$T/2$	$P_{dBm} - 6\text{dB}$

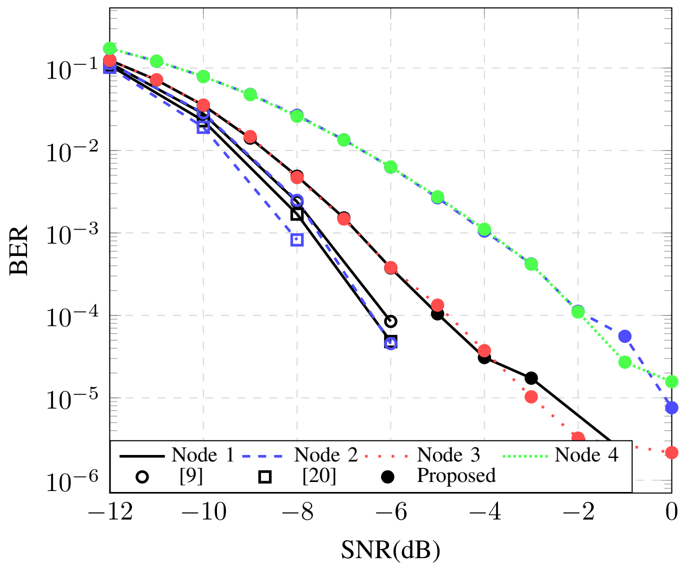

Fig. 2: The BER curves show the performance of the proposed method and those proposed in [20] and [9].

SIC algorithm to separate the nodes 1 and 2 while, in [9], the LoRa receiver exploits quasi-orthogonal property between up and down chirps. From Figure 2, we see that both approaches have similar performances.

In the proposed scheme, the nodes served with LoRa signal delayed by $T/2$ has a performance loss of 2 dB. This is because in the process of decoding the information of these signals, there is cross product of down-chirp delayed signal with up-chirp signal. In this situation, the quasi-orthogonality condition does not hold. One way to improve the performance of nodes 2 and 4 is to increase the code rate, as showin in 3. With this approach, the system increases the robustness by adding redundancy to the LoRa packet. However, we still observe a gap of 3 dB when we compare to nodes 2 and 4 with nodes 1 and 3.

Fig. 4 shows that all nodes converge to the same BER performance if the system increases the spreading factor. Such a convergence is observed at spreading factor 10 and to higher values. This is mainly because the higher the spreading factor is, the more robust to interference the LoRa signal is.

Fig. 5 shows the BER performance of the each node under different spreading factors assuming non-perfect time-frequency synchronization. To correct the time-frequency shift, we implement the algorithm proposed in [16]. From the curves, we can see that nodes whose LoRa signal a dele

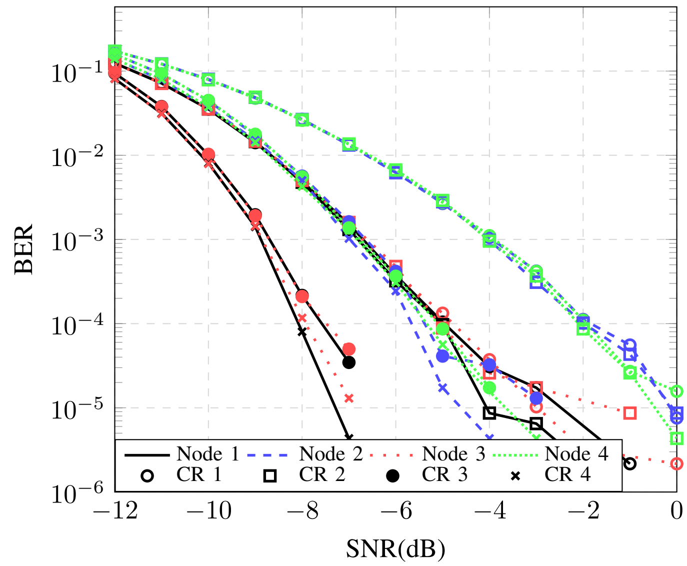

Fig. 3: BER curves for different values of CR assuming perfect time/frequency synchronizations.

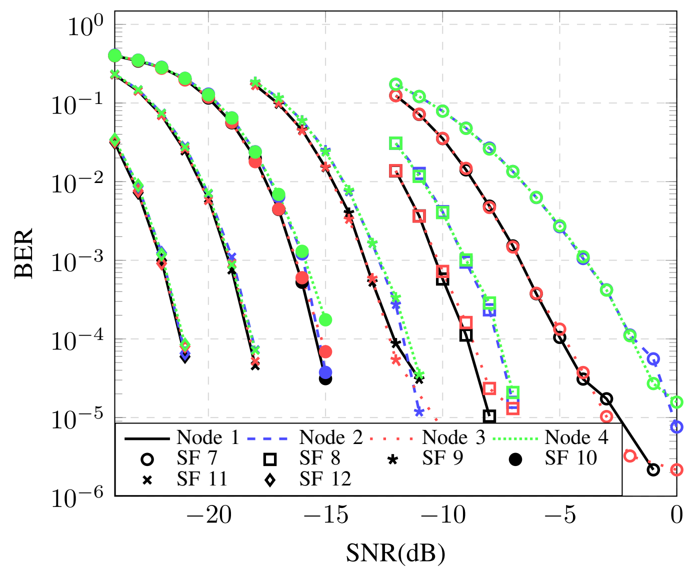

Fig. 4: BER curves for nodes perfect time/frequency synchronizations.

VI. Conclusions

This work presented a novel LoRa transmission scheme that increases the spectral efficiency by serving at least 4 nodes at the same time. The BER curves of the nodes are similar considering SF 9 to 12, which demonstrates the quasi-orthogonality property of the proposed scheme. Assuming SF 7 to 8, it is visible that nodes whose signals are delayed suffer with residual interference caused by the de-synchronized cross product between the up-chirp and down-chirp signal. More specifically, the quasi-orthogonality property of up-chirp and

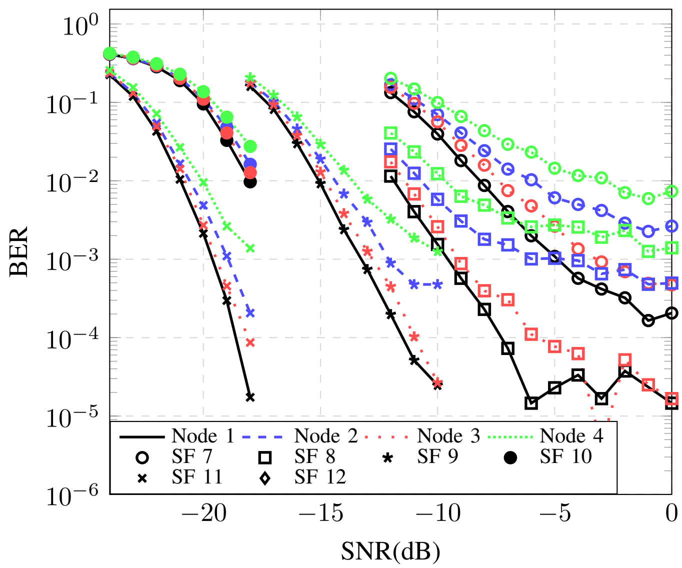

Fig. 5: BER curves for nodes non-perfect time/frequency synchronizations.

down-chirp is true if they are synchronized. The non-perfect time-frequency compensation breaks quasi-orthogonality condition at low spreading factors, such as SF 7 to 8. To more robust configurations, this property still holds.

As prospects, we plan to propose an optimization of the power diversity distribution to improve the BER performance of lower spreading factors. Additionally, we intend investigating how PAPR behaviours by applying the proposed scheme.

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