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Complexity-Efficient Sidelink Synchronization Signal Detection Scheme for Cellular Vehicle-to-Everything Communication Systems

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Abstract: Synchronization is a challenging issue in vehicle-to-everything (V2X) cellular communication, especially when V2X devices need to directly communicate with each other outside the network coverage area. By adopting the maximum likelihood principle, we propose joint detection of the sidelink secondary synchronization signal (SL-SSS) and carrier frequency offset (CFO) in a V2X system using 5G new radio sidelink. We formulate an effective joint coherent synchronization scheme for cellular V2X applications by decoupling the estimation of the sidelink identity and CFO, which requires a priori knowledge of channel state information. To verify the feasibility of the proposed detection scheme, we derive a simplified implementation of the proposed SL-SSS detector and a closed-form expression for the detection probability. Simulation results demonstrate that the proposed SL-SSS detector exhibits either comparable performance in terms of detection probability or reduced complexity when compared with conventional SL-SSS detectors. Using the proposed method in the cellular V2X system enables V2X devices to achieve robust synchronization with reduced power consumption during the initial synchronization procedure, while also offering valuable insights for designing a simple, efficient sidelink synchronization receiver.

Keywords: 5G new radio; sidelink secondary synchronization signal; maximum likelihood detection; vehicle to everything

MSC: 94A13

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1. Introduction

Device-to-device (D2D) communication is a prime enabler that creates direct interaction between 5G-enabled user devices without the participation of the cellular network [1]. Sidelink D2D communication is considered a key technology in advancing wireless cellular networks from 4G long-term evolution-advanced (LTE-A) to 5G new radio (NR). Standardization developments towards 5G NR in 3rd-generation partnership project (3GPP) releases 16 and 17 focus on feature enhancements for cellular vehicle-to-everything (V2X) communication using NR sidelink [2–4]. The main scenario of NR sidelink addresses D2D-enabled V2X use cases, such as vehicle platooning, advanced driving, extended sensors, and remote driving [5–7]. A main challenge in 5G NR-V2X applications is meeting the rigorous reliability and latency demands while simultaneously enhancing coverage using relays [8]. In order to achieve these goals, releases 16 and 17 establish the role of 5G in supporting advanced V2X services and ensuring vehicle quality of service (QoS) support. Thus, V2X communication is anticipated to play a pivotal role in 5G NR systems [4].

In NR-V2X cellular communication, synchronization information can be periodically broadcasted with source user equipment (UE) to increase the synchronization coverage of

a synchronization source and to facilitate multiple UE devices to align the time reference with each other [9,10]. When UE devices are in different, non-synchronized cells, or one or more UE devices are out of coverage, UE devices have to synchronize with one another via sidelink synchronization signal (SLSS) [11], which is transmitted within a sidelink synchronization signal block (SL-SSB). Figure 1 illustrates the synchronization modes of V2X networks. As shown in Figure 1, achieving synchronization across a network, especially in situations with partial coverage and in out-of-coverage scenarios where V2X UE devices communicate directly among themselves, poses a big challenge. In each SL-SSB, a sidelink primary synchronization signal (SL-PSS) and a sidelink secondary synchronization signal (SL-SSS) together carry information such as the source identifier of the transmitting UE, as well as synchronization information. At the beginning of communication, the SLSS is used for timing- and frequency-offset estimation. By identifying the SLSS transmitted by a source UE device, nearby UE devices can be synchronized with the source UE device and can estimate the symbol timing offset (STO) as well as carrier frequency offset (CFO) [12]. Once the initial STO and CFO have been removed from the time domain, a UE device attempts to acquire a sidelink identity by decoding the sidelink synchronization identity (SSID) information transmitted by the SLSS, which includes both the in-coverage indicator and the SLSS identifier (SLSSID) [13]. The problem of detecting the SLSS can be primarily divided into two main challenges. Firstly, the V2X UE device lacks information about the system timing, and secondly, the frequency of the local oscillator is not fully synchronized with the network [14]. Consequently, significant time and frequency uncertainties may be present between the V2X UE and the network, posing a substantial challenge to initial sidelink acquisition [15]. To establish robust synchronization under such demanding conditions, various synchronization strategies have been investigated in the existing literature [14–18].

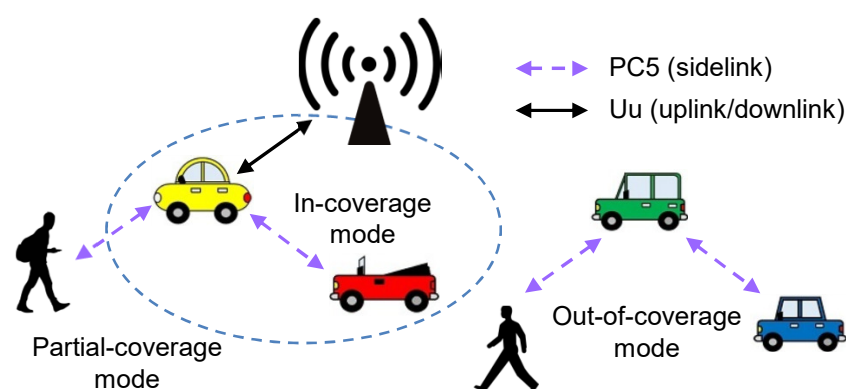


Figure 1. Synchronization modes of V2X cellular networks.

The first step of the synchronization procedure on the UE side is the detection of the SL-PSS and the determination of the in-coverage indicator transmitted in the SL-PSS. After SL-PSS detection, the UE device attempts to detect the SLSSID transmitted in the SL-SSS [16]. Generally, there are two main strategies for detecting the SL-SSS, which can be performed in a coherent or non-coherent manner [17–19]. Once the SL-PSS has been identified, the UE device becomes ready for obtaining channel state information (CSI) to coherently detect the following SL-SSS. However, an interfering SL-PSS received from a neighboring next-generation nodeB (gNodeB) may deteriorate the performance of the initial CSI estimation. Due to the difficult task of CSI estimation, differential detection is commonly used for non-coherent SL-SSS detection (NSD), thereby avoiding any CSI estimation [19]. In this way, the effect of multipath fading and timing offset is removed from the received SL-SSS without resorting to the use of CSI. In spite of its simple implementation and robustness to Doppler frequency, the performance of the NSD approach heavily depends on the amount of frequency selectivity over the signal bandwidth. Since the SL-PSS is transmitted on two consecutive orthogonal frequency division multiplexing (OFDM) symbols, average channel estimation of the SL-PSS becomes feasible and thus provides

an improved CSI estimate. Using available CSI, various coherent SL-SSS detection (CSD) schemes have been proposed in the literature [17,18]. However, in situations where the time selectivity of the channel prevents accurate CSI estimation and the Doppler effect can be high enough, the performance of the CSD methods deteriorates accordingly. In [18], the two binary m-sequences comprising a SL-SSS can be detected independently without relying on mutual information, which contributes to simplifying the complexity of the SL-SSS detector. Nevertheless, this computational advantage comes at the expense of sacrificing detection performance. After successfully detecting SL-SSS, the UE device gains access to important system information transmitted via a physical broadcast channel (PBCH). To decode PBCH successfully, it is essential to track and eliminate the remaining residual CFO (RCFO) since it persists even after CFO compensation. For this reason, it is important to design a cost-effective and still high-performance SL-SSS detection method for 5G V2X vehicular communications. Additionally, the initial synchronization procedure in 5G NR-V2X systems becomes notably challenging due to the necessity to evaluate 336 potential SL-SSS candidates for detection.

This paper presents a reduced-complexity formulation for SL-SSS detection in cellular V2X communication systems, supported by a coherent algorithm based on a priori knowledge of the CSI. The main contributions of this paper are stated as follows.

- By utilizing the cyclic-shifted property of SL-SSS derived from m-sequence and adopting the maximum likelihood (ML) detection principle, we design a computationally efficient SL-SSS detection method to simultaneously estimate the SLSSID and RCFO in a decoupled manner without sacrificing detection performance.
- The benefit of employing the identical correlation function for SLSSID detection is the capability to concurrently estimate the RCFO, thus resulting in a more straightforward design for the synchronization receiver.
- Such a design makes it easier to achieve robust joint detection of the SLSSID and RCFO in the presence of significant frequency-selective fading and STO.
- In order to validate the viability of the proposed SL-SSS detection method, we compare the performance of the proposed and conventional SL-SSS detectors in terms of detection probability and computational complexity.
- Simulation results confirm that the proposed SL-SSS detection method is computationally efficient and achieves detection performance comparable with that of existing approaches, thereby making it as a strong candidate for synchronization receivers in 5G NR-V2X systems.

The reminder of this paper is as follows. Section 2 presents the system model for 5G NR-V2X communication. Section 3 highlights the reduced-complexity SL-SSS synchronization method. The numerical results, demonstrating the benefits of the proposed detector, are presented in Section 4. Finally, the paper is summarized in Section 5.

2. System Model

2.1. Signal Model

We adopt an OFDM system which has N equi-width subcarriers. A time-domain OFDM symbol with a duration of T is generated through the application of an N -point inverse discrete Fourier transform (IDFT) to the complex data symbol $X_l(v)$, $v = 0, 1, \dots, N - 1$. In order to preserve inter-symbol interference free transmission, a guard interval (GI) of duration T_g is appended to the useful part of the OFDM signal, leading to blocks of length $T_u = T + T_g$. For a given symbol index l , thus, the time-domain modulated signal with duration $T_u = (N + N_g)T_s$ is described as

$$x_l(n) = \sum_{v=0}^{N-1} X_l(v) e^{j2\pi v n / N} \quad (1)$$

where T_s is the amount of time between consecutive output samples, N_g is the number of cyclic prefix (CP) samples used for GI insertion, $n = -N_g, -N_g + 1, \dots, N - 1$ is the time sample index, and $X_l(v)$ is the complex-valued data symbol.

In this paper, we consider the situation in which there remain an uncompensated integer CFO (ICFO), RCFO, and residual STO (RSTO) after the initial synchronization process in the pre-DFT stage [20]. As reported in [18], the RSTO has little impact on the detection performance during the post-DFT synchronization stage. Supposing that the RSTO is known to the receiver, the transmit signal $x_l(n)$ linearly convolves with the multi-path fading channel characterized by channel impulse response $h_l(n)$. In the time domain, then, the received signal during the l -th period is represented as

$$y_l(n) = e^{j2\pi l(\eta+\varepsilon)N_s/N} e^{j2\pi(\eta+\varepsilon)n/N} h_l(n) \otimes x_l(n) + z_l(n), \quad n = -N_g, -N_g + 1, \dots, N - 1 \quad (2)$$

where η is the normalized ICFO with respect to the frequency spacing between adjacent subcarriers denoted by Δ_f , ε is the normalized RCFO with respect to Δ_f , $N_s = N + N_g$, $\otimes$ is the linear convolution operator, and $z_l(n)$ is the zero-mean additive white Gaussian noise (AWGN) with variance σ_z^2 . Assuming that the GI duration is sufficiently large to absorb the multi-path channel effect, it can be safely discarded at the receiver. After the DFT process is performed over the CP-removed signal, the demodulated OFDM symbol under the effect of normalized CFO turns into [21]

$$Y_l(v) \approx H_l(v - \eta) X_l(v - \eta) e^{j2\pi \zeta (lN_s + N_g)/N} + I_l(v) + Z_l(v) \quad (3)$$

where l denotes the symbol index, $\zeta = \eta + \varepsilon$, $H_l(v)$ denotes the CSI following the complex Gaussian distribution $\mathcal{CN}(0, \sigma_H^2)$, $I_l(v)$ denotes the inter-channel interference (ICI) term having variance σ_I^2 , and $Z_l(v)$ denotes the AWGN term having $Z_l(v) \sim \mathcal{CN}(0, \sigma_z^2)$. The notation $x \sim \mathcal{CN}(y, z)$ indicates that a complex random variable x follows a complex normal distribution with mean y and variance z .

2.2. Synchronization Signal

The NR-V2X supports 672 physical-layer SSIDs. Each SSID is generated by uniquely combining 1 SL-PSS and 1 SL-SSS out of the 2 SL-PSS and 336 SL-SSS possible candidates, respectively. As a consequence, each SSID is a unique number assigned according to $N_{SSID} = s + 336u$, where $u \in \{0, 1\}$ and $s \in \{0, 1, 2, \dots, 335\}$ stand for identifiers of the SL-PSS and SL-SSS out of the SL-PSS and SL-SSS possible candidates, respectively. Depending on the in-coverage or out-of-coverage statuses of the UE devices, the number of possible SSIDs is partitioned into two groups. For the in-coverage case ($u = 0$), the UE device periodically sends SL-PSS sequences indicated by $N_{SSID} \leq 335$. For the out-of-coverage scenarios ($u = 1$), on the other hand, the UE device transmits SL-PSS sequences indicated by $N_{SSID} > 335$. In order to provide higher coverage, the SL-PSS and SL-SSS are both broadcasted twice consecutively within each SL-SSB.

In the frequency domain, the SL-PSS is generated by a maximum length sequence (m-sequence) with a length of 127 points, thus occupying 127 subcarriers of the SL-SSB bandwidth. By cyclically shifting a basic m-sequence $c(n')$, one obtain two unique SL-PSS symbols as follows:

$$m_u(n') = 1 - 2c([n' + 22 + 43u] \bmod N_m), \quad 0 \leq n' < N_m \quad (4)$$

where $u \in \{0, 1\}$ serves as an identifier of the SL-PSS among the candidate sequences and $N_m = 127$ denotes the length of SL-PSS symbol. Let us denote $M_u(v)$ as the SL-PSS symbol transmitted at subcarrier v , which is defined as

$$M_u(v) = \begin{cases} m_u(v - 2), & 2 \leq v \leq 128 \\ 0, & \text{otherwise} \end{cases} \quad (5)$$

where the same SL-PSS sequence is allocated to the second and third OFDM symbols of each SSB. The SL-SSS is generated by concatenating two m-sequences to overcome poor cross-correlation issue, which forms a gold sequence with a length of also 127 points. The generation of the SL-SSS is given by

$$g_s(n') = [1 - 2c_0([n' + 15[s/112] + 5u] \bmod N_m)] \times [1 - 2c_1([n' + (s \bmod 112)] \bmod N_m)], 0 \leq n' < N_m \quad (6)$$

where $s \in \{0, 1, 2, \dots, 335\}$ serves as an identifier of the SL-SSS among the candidate sequences and $\lfloor \cdot \rfloor$ stands for the floor operation. The formulation of two sequences $c_0(n')$ and $c_1(n')$ is described in detail in [3]. Identically, $g_s(n')$ is allocated onto N_m SL-SSS subcarriers with $\{2 \leq v \leq 128\}$, producing $G_s(v)$ as follows:

$$G_s(v) = \begin{cases} g_s(v-2), & 2 \leq v \leq 128 \\ 0, & \text{otherwise} \end{cases} \quad (7)$$

where the same SL-SSS sequence is mapped to the fourth and fifth OFDM symbols of each SSB. This paper focuses on the SL-SSS synchronization in NR-V2X vehicular systems.

2.3. SL-SSS Detection

In the literature, SL-SSS detection systems are classified into two categories: coherent detection and non-coherent detection [17–19]. First, the CSD method uses the CSI derived from SL-PSS. In the CSD approach, the cost function is described as [17]

$$\Omega_{csd}(c) = \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \tilde{Y}_{l+i}^*(v + \eta) Y_{l+i+2}(v + \eta) G_c^*(v) \quad (8)$$

where $c \in \{0, 1, 2, \dots, 335\}$ is the available trial value of s , $\mathcal{G} = \{v | 2 \leq v \leq 128\}$ with $N_m = 127$ elements, and $\tilde{Y}_{l+i}(v + \eta) = Y_{l+i}(v + \eta) P_u^*(v)$ is the received signal from which the SL-PSS sequence is compensated during the second and third OFDM symbol periods of each SSB. On the other hand, using the differential correlation between adjacent subcarriers is one of the most popular NSD strategies. In this case, the cost function is given by [19]

$$\Omega_{nsd}(c) = \sum_{i=0}^1 \sum_{v \in \mathcal{D}} Y_{l+i+2}(v + \hat{\eta}) Y_{l+i+2}^*(v + \hat{\eta} - 1) G_c^*(v) G_c(v - 1) \quad (9)$$

where $\mathcal{D} = \{v | 3 \leq v \leq 128\}$ with $N_m - 1$ elements. For both the CSD and NSD methods, the estimate $\hat{s}$ can be found by locating a maximum of $\Omega_{csd}(c)$ or $\Omega_{nsd}(c)$ over 336 candidate values of c . This process requires computationally expensive operations caused by a large number of possible hypotheses. Therefore, the development of a computationally efficient SL-SSS detection method holds significant importance in the 5G NR-V2X system.

3. Proposed SL-SSS Detection Method for V2X Systems

This section presents joint synchronization of SL-SSS and RCFO, which is based on the ML detection principle. In order to evaluate the usefulness of the proposed SL-SSS detector, we develop a simplified alternative and calculate its corresponding analytical detection probability.

For a simple presentation, assume that the SL-PSS with an in-coverage indicator u is present on the l -th OFDM symbol within each SSB such that $M_u(v) = X_l(v) = X_{l+1}(v)$ and $G_s(v) = X_{l+2}(v) = X_{l+3}(v)$. To counteract the impacts of interference and fading channel, we use the average correlation among the received four consecutive SLSSs. As seen from (6), the SL-SSS can be represented by the concatenation of two m-sequences as follows: $G_s(v) = G_{0,s_0}(v) G_{1,s_1}(v)$ for $s_0 = \lfloor s/112 \rfloor \in \{0, 1, 2\}$ and $s_1 = (s \bmod 112) \in \{0, 1, 2, \dots, 111\}$, where $G_{0,s_0}(v)$ and $G_{1,s_1}(v)$ stand for m-sequences mapped from $c_0(n')$

and $c_1(n')$ in a subcarrier-wise manner, respectively. Based on in-coverage indicator u detected using the SL-PSS, $G_{0,s_0}(v)$ forms one of three m-sequences created by applying a cyclic shift of length $15s_0 + 5u$ to a basic code and $G_{1,s_1}(v)$ is designed with the same sequence for all three sequences $G_{0,s_0}(v)$'s ($s_0 = 0, 1, 2$), therefore generating a total number of 336 SLSSIDs. According to this formulation, an SLSSID s is associated with a pair (s_0, s_1) in such a way that $s = s_1 + 112s_0$.

3.1. Approximate ML Detection Scheme

In this section, we consider that the OFDM receiver collects four observation SLSS vectors of length N_m , $\mathbf{Y}_l, \mathbf{Y}_{l+1}, \mathbf{Y}_{l+2}, \mathbf{Y}_{l+3}$, to detect the SLSSID and RCFO. Denote $\mathbf{Y}_{l+d} = [Y_{l+d}(2), \dots, Y_{l+d}(128)]^T$ ($d = 0, 1, 2, 3$). For the simplicity of notations, let us consider a situation where the ICFO η and in-coverage indicator u are perfectly known at the UE receiver. Furthermore, $G_{a,b}(v) = G_{0,a}(v)G_{1,b}(v)$, where $a \in \{0, 1, 2\}$ is the possible identifier of s_0 and $b \in \{0, 1, 2, \dots, 111\}$ is the possible identifier of s_1 . Applying the ML estimation principle, one can obtain the conditional joint probability density function (PDF) of the SLSS vectors $\mathbf{Y}_{l+d}$ ($d = 0, 1, 2, 3$) as

$$f(\mathbf{Y}_l, \mathbf{Y}_{l+1}, \mathbf{Y}_{l+2}, \mathbf{Y}_{l+3} | a, b, \varepsilon) = \frac{1}{(2\pi\sigma^2)^{2N_m}} \exp \left\{ -\frac{1}{2\sigma^2} \times \sum_{i=0}^3 \sum_{v \in \mathcal{G}} \left| \tilde{Y}_{l+i+2}^{a,b}(v + \eta) - e^{j4\pi(\eta+\varepsilon)\varrho} \tilde{Y}_{l+i}(v + \eta) \right|^2 \right\} \quad (10)$$

where σ^2 stands for the appropriate noise variance, $\varrho = (N + N_g)/N$, and $\tilde{Y}_{l+i+2}^{a,b}(v + \eta) = Y_{l+i+2}(v + \eta)G_{a,b}^*(v)$ stands for the received signal from which the SL-SSS sequence is compensated during the fourth and fifth OFDM symbol periods of each SSB.

The ML optimization problem is to jointly choose the SLSSID and RCFO in order to maximize the conditional PDF $f(\mathbf{Y}_l, \mathbf{Y}_{l+1}, \mathbf{Y}_{l+2}, \mathbf{Y}_{l+3} | a, b, \varepsilon)$ with respect to (a, b, ε) , which is described as

$$(\hat{s}_0, \hat{s}_1, \hat{\varepsilon}) = \arg \max_{(a, b, \varepsilon)} f(\mathbf{Y}_l, \mathbf{Y}_{l+1}, \mathbf{Y}_{l+2}, \mathbf{Y}_{l+3} | a, b, \varepsilon). \quad (11)$$

Since $M_u(v)$ has already been detected before the process of detecting SL-SSS, the UE device tries to jointly detect the SLSSID and RCFO by performing

$$(\hat{s}_0, \hat{s}_1, \hat{\varepsilon}) = \arg \max_{(a, b, \varepsilon)} \Psi(a, b, \varepsilon) \quad (12)$$

where

$$\Psi(a, b, \varepsilon) = - \sum_{i=0}^3 \sum_{v \in \mathcal{G}} \left| \tilde{Y}_{l+i+2}^{a,b}(v + \eta) - e^{j4\pi(\eta+\varepsilon)\varrho} \tilde{Y}_{l+i}(v + \eta) \right|^2. \quad (13)$$

After some arithmetic calculations, (13) is simplified as

$$\Psi(a, b, \varepsilon) = 2\Re \left\{ \Omega(a, b) e^{-j4\pi\varepsilon\varrho} \right\} - \Xi(a, b) \quad (14)$$

where $\Re\{\cdot\}$ stands for a real part of the enclosed quantity,

$$\Omega(a, b) = \sum_{i=0}^3 \sum_{v \in \mathcal{G}} \tilde{Y}_{l+i}^*(v + \eta) Y_{l+i+2}(v + \eta) G_{a,b}^*(v) e^{-j4\pi\eta\varrho}, \quad (15)$$

and

$$\Xi(a, b) = \sum_{i=0}^3 \sum_{v \in \mathcal{G}} \left(|\tilde{Y}_{l+i}(v + \eta)|^2 + |\tilde{Y}_{l+i+2}^{a,b}(v + \eta)|^2 \right). \quad (16)$$

Since $\Xi(a, b)$ is independent of both a and b , it can be dropped from the optimization task in (12), resulting in an approximate ML (AML) estimate of (s_0, s_1, ε) simply as

$$(\hat{s}_0, \hat{s}_1, \hat{\varepsilon}) = \arg \max_{(a, b, \varepsilon)} \Re \left\{ \Omega(a, b) e^{-j4\pi\varepsilon\varrho} \right\}. \quad (17)$$

For each trial value of (a, b) , the AML estimator performs an exhaustive search of ε by appropriately quantizing the possible RCFO values, which incurs a heavy computational load. In order to solve the complexity issue involved with maximizing (17) with respect to (s_0, s_1, ε) , a low-complexity alternative is presented in the following section.

3.2. Low-Complexity SL-SSS Detection Scheme

Since the superscripts (a, b) are omitted from $\Xi(a, b)$ and it is apparent from (13) that $\Psi(a, b, \varepsilon)$ is negative, the cost function $\Psi(a, b, \varepsilon)$ reaches its maximum when $\Re \{ \Omega(a, b) e^{-j4\pi\varepsilon\varrho} \}$ is maximized. To achieve this goal, it is necessary that $\angle \{ \Omega(a, b) e^{-j4\pi\varepsilon\varrho} \} = \angle \{ \Omega(a, b) \} - 4\pi\varepsilon\varrho = 0$. Based on this implication, ε can be estimated in a closed form as follows:

$$\hat{\varepsilon} = \frac{1}{4\pi\varrho} \angle \{ \Omega(a, b) \} \quad (18)$$

where $\angle$ is the angle operation. Note that the estimation range of (18) is limited by $|\varepsilon| \leq 1/(4\varrho)$. Using the fact that $e^{-j4\pi\hat{\varepsilon}\varrho} = \Omega^*(a, b)/|\Omega(a, b)|$ and putting this notation in $\Re \{ \Omega(a, b) e^{-j4\pi\hat{\varepsilon}\varrho} \}$, it is obvious that the quantity $\Re \{ \Omega(a, b) e^{-j4\pi\hat{\varepsilon}\varrho} \}$ is reduced to $|\Omega(a, b)|$. By using this formulation, a simplified estimation $(\hat{s}_0, \hat{s}_1)$ is performed as

$$(\hat{s}_0, \hat{s}_1) = \arg \max_{(a, b)} |\Omega(a, b)| \quad (19)$$

which implies that the detection problem of (17) is decoupled in a sequential manner. By collecting the estimated pair $(\hat{s}_0, \hat{s}_1)$, we can determine $\hat{s}$ as $\hat{s} = \hat{s}_1 + 112\hat{s}_0$. Then, one can estimate ε in a decoupled manner by substituting $(\hat{s}_0, \hat{s}_1)$ into (18).

As seen from (19), we perform a hypothesis test for potential SL-SSS candidates to decide one of the 336 SLSSIDs, which still remains computationally very demanding. To address this complexity problem, we propose a low-complexity SLSSID detection method without compromising the detection performance of (19). A close looking at (6) informs us that $G_{0,1}(v)$ and $G_{0,2}(v)$ are generated by cyclically shifting $G_{0,0}(v)$. Therefore, three m-sequences are related to each other as $G_{0,s_0}(v) = G_{0,0}(v)B_{s_0}(v)$, where $B_{s_0}(v) = G_{0,s_0}(v)/G_{0,0}(v) \in \{-1, 1\}$ for $s_0 = 0, 1, 2$ is the phase difference factor between $G_{0,0}(v)$ and $G_{0,s_0}(v)$'s for $s_0 = 1, 2$ that are cyclic-shifted versions of $G_{0,0}(v)$. Inspired by this observation, the cost function is rewritten by

$$\Omega(a, b) = \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \tilde{Y}_{l+i}^*(v + \eta) Y_{l+i+2}(v + \eta) \bar{G}_{0,b}^*(v) B_a^*(v) e^{-j4\pi\eta\varrho} \quad (20)$$

where $\bar{G}_{0,b}(v) = G_{0,0}(v)G_{1,b}(v)$ is the partial SL-SSS for $a = 0$ and $b \in \{0, 1, 2, \dots, 111\}$ whose possible number of candidate sequences is 112. Using the cyclic-shifted property of m-sequence, two different auto-correlations with local template $G_{a,b}(v)$ for the remaining hypothesized values $a \in \{1, 2\}$ and $b \in \{0, 1, 2, \dots, 111\}$ can be implemented by simply compensating only the polarity of the pre-computed autocorrelation for $a = 0$. This compensation is performed on a per-subcarrier basis according to the phase difference factor $B_a(v)$ ($a = 1, 2$). To complete (20), therefore, only one auto-correlation needs to be calculated for each trial value of b , which helps reduce the computational burden by

minimizing the number of required multiplication operations. Substituting (3) into (20), we have

$$\begin{aligned}\Omega(a, b) = & \sum_{i=0}^1 \sum_{v \in \mathcal{G}} |H_{l+i}(v)|^2 |M_u(v)|^2 G_s(v) \bar{G}_{0,b}^*(v) B_a^*(v) e^{j4\pi\epsilon\varrho} \\ & + \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \bar{I}_{l+i}(v) G_{a,b}^*(v) + \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \bar{Z}_{l+i}(v) G_{a,b}^*(v)\end{aligned}\quad (21)$$

where $\tilde{\Omega}(a, b) = \Omega(a, b)e^{-j4\pi\epsilon\varrho}$ is a complex Gaussian random variable (RV),

$$\begin{aligned}\bar{I}_{l+i}(v) = & H_{l+i}^*(v) |M_u(v)|^2 I_{l+i+2}(v) e^{-j2\pi\zeta((l+i)N_s + N_g)/N} \\ & + H_{l+i+2}(v) G_s(v) M_u(v) I_{l+i}^*(v) e^{j2\pi\zeta((l+i+2)N_s + N_g)/N} + M_u(v) Z_{l+i}^*(v) I_{l+i+2}(v) \\ & + M_u(v) Z_{l+i+2}(v) I_{l+i}^*(v) + M_u(v) I_{l+i}^*(v) I_{l+i+2}(v)\end{aligned}\quad (22)$$

and

$$\begin{aligned}\bar{Z}_{l+i}(v) = & H_{l+i}^*(v) |M_u(v)|^2 Z_{l+i+2}(v) e^{-j2\pi\zeta((l+i)N_s + N_g)/N} \\ & + H_{l+i+2}(v) G_s(v) M_u(v) Z_{l+i}^*(v) e^{j2\pi\zeta((l+i+2)N_s + N_g)/N} + M_u(v) Z_{l+i}^*(v) Z_{l+i+2}(v).\end{aligned}\quad (23)$$

Note that $\bar{I}_{l+i}(v)$ and $\bar{Z}_{l+i}(v)$ are treated as zero-mean complex Gaussian RVs with variances σ_I^2 and σ_Z^2 , respectively. Assuming $g = b + 112a$ (hypothesis H_1), substituting (3) into (20) yields

$$\begin{aligned}\Omega(a, b) = & \sum_{i=0}^1 \sum_{v \in \mathcal{G}} |H_{l+i}(v)|^2 |M_u(v)|^2 |G_s(v)|^2 e^{j4\pi\epsilon\varrho} \\ & + \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \bar{I}_{l+i}(v) G_{a,b}^*(v) + \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \bar{Z}_{l+i}(v) G_{a,b}^*(v)\end{aligned}\quad (24)$$

where we assume $H_l(v) \approx H_{l+1}(v)$.

3.3. Performance Analysis

3.3.1. SL-SSS Detection

In this subsection, the detection probability of (19) is derived in a closed form. For ease of derivation, we suppose that the flat-fading channel is slowly time-variant, i.e., $H_l(v) \approx H_{l+1}(v)$. Let $\mathbb{E}\{\cdot\}$ denote the expected value of the enclosed quantity. The detection probability $P_d = \text{Prob}\{(\hat{s}_0, \hat{s}_1) = (s_0, s_1)\}$ is the probability that (s_0, s_1) (or equivalently s) are correctly decided. First, we consider the hypothesis H_1 that $(a, b) = (s_0, s_1)$ in (24). Conditioned on $\beta = |H_l(v)|^2$, we find that $\tilde{\Omega}(a, b) = \Omega(a, b)e^{-j4\pi\epsilon\varrho} \sim \mathcal{CN}(\mu, \sigma_1^2)$ with $\mu = 2N_m E_X^2 \beta$ and $\sigma_1^2 = 2N_m E_X (\sigma_I^2 + \sigma_Z^2)$, where $E_X = \mathbb{E}\{|M_u(v)|^2\} = \mathbb{E}\{|G_s(v)|^2\}$, $\sigma_I^2 = 2E_X^2 \sigma_I^2 \beta + 2E_X \sigma_I^2 \sigma_Z^2 + E_X \sigma_I^4$, $\sigma_Z^2 = 2E_X^2 \sigma_Z^2 \beta + E_X \sigma_Z^4$, and $\sigma_I^2 \approx E_X \sigma_H^2 \epsilon^2 \pi^2 / 3$. In this assumption, the PDF of $z = |\tilde{\Omega}(a, b)| = |\Omega(a, b)|$ in (24) is Rician distributed and is given by

$$f_{H_1}(z) = \frac{2z}{\sigma_1^2} e^{-\frac{z^2 + \mu^2}{\sigma_1^2}} I_0\left(\frac{2z\mu}{\sigma_1^2}\right)\quad (25)$$

where $I_0(\cdot)$ denotes the modified Bessel function of the first kind and zeroth order. In the case when $(a, b) \neq (s_0, s_1)$ (null hypothesis H_0), $\tilde{\Omega}(a, b) \sim \mathcal{CN}(0, \sigma_0^2)$. For $N_m \gg 1$, it can be seen that $\sigma_0^2 \approx \sigma_1^2$. As a consequence, $|\Omega(a, b)|$ is Rayleigh-distributed as $f_{H_0}(z) = 2z/\sigma_0^2 e^{-z^2/\sigma_0^2}$.

Thus, making use of $f_{H_0}(z)$ and $f_{H_1}(z)$, the probability that (s_0, s_1) are correctly detected and conditioned on β becomes

$$P_d(\beta) = \int_0^\infty f_{H_1}(z) \left[\int_0^z f_{H_0}(y) dy \right]^{335} dz. \quad (26)$$

With the aid of binomial expansion, $P_d(\beta)$ is found to be

$$\begin{aligned} P_d(\beta) &= \sum_{g=0}^{335} (-1)^g \binom{335}{g} \int_0^\infty \frac{2z}{\sigma_1^2} e^{-\frac{z^2(1+g)}{\sigma_1^2} - \nu^2} I_0\left(\frac{2z\mu}{\sigma_1^2}\right) dz \\ &= \sum_{g=0}^{335} (-1)^g \binom{335}{g} \frac{1}{1+g} e^{-\frac{g\nu^2}{1+g}} \end{aligned} \quad (27)$$

where $\nu = \mu/\sigma_1$ is the ratio of mean to standard deviation of a complex random variable $\tilde{\Omega}(a, b)$. In (27), ν is defined as

$$\nu = \sqrt{\frac{2N_m}{2(1/\gamma_z + 1/\gamma_i) + (1/\gamma_z + 1/\gamma_i)^2}} \quad (28)$$

where $\gamma_z = \beta E_X/\sigma_Z^2$ is the instantaneous signal-to-noise ratio (SNR) and $\gamma_i = \beta E_X/\sigma_I^2$ is the instantaneous signal-to-ICI ratio (SIR). In order to find a closed-form formula, (27) is averaged over β , namely

$$\begin{aligned} P_d &= \int_0^\infty P_d(\beta) f(\beta) d\beta \\ &= \sum_{g=0}^{335} (-1)^g \binom{335}{g} \frac{1}{(1+g)\sigma_H^2} \int_0^\infty e^{-\frac{g\nu^2}{1+g}} e^{-\frac{\beta}{\sigma_H^2}} d\beta \end{aligned} \quad (29)$$

where $f(\beta) = \sigma_H^{-2} e^{-\frac{\beta}{\sigma_H^2}}$ is the PDF of β . If the SNR is relatively high, the second squaring term of the denominator of (28) is negligible. Ignoring this term, ν is approximated as

$$\nu \approx \sqrt{\frac{N_m \gamma_z \gamma_i}{\gamma_z + \gamma_i}}. \quad (30)$$

Substituting (30) into (29), and after some mathematical simplifications, one obtains the following unconditional probability

$$P_d = \sum_{g=0}^{335} (-1)^g \binom{335}{g} \frac{\bar{\gamma}_z + \bar{\gamma}_i}{g N_m \bar{\gamma}_z \bar{\gamma}_i + (1+g)(\bar{\gamma}_z + \bar{\gamma}_i)} \quad (31)$$

where $\bar{\gamma}_z = \mathbb{E}\{\beta\} E_X/\sigma_Z^2$ is the average SNR and $\bar{\gamma}_i = \mathbb{E}\{\beta\} E_X/\sigma_I^2$ is the average SIR.

3.3.2. RCFO Estimation

We consider the scenario in which the estimates $(\hat{s}_0, \hat{s}_1)$ are available to the receiver without any error. With this assumption, it is easy to see that instead of (21), we now have

$$\Omega(\hat{s}_0, \hat{s}_1) = 2E_X^2 e^{j4\pi\epsilon\varrho} \sum_{v \in \mathcal{G}} |H_l(v)|^2 \left[1 + \frac{\sum_{v \in \mathcal{G}} \bar{I}(v) + \sum_{v \in \mathcal{G}} \bar{Z}(v)}{2E_X^2 e^{j4\pi\epsilon\varrho} \sum_{v \in \mathcal{G}} |H_l(v)|^2} \right] \quad (32)$$

where $\bar{I}(v) = \sum_{i=0}^1 \bar{I}_{l+i}(v) G_{a,b}^*(v)$ and $\bar{Z}(v) = \sum_{i=0}^1 \bar{Z}_{l+i}(v) G_{a,b}^*(v)$. Assuming a significantly large SNR and applying power boosting to the SLSS, we can confidently assert that $\Re\{\sum_{v \in \mathcal{G}} \bar{I}(v) e^{-j4\pi\epsilon\varrho} + \sum_{v \in \mathcal{G}} \bar{Z}(v) e^{-j4\pi\epsilon\varrho}\} / (2E_X^2 \sum_{v \in \mathcal{G}} |H_l(v)|^2) \ll 1$. Consequently, for high SNR, it follows that

$$\begin{aligned}\angle\Omega(\hat{s}_0, \hat{s}_1) &= 4\pi\epsilon\varrho + \angle \left[1 + \frac{\sum_{v \in \mathcal{G}} \bar{I}(v) + \sum_{v \in \mathcal{G}} \bar{Z}(v)}{2E_X^2 e^{j4\pi\epsilon\varrho} \sum_{v \in \mathcal{G}} |H_l(v)|^2} \right] \\ &\approx 4\pi\epsilon\varrho + \angle e^{j \tan^{-1} \left(\frac{\sum_{v \in \mathcal{G}} \Im\{\bar{I}(v)\} + \sum_{v \in \mathcal{G}} \Im\{\bar{Z}(v)\}}{2E_X^2 \sum_{v \in \mathcal{G}} |H_l(v)|^2} \right)}\end{aligned}\quad (33)$$

where $\hat{I}(v) = \bar{I}(v)e^{-j4\pi\epsilon\varrho}$, $\hat{Z}(v) = \bar{Z}(v)e^{-j4\pi\epsilon\varrho}$, $\Im\{\cdot\}$ represents the imaginary part and $\tan^{-1}(x)$ refers to the arctangent function applied to a real number x . It is important to mention that $\bar{I}(v)$ and $\bar{Z}(v)$ hold the same statistical equivalence as $\hat{I}(v)$ and $\hat{Z}(v)$, respectively. Since the estimated error $\Delta_\epsilon = \hat{\epsilon} - \epsilon$ is significantly small and $\angle e^{j \tan^{-1}(x)}$ approximates $\angle e^{jx} = x$ with increasing SNR, we can easily conclude that

$$\Delta_\epsilon \approx \frac{1}{4\pi\varrho} \frac{\sum_{v \in \mathcal{G}} \Im\{\hat{I}(v)\} + \sum_{v \in \mathcal{G}} \Im\{\hat{Z}(v)\}}{2E_X^2 N_m \beta}. \quad (34)$$

It is clear from (34) that the MSE of the RCFO estimation scheme is represented by

$$\mathbb{E}\{|\Delta_\epsilon|^2\} = \left(\frac{1}{8\pi\varrho N_m E_X^2} \right)^2 \sum_{v \in \mathcal{G}} \mathbb{E}\{|\Im\{\hat{I}(v)\}/\beta|^2\} + \left(\frac{1}{8\pi\varrho N_m E_X^2} \right)^2 \sum_{v \in \mathcal{G}} \mathbb{E}\{|\Im\{\hat{Z}(v)\}/\beta|^2\} \quad (35)$$

and we have $\mathbb{E}\{\Im\{\hat{I}(v)\}\} = \mathbb{E}\{\Im\{\hat{Z}(v)\}\} = 0$. Upon observing (22) and (23), it becomes apparent that the variances of $\hat{I}(v)$ and $\hat{Z}(v)$ are computed as $\sigma_I^2 = 4E_X^3 \sigma_I^2 \beta + 4E_X^2 \sigma_I^2 \sigma_Z^2 + 2E_X^2 \sigma_I^4$, $\sigma_Z^2 = 4E_X^3 \sigma_Z^2 \beta + 2E_X^2 \sigma_Z^4$, respectively. Hence, it is evident that

$$\frac{\mathbb{E}\{|\Im\{\hat{I}(v)\}/\beta|^2\}}{E_X^4} = \frac{4}{\gamma_i} + \frac{4}{\gamma_i \gamma_z} + \frac{2}{\gamma_i^2} \quad (36)$$

and

$$\frac{\mathbb{E}\{|\Im\{\hat{Z}(v)\}/\beta|^2\}}{E_X^4} = \frac{4}{\gamma_z} + \frac{2}{\gamma_z^2}. \quad (37)$$

By combining (35)–(37), we eventually arrive at

$$\mathbb{E}\{|\Delta_\epsilon|^2\} = \frac{1}{16\pi^2 \varrho^2 N_m} \left(\frac{1}{\gamma_i} + \frac{1}{\gamma_i \gamma_z} + \frac{1}{2\gamma_i^2} + \frac{1}{\gamma_z} + \frac{1}{2\gamma_z^2} \right). \quad (38)$$

4. Numerical Results and Analysis

In this section, we conduct extensive simulation experiments to verify the effectiveness of the SL-SSS detectors.

4.1. Simulation Setup

We consider an OFDM system employing the number of DFT points $N = 2048$ and the CP size $N_g = 144$. It operates within a transmission bandwidth of 40 MHz with a frequency separation $\Delta_f = 30$ kHz in the 6 GHz frequency band, resulting in a sampling time period of $T_s = 1/61.44 \mu\text{s}$. For the simulations, a 5G tapped delay line-B model is used for non-line-of-sight scenarios [22]. Two different values of delay spreads are considered to represent two channel profiles, which are very short delay (Vsd) spread and very long delay (Vld) spread. Unless otherwise mentioned, the Doppler frequency is set to be $D_f = 300$ Hz, corresponding to the UE velocity of 54 km/h [23], and all performance results were obtained through simulations.

To demonstrate the advantages of the proposed SL-SSS detection scheme, we consider three different SL-SSS detection schemes as benchmarks [17–19]. The first benchmark is the coherent approach and the second benchmark is the non-coherent method with the cost functions provided in (8) and (9), respectively. The last benchmark is a reduced SL-SSS

detection (RSD) method that identifies the SL-SSS by individually maximizing two cost functions given by [18]

$$\Omega_{rsd1}(b) = \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \tilde{Y}_{l+i}^*(v + \eta) Y_{l+i+2}(v + \eta) \left\{ \sum_{a=0}^2 G_{0,a}^*(v) \right\} G_{1,b}^*(v) e^{-j4\pi\eta\epsilon} \quad (39)$$

and

$$\Omega_{rsd0}(a) = \sum_{i=0}^1 \sum_{v \in \mathcal{G}} \tilde{Y}_{l+i}^*(v + \eta) Y_{l+i+2}(v + \eta) G_{0,a}^*(v) G_{1,\hat{b}}^*(v) e^{-j4\pi\eta\epsilon}. \quad (40)$$

4.2. Complexity Evaluation

First, we evaluate the computational load of the SL-SSS detectors, assuming six floating point operations (flops) for each complex multiplication, two flops for each complex addition, and three flops for each complex magnitude [24]. For $\mathcal{G}$ and $i = 0, 1$, the CSD method computes $\tilde{Y}_{l+i}^*(v + \hat{\eta}) Y_{l+i+2}(v + \hat{\eta})$ with $16N_m$ flops, which are common for every hypothesis. Then, $|\Omega_{csd}(c)|$ is implemented using $8N_m$ flops for each trial, resulting in a total of $2704N_m$ flops to obtain the estimate $\hat{s}$ by locating a maximum of $|\Omega_{csd}(c)|$. Similarly, the NSD approach uses $12(N_m - 1)$ flops to evaluate the term $Y_{l+i+2}(v + \hat{\eta}) Y_{l+i+2}^*(v + \hat{\eta} - 1)$ for $\mathcal{D}$ and $i = 0, 1$. For every hypothesis, the cost function $|\Omega_{nsd}(c)|$ is computed using $8N_m - 8$ flops. Consequently, $2700(N_m - 1)$ flops are required to obtain the SSID $\hat{s}$ employing the NSD approach. In the case of the RSD approach, $|\Omega_{rsd1}(b)|$ is evaluated using $8N_m + 6$ flops for each trial and then a total of $16N_m + 112(8N_m + 6)$ flops is used to obtain the SSID $\hat{s}_1$. In this way, the computation of $|\Omega_{rsd0}(a)|$ for $a \in \{0, 1, 2\}$ involves $3(8N_m + 6)$ flops. As a result, the RSD approach can be implemented with $936N_m + 690$ flops. Finally, we evaluate the complexity of the proposed approach. Considering all hypotheses associated with (20), the proposed method requires $112(4N_m + 6)$ real multiplications (RMs) and $336(4N_m - 3)$ real additions, which are eventually converted to $1792N_m + 672$ flops. Table 1 summarizes the computational complexity of SL-SSS detection methods. As one can see from (18), the RCFO detector is realized with one angle operation and one RM.

Table 1. Complexity of the SL-SSS detection methods.

Method	Number of Flops
CSD method	$2704N_m$
NSD method	$2700(N_m - 1)$
RSD method	$936N_m + 690$
Proposed method	$1792N_m + 672$

4.3. Performance Evaluation

Figure 2 presents $\text{Prob}\{(\hat{s}_0, \hat{s}_1) = (s_0, s_1)\}$ for the conventional RSD and proposed algorithms under both AWGN and fading conditions, as a function of average SNR. To validate the accuracy of the derived detection probability, we consider the case where $D_f = 0$ Hz and RSTO is equal to zero. However, the impact of Doppler frequency and RSTO on the performance of the proposed method will be examined in the following example. Although not shown in Figure 2, the detection curves for both the CSD and proposed schemes are nearly identical because they use the same cost function. On the other hand, it is observed that the proposed SL-SSS detector has a competitive performance improvement over the RSD method in terms of detection probability. And, more importantly, the analytical curves are very well matched with the simulated curves in the presence of frequency offset values under the assumption of perfect ICFO and SL-PSS detection. These observations confirm the effectiveness of the proposed SL-SSS detection method and the accuracy of theoretical expressions derived in Section 3.3.

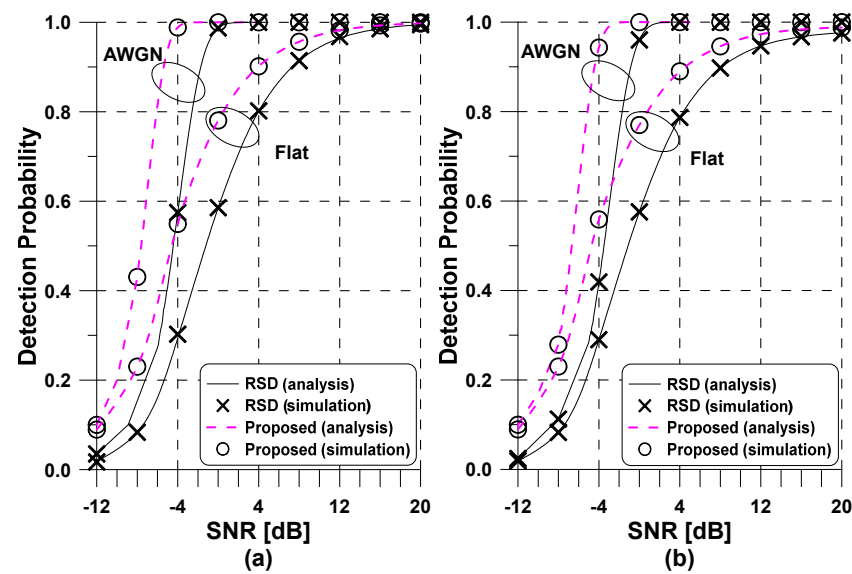


Figure 2. Performance of the proposed SL-SSS detectors in AWGN and flat-fading channels: (a) $\varepsilon = 0$, (b) $\varepsilon = 0.1$.

Figure 3 shows the performance comparison of different SL-SSS detection methods versus average SNR under various channel models, assuming that RSTO is equal to zero. While the AML method offers a substantial gain over the proposed SL-SSS detection method, its heavy computational complexity severely limits the efficiency. As we can see, the proposed SL-SSS detection scheme attains comparable detection performance to other baselines such as the NSD and CSD methods, which suffer from significant computational load caused by a huge number of possible hypothesis testing outcomes. Consequently, the proposed method is an effective approach to compromise a better trade-off between detection performance and computational complexity. In the case of the Vld channel, the capacity enhancement of the presented SL-SSS detectors becomes more significant as the maximum channel delay spread increases, which is due to multipath diversity.

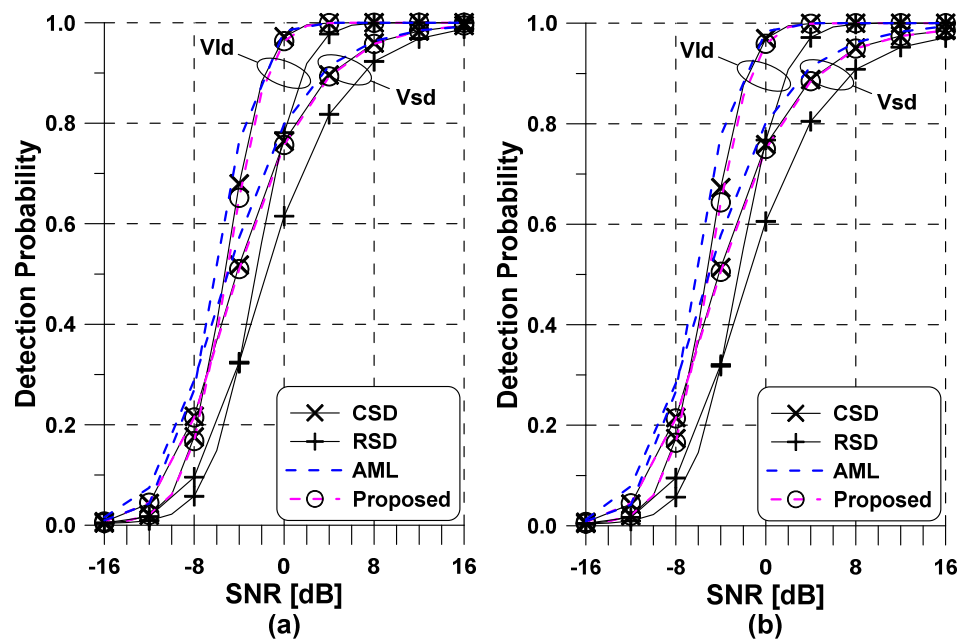


Figure 3. Performance comparison between the conventional and proposed SL-SSS detectors with respect to SNR values: (a) $\varepsilon = 0$, (b) $\varepsilon = 0.1$.

To further evaluate the usefulness of the proposed SL-SSS detector, we consider the overall detection performance denoted by $\text{Prob}\{(\hat{\eta}, \hat{u}, \hat{s}_0, \hat{s}_1) = (\eta, u, s_0, s_1)\}$ under different STO and Doppler frequency conditions. For this purpose, η is randomly chosen within a range from -4 to 4 . Figure 4 exhibits the performance of various SL-SSS detection methods when the UE device operates at an SNR of 4 dB. As expected, increased UE velocity severely degrades the performance of the detection methods employing coherent strategy, while the AML and NSD methods are less sensitive to the amount of Doppler frequency. Therefore, we can conclude that the proposed approach can attain near-optimal detection performance while keeping lower computational costs compared with other methods.

Figure 5 illustrates $\mathbb{E}\{|\Delta_\varepsilon|^2\}$ of the RCFO detection scheme as a function of SNR under various channel conditions. To verify the feasibility of the RCFO estimation scheme, we provide a Cramer–Rao lower bound (CRLB) as a benchmark [21]. In Figure 5a, it is evident that the MSE of the RCFO estimation scheme closely approximates the CRLB for moderate SNR levels in the case of the AWGN channel. Conversely, the presence of a Doppler frequency introduces a loss of orthogonality in the frequency-selective fading channel, which in turn impairs the estimation accuracy of the RCFO detector. Another notable observation regarding the suggested RCFO detector is that the simulated MSE demonstrates substantial concordance with the theoretical MSE when the SNR exceeds -2 dB. Meanwhile, if the SNR value falls below a certain threshold, the detection capability of the RCFO detection method is significantly impaired due to the constrained estimation range, which is also confirmed by Figure 5b. As demonstrated in (18), we observe that the estimation range of the RCFO estimator is constrained to $|\varepsilon| \leq 1/(4\varrho) \approx 0.24$. Within this estimation range, it becomes apparent that the detection performance deteriorates around $\varepsilon = 0$ as the RCFO value increases. This phenomenon is primarily attributed to the existence of an irreducible ICI term that persists even as the SNR approaches infinity.

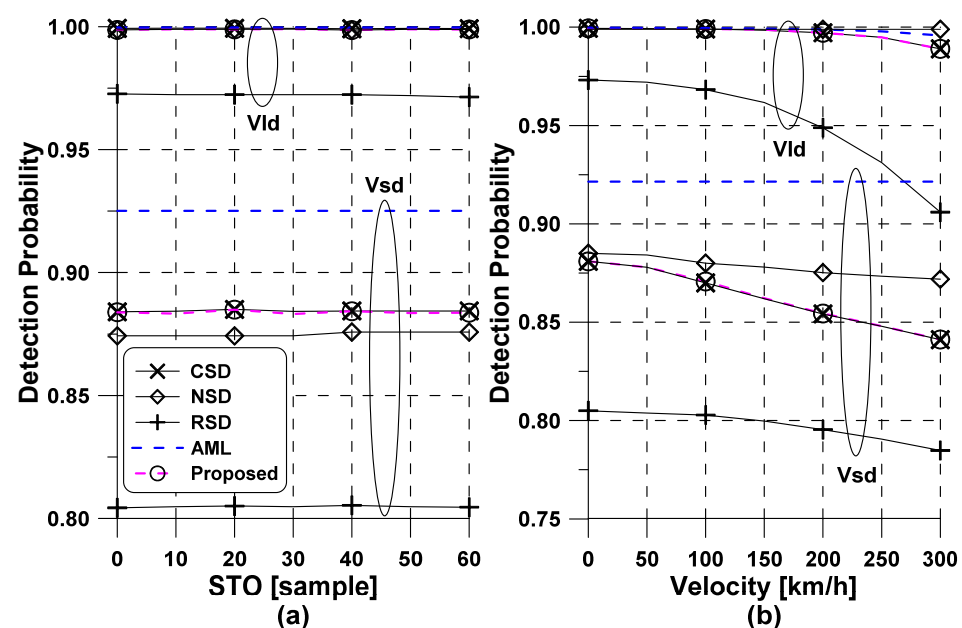


Figure 4. Performance comparison between the conventional and proposed SL-SSS detectors versus (a) STO and (b) UE speed when SNR = 4 dB.

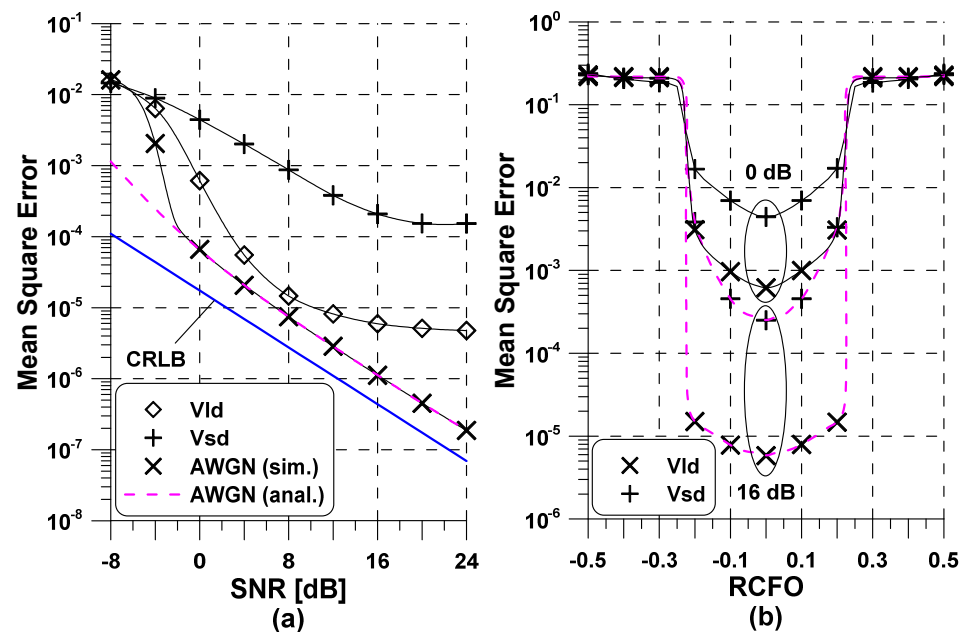


Figure 5. MSE of the RCFO estimation scheme under various channel conditions: (a) $\varepsilon = 0$, (b) $SNR = 0$ and 16 dB.

5. Concluding Remark

In this paper, a computationally effective SL-SSS synchronization scheme is presented in a cellular NR-V2X communication system. Taking advantage of the cyclic-shifted property between SL-SSS sequences, we highlight a low-complexity implementation of the proposed SL-SSS detector. The proposed method offers the advantage of simultaneously estimating RCFO by utilizing the same cost function employed for SLSSID detection. Therefore, the total complexity of the initial synchronization procedure can be considerably reduced. We derive the detection probability of the proposed SLSSID detector and the MSE of the RCFO estimator, demonstrating their accuracy through simulation results. Numerical results show that the computational complexity of the proposed SL-SSS detector is reduced compared with the conventional SL-SSS detector while maintaining the same detection performance. Applying the proposed synchronization method to the 5G NR-V2X communication system results in a reduction in computational complexity by over 66%, leading to decreased power consumption and ultimately enhancing the battery life of NR-V2X terminals.

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