Time-Delay-Estimation-Liked Detection Algorithm for LoRa Signals Over Multipath Channels

Yurong Guo and Zujun Liu

***Abstract*—LoRa promises long range reliable communication for the Internet of Things (IoT). However, multipath fading severely affects LoRa's BER performance. In this letter, we propose a simple detection algorithm for the frame-based LoRa PHY signals mainly composed of a preamble and a variable-length PHY-payload. We first prove that detecting the frame-based LoRa signals over multipath fading channels can be formulated as a Time-delay-estimation (TDE)-liked problem. Based on this, a cyclic cross-correlation implemented by a matched filter is utilized to detect the LoRa signals. Additionally, the proposed TDE-liked detection algorithm is insensitive to the integer frequency offset due to the nature of implying a differential operation. The simulation results indicate that, the proposed algorithm significantly outperforms the existing detection algorithms for the LoRa signals over multipath fading channels. Under the multipath fading channel, the BER performance of the proposed algorithm is slightly inferior to that in the AWGN channel while the existing scheme fails to work.**

***Index Terms*—LoRa, Internet of Things (IoT), cross-correlation, multipath channels.**

I. Introduction

LoRa is an attractive Low Power Wide Area Network (LPWAN) physical layer technology and promises long range low power communication for the Internet of Things (IoT) [1], [2]. LoRa modulation is based on chirp spread spectrum (CSS), where the chirp modulation rates depends on the spreading factors [3], [4].

LoRa signals use the initial phase of a CSS signal to carry information, so it is also recognized as a frequency shift chirp modulation (FSCM) [5]. For detecting LoRa signals, the existing scheme adopts the direct non-coherent detection algorithm (DNC) [5]–[8]. However, many works show that LoRa's BER performance will decline dramatically over multipath fading channels. For instance, compared with the performance under the AWGN channel, LoRa signals would suffer from 3dB performance loss over a 2-path fading channel [5] and a performance degradation over 20dB will occur under Rayleigh fading channels when BER $= 10^{-3}$ [6]. The measurement study also finds that the performance loss under non-line-of-sight (NLOS) environments is severely affected [9]. Therefore, LoRa is incapable of supporting its promising long range communication in urban environment [6].

To resist the multipath fading, a frequency domain equalization (FDE) before dechirping has been proposed for the CSS system [10]. However, due to the low SNR regime of interests and the noise amplification, FDE cannot effectively address the issue of LoRa signals over the multipath channels. Based on the fact that a LoRa PHY frame is generally composed of a preamble and a PHY payload carrying the messages [11], we propose a simple detection algorithm to overcome the severe negative influence caused by multipath fading. We first formulate detecting the frame-based LoRa signals as a time-delay-estimation (TDE)-liked problem. And then we adopt the cyclic cross-correlation implemented by a matched filter to detect LoRa signals. Through simulations, the proposed TDE-liked detection algorithm shows a significant BER performance gain and promises high reliable communication for LoRa signals over multipath fading channels.

II. System Model

The basis chirp signal in analog form is written by [12]

$$c(t) = e^{j2\pi \int_0^t f(t)d(t)} = e^{j2\pi \cdot \frac{t^2}{2} \cdot \frac{B}{T_s}} \quad (1)$$

where $0 \leq t < T_s$, B is the signal bandwidth and T_s is the chirp symbol period. The frequency of the continuous-time basis chirp signal is $f(t) = \frac{t}{T_s} \cdot B$.

After sampling, the basis chirp signal in discrete-form is given as

$$c(kT) = \frac{1}{\sqrt{2^{SF}}} e^{j2\pi \frac{(kT)^2}{2} \frac{B}{T_s}} = \frac{1}{\sqrt{2^{SF}}} e^{\pi \frac{k^2}{2^{SF}}} \quad (2)$$

where $T = \frac{1}{B}$ represents the sampling interval and $T_s = 2^{SF}\, T$ is the chirp symbol period. The value of k ranges from 0 to $2^{SF} - 1$ where $SF \in \{7, 8, 9, \ldots, 12\}$ is the spread factor.

The modulated basic chirp with the modulation information p is given by [12]

$$c(kT,\ p) = \frac{1}{\sqrt{2^{SF}}} e^{j\pi [(p+k)\ mod\ 2^{SF}]^2 \frac{1}{2^{SF}}} \quad (3)$$

where $p \in \{0, 1, \ldots, 2^{SF} - 1\}$ is a non-negative integer number converted from SF binary information bits.

In order to ensure the phase continuity between the consequent chirp symbols, we slightly modified the LoRa modulation signal by introducing a phase of $e^{-j\pi \frac{p^2}{2^{SF}}}$ as

$$c(kT,\ p) = \frac{1}{\sqrt{2^{SF}}} e^{j\pi [(p+k) mod 2^{SF}]^2 \frac{1}{2^{SF}}} \cdot e^{-j\pi \frac{p^2}{2^{SF}}} \quad (4)$$

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It can be easily obtained that, for arbitrary $(p + k) \in \mathbb{N}_+$, Eq. (4) can be simplified to

$$c(kT,\ p) = \frac{1}{\sqrt{2^{SF}}} e^{j\pi(k^2+2pk)\frac{1}{2^{SF}}} \tag{5}$$

Here we assume that the multipath fading channel denoted by $h(k)$ has L paths and is of block fading, i.e., $h(k)$ keeps invariant within a LoRa PHY frame. Then, the received signal over $h(k)$ is written by

$$r(p,\ kT) = c(p,\ kT) \otimes h(kT) + n(p,\ kT) \tag{6}$$

where $\otimes$ is convolution operation and $n(p,\ kT)$ is the additive white Gaussian noise.

For LoRa demodulation, the existing widely used scheme is the DNC detection [5]–[8]. To be more specific, it firstly executes the inner product as

$$\begin{aligned} y(q) &= \langle r(p,kT), c(q,kT)\rangle \\ &= \sum_{k=0}^{2^{SF}-1} r(p,\ kT) \cdot c^*(q,kT) \\ &= \sum_{k=0}^{2^{SF}-1} r(p,\ kT) e^{-j\pi\frac{k^2}{2^{SF}}} \frac{1}{\sqrt{2^{SF}}} e^{-j2\pi qk\frac{1}{2^{SF}}} \end{aligned} \tag{7}$$

where $<>$ depicts the inner product operation and $q \in \{0, 1, \ldots, 2^{SF} - 1\}$ depicts the index of candidate reference symbols.

From (7), the inner product process consists of two steps: the sample-based multiplying the received signal by the local down-chirp $e^{-j\pi\frac{k^2}{2^{SF}}}$ (i.e., dechirp) and taking Discrete Fourier Transform (DFT) on the dechirped signal. Hereafter, the detection of LoRa symbols reverts to selecting the index of the DFT output that exhibits the highest magnitude, i.e.,

$$\widehat{p} = \arg\max_q(|y(q)|) \tag{8}$$

where q can also be viewed as the index of the DFT outputs in frequency domain.

III. TDE-Liked Detection Algorithm for the Frame-Based LoRa Signals

A LoRa PHY frame is mainly composed of a preamble and a variable-length payload. The preamble contains several unmodulated CSS symbols and the payload contains several modulated CSS symbols. In this section, we will propose a TDE-liked detection algorithm for the frame-based LoRa signals over multipath channels and utilize the cyclic cross correlation to detect the message.

A. TDE-Liked Detection Problem Formulation

For the sake of discussion, here we refer to the DNC demodulation as the process including the inner production given in (7) (dechirp and DFT) and taking modulus on the output of the inner production.

Proposition I: After performing the DNC demodulation on the frame-based LoRa signals over multipath channels, detecting the information p can be formulated as a TDE-liked problem.

Proof: Performing the inner product between the received LoRa payload signal $r(p,\ kT)$ and the basis chirp signal is given by

$$\begin{aligned} y_1(q) &= \langle r(p,kT), c(q,kT)\rangle \\ &= \frac{1}{\sqrt{2^{SF}}} \sum_{k=0}^{2^{SF}-1} \sum_{i=0}^{L-1} h(i) c(p + kT - i) e^{-j\pi(2qk+k^2)\frac{1}{2^{SF}}} + n_1 \\ &= \frac{\pi}{2^{SF-1}} \sum_{i=0}^{L-1} h((i)\ mod\ 2^{SF}) e^{j\frac{\pi i^2}{2^{SF}}} e^{\frac{-j2\pi pi}{2^{SF}}} \delta(q + i - p) + n_1 \end{aligned} \tag{9}$$

where $\delta()$ is Dirac function, n_1 depicts the inner product result of the AWGN $n(p,\ kT)$ and $c(kT)$. Taking modulus on $y_1(q)$, we can obtain

$$|y_1(q)| = |\frac{\pi}{2^{SF-1}} \sum_{i=0}^{L-1} h((i)\ mod\ 2^{SF}) e^{\frac{j\pi i^2}{2^{SF}}} e^{\frac{-j2\pi pi}{2^{SF}}} \delta(q + i - p)| + C_1 \tag{10}$$

where C_1 is the noise term. (10) can be further written by

$$\begin{aligned} |y_1(q)| &= |\frac{\pi}{2^{SF-1}} h((-(q - p))\ mod\ 2^{SF})\ | + C_1 \\ &\triangleq \widehat{h}(q - p) + C_1 \end{aligned} \tag{11}$$

The LoRa preamble is composed of several unmodulated basis CSS signals which can be considered as the modulated CSS signal with $p = 0$. Therefore, similar to the derivation of (11), performing the direct non-coherent demodulation on the LoRa preamble over $h(k)$ yields

$$\begin{aligned} |y_2(q)| &= |\frac{\pi}{2^{SF-1}} h((-q)\ mod\ 2^{SF})| + C_2 \\ &\triangleq \widehat{h}(q) + C_2 \end{aligned} \tag{12}$$

where C_2 is the noise term.

From (11) and (12), we observe that $\widehat{h}(q - p)$ is the version of cyclic shifting $\widehat{h}(q)$ with p in frequency domain. Here we can find p by using the classical cyclic cross-correlation method to obtain the maximum value, so it can be formulated as an estimation problem as

$$\widehat{p} = \arg\max_\tau(R(\tau)) \tag{13}$$

where $R(\tau) = \mathbb{E}\{|y_2(q)||y_1(q + \tau)|\}$.

Such estimation problem is very similar to the classic time delay estimation problem [13], except that the delay tag τ is in frequency domain. ■

Generally, the preamble is used for synchronization, but not for directly detecting the message from the payload. Hence the DNC detection of p is $\widehat{p} = \arg\max_q(|y_1(q)|)$. From (11) and (12), we observe that the DFT outputs after dechirping in frequency domain reflect the channel impulse response. The fact that the DFT outputs are cyclic extended by $h(k)$ increases the error probability of estimating the index corresponding to the highest magnitude, so eliminating the influence of the channel is crucial for the acquisition of p.

B. Cyclic Cross-Correlation for Estimating Message

TDE aims at measuring the relative time difference of arrival (TDOA) among spatially separated sensors. As well known, two classical methods are usually utilized to solve the TDE

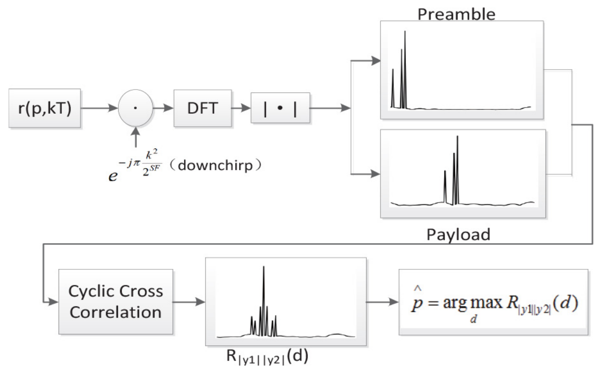

Fig. 1. Block diagram of the proposed receiver algorithm.

problem, i.e., cross-correlation (CC) and generalized cross-correlation (GCC) [13]. Due to the existence of multiple cross-correlation and matrix inversion, GCC-based methods involves high computational complexity and noise amplification. Moreover, $p \in \{0, \ldots, 2^{SF} - 1\}$ is considered as the index in frequency domain locating and the DFT outputs are cyclic extended by $h(k)$. Therefore, we here utilize the cyclic cross-correlation to solve the TDE-liked problem.

Fig. 1 illustrates the block diagram of the proposed receiver algorithm. The DFT outputs of multiple preamble symbols can be averaged to obtain more accurate channel information for better BER performance. The DNC demodulation is executed on the received preamble and payload, respectively. Then, the cyclic cross correlation of $|y_2(q)|$ and $|y_1(q)|$ is calculated. Finally, the message is obtained by selecting the index of the output of the cyclic cross correlation that has the highest magnitude. The cyclic cross-correlation between $|y_1|$ and $|y_2|$ is computed by

$$\begin{aligned} R_{|y_2||y_1|}(d) &= \sum_{q=0}^{2^{SF}-1} |y_2(q)||y_1((q+d)\, mod\ 2^{SF})| \\ &= \sum_{q=0}^{2^{SF}-1} (\widehat{h}(q)\widehat{h}((q-p+d)\, mod\ 2^{SF}) + \widehat{h}(q)C_1 \\ &\quad + \widehat{h}((q-p+d)\, mod\ 2^{SF})C_2 + C_1C_2) \end{aligned} \tag{14}$$

Then, the estimation of p equals to the value of d maximizing $R_{|y_2||y_1|}(d)$, i.e.,

$$\widehat{p} = \arg\max_d R_{|y_2||y_1|}(d) \tag{15}$$

Without loss of generality, the delay spread of $h(k)$ is normally much less than the length of the CSS symbol (2^{SF}). To further reduce the receiver complexity, we can implement the cyclic cross correlation by a matched filter (MF). We only keep the preamble DFT outputs whose magnitudes are greater than a threshold for the matched filtering, as the following

$$|\overline{y_2}| = \begin{cases} |y_2(q)| & |y_2(q)| \geq \alpha |y_2|_{max} \\ 0 & |y_2(q)| < \alpha |y_2|_{max} \end{cases} \tag{16}$$

where $|y_2|_{max}$ is the maximum in $|y_2(q)|$, $\alpha \in (0,1)$ is the threshold parameter which selects the DFT outputs having relatively large gain for the cyclic cross-correlation. In other words, the multipath channel spreads the DFT outputs and thus decreases the effective SNR of the p-th output supposed to have the maximum magnitude. The matched filter can enhance the effective SNR. The proposed receiver algorithm is very simple because it just add an extra cyclic cross correlation implemented by a matched filter to the existing direct non-coherent receiver.

It is worth noting that the TDE-liked detection algorithm is insensitive to the integer frequency offset. Assuming it exists an integer offset $\Delta f_I = \varepsilon_I B/2^{SF}$ in the LoRa system, where $\varepsilon_I \in \mathbb{Z}$. And ε_I will cause the shift of the DFT outputs in frequency domain. Based on the derivation of (11) and (12), we can respectively write the preamble and payload after the direct non-coherent demodulation as

$$\begin{cases} \widehat{|y_1(q)|} \triangleq \widehat{h}(q-p-\varepsilon_I) + \widehat{C}_1 \\ \widehat{|y_2(q)|} \triangleq \widehat{h}(q-\varepsilon_I) + \widehat{C}_2 \end{cases} \tag{17}$$

The cyclic cross-correlation between $\widehat{|y_1(q)|}$ and $\widehat{|y_2(q)|}$ is equivalent to executing a differential operation in frequency domain, which can remove the effect of ε_I. However, the TDE-liked algorithm is still sensitive to the fractional frequency offset. It needs to estimate and compensate the fractional frequency offset for utilizing the TDE-liked algorithm.

IV. Simulation Results

This section provides the simulation results for comparing the BER performance of the proposed TDE-liked algorithm, the direct non-coherent detection algorithm [5]–[8] and the frequency domain equalization (FDE) [10] algorithm, which respectively are denoted by "MF", "DNC" and "FDE" in the following. In the FDE algorithm, a frequency domain equalizer [10] is used before the direct non-coherent detection. In the simulations, the channel estimation is averaged over 8 preamble symbols, $\alpha = \frac{1}{4}$ is set for the proposed algorithm, and the real channel response is used for the frequency equalization in "FDE".

Fig. 2 presents the BER performance comparisons for the three algorithms with $B = 125KHz$, $SF = 7$ and $SF = 10$ over the 2-path channel with impulse response $h(k) = \sqrt{0.8}\delta(k) + \sqrt{0.2}\delta(k-1)$ [5] and $SF = 7$ over the ITU Vehicular B (VehB) channel [14]. The power profile of the VehB channel is [-2.5 0 -128 -100 -25.2 -16] in dB with a relatively delay profile [0 300 8900 12900 17100 20000] in ns. We observe that "DNC" performs almost 3dB worse under the 2-path channel than AWGN channel, because the power of the DFT outputs is split to two positions in frequency domain due to the 2-path fading channel. The proposed algorithm exhibits about 2 dB gain compared to "DNC" and "FDE", in the case of $SF = 7$ and $BER = 10^{-3}$. With the higher $SF = 10$, the performance gains over "DNC" and "FDE" are 1dB and 1.8dB, respectively. The TDE-liked algorithm with $SF = 7$ under the VehB channel shows significant gains compared to the existing detection algorithms. We can see that, with the same SF the proposed algorithm under the VehB channel shows more gains compared to the 2-path channel. Under the 2-path and the VehB channels, the proposed algorithm is slightly inferior to that in the ideal AWGN channel.

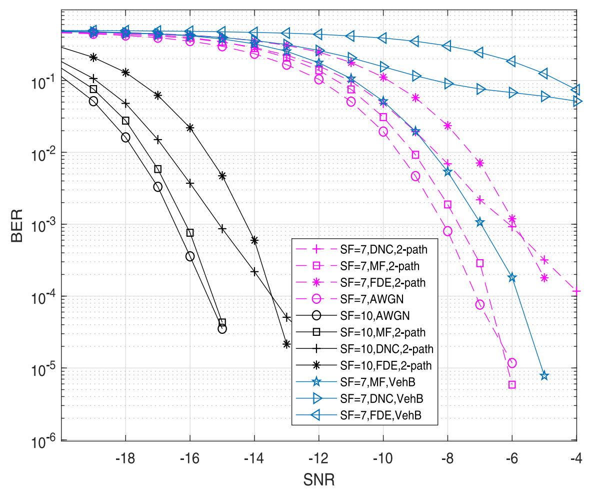

Fig. 2. BER performance comparisons with $B = 125KHz$, $SF = 7$ and $SF = 10$ over 2-path fixed channel and $SF = 7$ over the VehB channel.

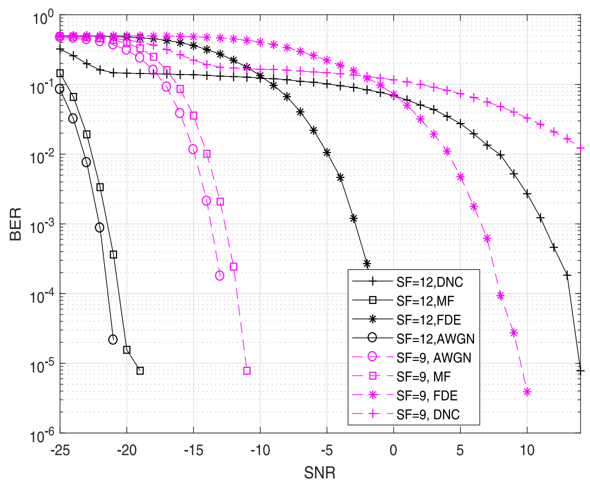

Fig. 3. BER performance comparison of different detection algorithm over VehB channel with $SF = 9$ and $SF = 12$, $B = 500KHz$.

In Fig. 3, the BER performance curves for the three algorithms with $SF = 9$ and $SF = 12$ under the VehB channel model are provided. "DNC" algorithm fails to work under such severely fading channel at the negative SNR regime even with the most robust $SF = 12$. "FDE" can compensate the channel fading to some extent and thus outperform "DNC", but it is still affected by noise amplification. The BER performance of the TDE-liked algorithm under the VehB channel shows around 20dB gain compared to "FDE" and is only 0.5dB inferior to that in the ideal AWGN channel.

From the above simulation results, the existing detection algorithms are severely affected by the multipath fading channels, thus they cannot ensure the promising reliable communication under such environments. The TDE-liked detection shows an excellent BER performance even under the severe fading channels, at the cost of merely adding an extra matched-filter to the direct non-coherent receiver. These good features demonstrated the proposed TDE-liked detection algorithm stands out as a competitive alternative to the LoRa systems under the multipath channels.

V. CONCLUSION

This letter proposed a TDE-liked detection algorithm for the frame-based LoRa signals over multipath channels. The TDE-liked algorithm is implemented by a simple cyclic cross correlation, and is also demonstrated to be insensitive to the integer frequency offset. Through the simulations, the proposed algorithm significantly outperforms the existing receiver algorithms under multipath channels. The BER performance of the proposed algorithm with larger SF under some severely fading channel is only 0.5dB different from that of AWGN channel. The proposed algorithm can promise a reliable communication with low complexity for LoRa systems over multipath channels.

REFERENCES

[1] F. Adelantado, X. Vilajosana, P. Tuset-Peiro, B. Martinez, J. Melia-Segui, and T. Watteyne, "Understanding the limits of LoRaWAN," *IEEE Commun. Mag.*, vol. 55, no. 9, pp. 34–40, Sep. 2017.

[2] A. Lavric and V. Popa, "Internet of Things and LoRa^{TM} low-power wide-area networks: A survey," in *Proc. Int. Symp. Signals Circuits Syst. (ISSCS)*, Iasi, Romania, 2017, pp. 1–5.

[3] J. Petajajarvi, K. Mikhaylov, A. Roivainen, T. Hanninen, and M. Pettissalo, "On the coverage of LPWANs: Range evaluation and channel attenuation model for LoRa technology," in *Proc. 14th Int. Conf. ITS Telecommun. (ITST)*, Copenhagen, Denmark, 2015, pp. 55–59.

[4] A. Waret, M. Kaneko, A. Guitton, and N. El Rachkidy, "LoRa throughput analysis with imperfect spreading factor orthogonality," *IEEE Wireless Commun. Lett.*, vol. 8, no. 2, pp. 408–411, Apr. 2019.

[5] L. Vangelista, "Frequency shift chirp modulation: The LoRa modulation," *IEEE Signal Process. Lett.*, vol. 24, no. 12, pp. 1818–1821, Dec. 2017.

[6] T. Elshabrawy and J. Robert, "Closed-form approximation of LoRa modulation BER performance," *IEEE Commun. Lett.*, vol. 22, no. 9, pp. 1778–1781, Sep. 2018.

[7] C. Goursaud and J. M. Gorce, "Dedicated networks for IoT: PHY/MAC state of the art and challenges," *EAI Endorsed Trans. Internet Things*, vol. 15, no. 1, pp. 4–7, Oct. 2015

[8] B. Reynders and S. Pollin, "Chirp spread spectrum as a modulation technique for long range communication," in *Proc. Symp. Commun. Veh. Technol. (SCVT)*, Nov, 2016, pp. 1–5.

[9] J. C. Liando, A. Gamage, A. W. Tengourtius, and M. Li, "Known and unknown facts of LoRa: Experiences from a large-scale measurement study," *ACM Trans. Sensor Netw.*, vol. 15, no. 2, pp. 16:1–16:35, Feb. 2019.

[10] K. Huang, R. Tao, and Y. Wang, "Study of frequency domain equalization for chirp spread spectrum systems," in *Proc. IEEE Int. Conf. Wireless Commun. Netw. Inf. Security*, Beijing, China, 2010, pp. 132–136.

[11] R. Ghanaatian, O. Afisiadis, M. Cotting, and A. Burg, "Lora digital receiver analysis and implementation," in *Proc. IEEE Int. Conf. Acoust. Speech Signal Process. (ICASSP)*, Brighton, U.K., 2019, pp. 1498–1502

[12] T. Elshabrawy and J. Robert, "Interleaved chirp spreading LoRa-based modulation," *IEEE Internet Things J.*, vol. 6, no. 2, pp. 3855–3863, Apr. 2019.

[13] C. Knapp and G. Carter, "The generalized correlation method for estimation of time delay," *IEEE Trans. Acoust. Speech Signal Process.*, vol. 24, no. 4, pp. 320–327, Aug. 1976.

[14] D. Lekomtcev, E. Kasem, and R. Marsalek, "MATLAB-based simulator of cooperative spectrum sening in real channel condition," in *Proc. 25th Int. Conf. Radioelektronika (RADIOELEKTRONIKA)*, Pardubice, Czech Republic, 2015, pp. 209–212.