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Improved Joint Estimation of Set Information and Frequency Offset for LTE Device-to-Device Communications

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Abstract: Device-to-device (D2D) communication is a key enabler to facilitate the realization of the Internet of Things in which direct communication between mobile users is allowed without the intervention of the base station. One of the most challenging issues in the D2D cellular network is a low-complexity and reliable initial synchronization. To solve this issue, an efficient joint detection of integer carrier frequency offset (IFO) and set information (SI) is proposed for D2D communication in a long term evolution (LTE) system. Unlike the conventional approach, the primary sidelink synchronization signal (PSSS) and secondary sidelink synchronization signal (SSSS) transmitted in the LTE-D2D system are used together to remove the effect of channel fading and symbol timing offset (STO). In addition, the conjugate property between two PSSS sequences is utilized. This design facilitates robust sequential detection of the IFO and SI without priori information on two synchronization signals when residual STO is present. It has been demonstrated that the inherent properties of the PSSS and SSSS sequences are effectively used for efficient cell search in the presence of residual STO, compared to existing methods.

Keywords: device-to-device; internet of things; sidelink synchronization; set information; frequency synchronization

1. Introduction

Orthogonal frequency division multiplexing (OFDM) has been shown to be an efficient modulation technique in wireless communication systems due to its excellent bandwidth efficiency and resistance to channel impairments. Long term evolution (LTE) is an attractive technology to be applied for a variety of wireless applications due to its ability to provide high spectral efficiency and low latency using OFDM transmission, which has been standardized by the third generation partnership project (3GPP) [1]. Recently, cellular internet of things (IoT) solutions such as narrowband IoT and device-to-device (D2D) communications have been paid much attention as a new communication paradigm in existing and future cellular LTE networks [2–8]. D2D is considered a promising technology that provides a mechanism for reusing LTE's uplink physical layer such as single-carrier frequency division multiple access (SC-FDMA) and reference signals [5,6]. Thus, IoT that enables ubiquitous connection among D2D devices without the essential control of the centralized supervision is a prime enabler to facilitate the evolution of routing the traffic via enhanced base station (eNodeB) from LTE-A systems to future 5G [8–11].

In the LTE-D2D network, it is a challenging task to maintain the synchronization especially for out-of-network coverage scenarios, wherein the D2D devices have to synchronize with each other.

To this end, the D2D device has not only to estimate symbol timing offset (STO) and carrier frequency offset (CFO) but also to capture the sidelink identity [12–18]. Toward this, a synchronization source user equipment (UE) periodically transmits two specified synchronization signals, named as primary sidelink synchronization signal (PSSS) and secondary sidelink synchronization signal (SSSS). Most of the existing synchronization schemes are designed for downlink (DL) detection and are not the best solution for sidelink synchronization, but they can be applied to the D2D scenarios [11–27]. The sidelink identity is acquired by combining the set information (SI) transmitted in the PSSS and the sidelink ID transmitted in the SSSS. The initial cell search procedure in the LTE system generally involves three steps. During the first step of the procedure, the D2D device exploits the redundant cyclic prefix (CP) to find the fractional frequency offset (FFO) and initial STO [12–14]. In the second step, the PSSS is used to retrieve the SI and detect integer frequency offset (IFO) [20–27]. After the completion of this stage, the SSSS is identified to determine the frame boundary and the sidelink ID [16–18].

Once the IFO and SI acquisition has been successfully achieved, the D2D UE can attempt to decode the SSSS and thus the correct detection of the PSSS is the most challenging task in the overall synchronization process. There are many works concentrated mainly on the detection of the PSSS in the literature [19–27]. Some of these methods perform in the time domain [19,20], while others use the frequency-domain samples obtained from the discrete Fourier transform (DFT) unit [21–27]. In [19,20], PSSS detection scheme has been presented to be performed by looking for the peak of the cross-correlation in the time domain. However, the performance of these approaches is significantly degraded in the case when the IFO is present. To solve this drawback, various differential correlation based PSSS detection schemes have been proposed in the frequency domain [21–23], wherein the IFO is estimated along with the PSSS. As for the IFO, it can be estimated either in a joint or decoupled way. The works in [21–24] estimate the IFO together during the PSSS match process, which is accomplished by evaluating a frequency-shifted version of the received PSSS and requires a large amount of computational complexity because of using correlation banks for a large number of hypotheses. Maximum likelihood (ML) detection has been efficiently adopted to detect the IFO and PSSS in [25,26], which are based upon the reduced-rank approximation of the channel and the PSSS matrix, respectively. Although ML detection can achieve optimal performance, it requires a computationally intensive exhaustive search. The differential correlation based method with low complexity has been presented in [27], wherein the joint detection problem of the IFO and SI is decoupled by utilizing the central-symmetric feature of the PSSS. A drawback of the sequential method is its sensitivity to residual STO. Furthermore, the performance of differential correlation based PSSS detection schemes is severely degraded because of a poor autocorrelation property of the differential PSSS. For the D2D UE's simple operation, therefore, it is of crucial importance to develop a reduced-complexity and robust joint synchronization scheme in the LTE-D2D system.

To address this issue, a robust scheme for the IFO recovery and SI detection is proposed in the LTE-D2D system. Unlike the existing methods, two received PSSS and SSSS sequences are used to sequentially detect the IFO and SI. The intrinsic properties associated with the PSSS and SSSS facilitate sequential detection without priori information on two synchronization signals. Simulation results demonstrate that the proposed sequential detection method offers strong resistance to symbol timing error and achieves a reduced complexity, compared to the conventional joint detection schemes.

The remainder of this paper is structured as follows. In Section 2, the OFDM system model and synchronization signal are introduced. Section 3 highlights the existing joint detection schemes as benchmark. In Section 4, the joint detection method is proposed in the LTE-D2D system. The effectiveness of the presented method is verified through computer simulations in Section 5, and a summary is given in Section 6.

2. System Description

2.1. System Model

We assume that the synchronization signal is transmitted without DFT precoding [11]. The system of interest uses OFDM symbols with N subcarriers for the data transmission. At the transmitter, the quadrature amplitude modulation (QAM) signal is converted into the time domain using an N -point inverse fast Fourier transform (IFFT) operation, which is equivalent to modulate every N symbols onto N orthogonal subcarriers. Then, a CP of N_g samples is appended at the front of OFDM symbol in order to avoid both the inter-symbol interference (ISI) and inter-carrier interference (ICI). Accordingly, the effective OFDM signal during the l -th period is generated as

$$x_l(m) = \sum_{k=-N/2}^{N/2-1} X_l(k) e^{j2\pi km/N}, \quad m = -N_g, -N_g + 1, \dots, N - 1, \quad (1)$$

where l denotes the index of the OFDM symbol and $X_l(k)$ is the frequency-domain QAM symbol at the k -th subcarrier over the l -th symbol duration. The remaining $X_l(k)$'s except for the synchronization signal are DFT-precoded signals.

The signal is transmitted over frequency-selective fading channels and corrupted by additive white Gaussian noise (AWGN). Because the performance of the conventional IFO estimation methods developed in [25–27] is marginally affected by the presence of the FFO, we assume the FFO is completely estimated. The accuracy of STO estimate may be considerably deteriorated in the presence of multipath channels. For this reason, we consider the scenario where there is a residual STO τ after initial STO estimation is performed. The received signal is sampled at a sampling frequency of $f_s = N\Delta_f$ with Δ_f being the subcarrier spacing. Consequently, the m -th time-domain sample over the l -th symbol duration $y_l(m)$, $m = -N_g, -N_g + 1, \dots, N - 1$, is written as [20]

$$y_l(m) = e^{j l \epsilon \rho} e^{j 2 \pi \epsilon (m - \tau) / N} \sum_{p=1}^L c_l(p) x_l(m - \tau_p - \tau) + z_l(m), \quad (2)$$

where $\rho = 2\pi N_g / N$, ϵ is the normalized IFO by the subcarrier spacing Δ_f , $c_l(p)$ denotes the channel impulse response with L multipaths, τ_p is the delay of the p -th path, and $z_l(m)$ is the m -th AWGN sample. It is assumed that the IFO is uniformly distributed in range $[-M, M]$ with M being the maximum value of the IFO that determined by the local oscillator's stability. Given the assumption that the CP is longer than the maximum excess delay of the channel, QAM symbols can be demodulated taking a FFT on the CP-removed time-domain sample $y_l(m)$ as follows

$$Y_l(k) = C_l(k - \epsilon) X_l(k - \epsilon) e^{-j 2 \pi (k - \epsilon) \tau / N} e^{j l \epsilon \rho} + Z_l(k), \quad (3)$$

where k represents the subcarrier index, $X_l(k)$ is the frequency-domain modulated QAM symbol at the k -th subcarrier, $C_l(k)$ is the channel frequency response (CFR) over the l -th period, and $Z_l(k)$ is a complex zero-mean AWGN with variance σ_Z^2 .

2.2. Synchronization Signal

Let us consider the sidelink of an LTE system. As depicted in Figure 1, one LTE radio frame has a duration 10 ms and consists of equally sized 20 slots each of 0.5 ms. Each slot in turn is made up of a number of SC-FDMA symbols which can be either seven or six based on the CP mode. In contrast to LTE DL, the D2D sidelink uses a pair of PSSS and SSSS, which are periodically transmitted in two consecutive OFDM symbols every 40ms. Physical sidelink broadcast channel (PSBCH) can be used as synchronization measurement for determining whether or not the out-of-coverage UE becomes a synchronization source, while demodulation reference signal (DMRS) can be used to produce channel

estimates for demodulation of the associated physical channel. An LTE sidelink network can support 336 different sidelink identities indexed by $N_{ID} \in \{0, 1, 2, \dots, 335\}$, which is retrieved from the combination of the SI embedded in the PSSS and the sidelink ID embedded in the SSSS. The SI present in the PSSS informs whether the synchronization source UE is in-network or out-of-network. The in-network-coverage case is numbered as $\{0, 1, 2, \dots, 167\}$, while the out-of-network scenario is numbered as $\{168, 169, \dots, 335\}$. The PSSS is generated based on Zadoff-Chu (ZC) sequence that has the property of constant amplitude zero autocorrelation. The ZC root sequence index u_i ($i = 0, 1$) is used to represent two sets corresponding to in-network and out-of-network cases. A UE within the network coverage transmits PSSS symbols with $u_0 = 26$ for $N_{ID} \leq 167$ and transmits PSSS symbols with $u_1 = 37$ otherwise. The PSSS is generated from a frequency-domain ZC sequence of length 63, with the middle part zero-padded to avoid modulating the DC subcarrier. Thus, the DC subcarrier and the rest of the subcarriers in the symbols where the PSSS is located are nulled and 31 subcarriers are mapped on each side of the DC subcarrier. Two PSSS sequences are constructed to have good periodic autocorrelation and cross-correlation properties [28]

$$P_i(k) = \begin{cases} e^{-j\pi u_i(k^2+992)/63-j\pi u_i k}, & k \in \mathcal{P} \\ 0, & \text{otherwise} \end{cases} \quad (4)$$

where $\mathcal{P} = \{k | -N_p/2 \leq k \leq N_p/2, k \neq 0\}$ is the set of PSSS subcarriers with N_p being the number of non-zero PSSS subcarriers.

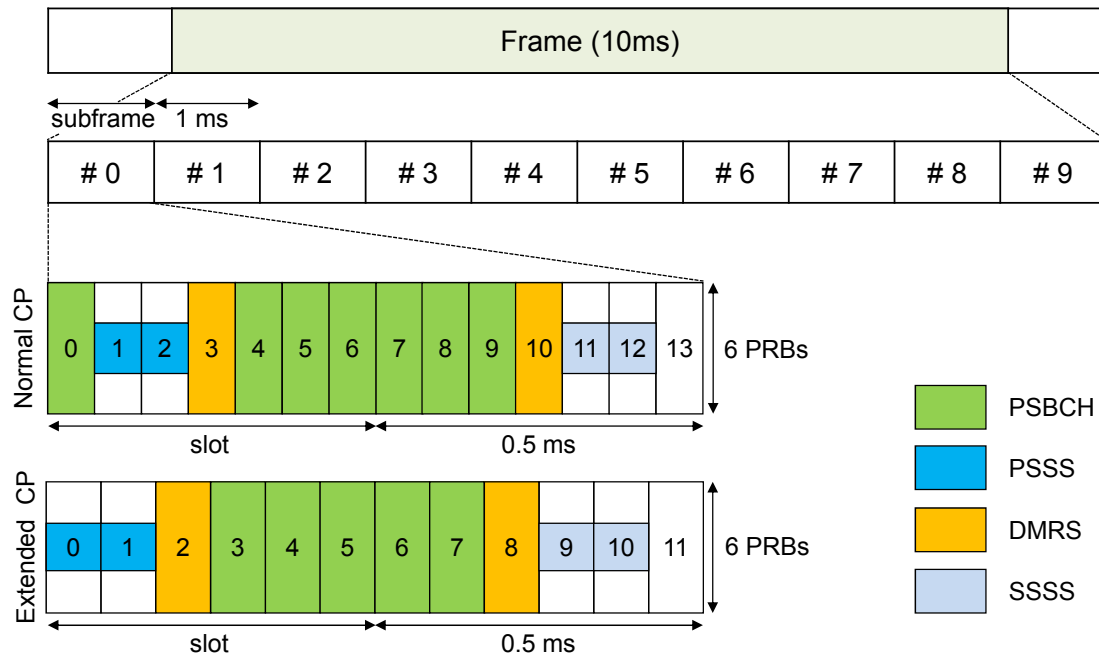


Figure 1. Frame structure in the long term evolution (LTE)-device-to-device (D2D) system.

The primary usage of the SSSS is to retrieve the sidelink identity and to acquire the beginning point of an LTE frame. The SSSS is a frequency-domain binary phase-shift keying (BPSK) sequence, which is an interleaved combination of two length-31 scrambled maximum length sequences $s_{w_0}(q)$ and $s_{w_1}(q)$, with $q = 0, 1, \dots, N_p/2 - 1$. In the case of the normal CP, the SSSS is generated as

$$H(2q) = \begin{cases} s_{w_0}(q)h_0(q) & \text{in SC-FDMA symbol 11} \\ s_{w_1}(q)h_0(q) & \text{in SC-FDMA symbol 12} \end{cases} \quad (5)$$

$$H(2q+1) = \begin{cases} s_{w_1}(q)h_1(q)d_{w_0}(q) & \text{in SC-FDMA symbol 11} \\ s_{w_0}(q)h_1(q)d_{w_1}(q) & \text{in SC-FDMA symbol 12} \end{cases} \quad (6)$$

where $h_i(q)$ and $d_{w_i}(q)$ represent the scrambling sequences. The subscripts w_0 and w_1 are directly related with the value of the sidelink identity. The sequence $H(q)$ is mapped onto central N_p SSSS subcarriers with indices $k \in \mathcal{P}$, yielding the SSSS sequence $S_j(k)$ with sidelink ID j . The arrangement of the PSSS and SSSS sequences differs depending on the CP type. For simple notation, we consider the situation where the first PSSS is transmitted at the l -th symbol, i.e., $X_l(k) = P_i(k)$. According to the CP type, the first SSSS is denoted as

$$S_j(k) = \begin{cases} X_{l+10}(k), & \text{for the normal CP} \\ X_{l+9}(k), & \text{for the extended CP} \end{cases} \quad (7)$$

where $j \in \{0, 1, 2, \dots, 167\}$. It is apparent from (3) that ϵ and u_i affect the frequency domain signal $Y_l(k)$ together, thus necessitating the use of joint detection in the traditional synchronization methods. The processing complexity of the IFO estimation can be significantly reduced when it is performed jointly with PSSS detection because there are fewer hypotheses in the PSSS matching process than in the SSSS.

Similar to 4G networks, the 5G new radio (NR) synchronization signal is a physical layer specific signal and helps UEs to detect the cell ID. Unlike 4G LTE, the 5G NR PSSS consists of one of three 127-symbols m-sequences and is allocated on the first symbol of each synchronization signal block (SSB), while the SSSS consists of one of 336 127-symbols gold sequences and is allocated on the third symbol of each SSB [29]. Due to the absence of the conjugate property of the synchronization signal, the proposed scheme cannot be directly applied in the 5G-NR system. The 5G-NR encompasses both an evolution of 4G networks and the NR, the synchronization issue in the 4G network is still important. In this paper, we focus on the initial synchronization scheme in the 4G LTE-D2D system.

3. Conventional Detection Scheme

To remove the impact of the channel on the detection performance, the works developed in [21–27] perform differential correlation, wherein the effect of STO is mitigated together during non-coherent detection. Based on the assumption that the l -th transmitted symbol $X_l(k)$ is equal to $P_i(k)$ and $C_l(k) \approx C_l(k-1)$, the differential correlation between neighboring subcarriers of the received PSSS sequence can be expressed as

$$\begin{aligned} \bar{Y}_l(k) &= Y_l(k)Y_l^*(k-1) \\ &\approx |C_l(k-\epsilon)|^2 D_i(k-\epsilon)e^{j2\pi\tau/N} + \bar{Z}_l(k), \quad k \in \mathcal{D}, \end{aligned} \quad (8)$$

where a^* denotes the complex conjugate of a , $|\cdot|$ denotes the magnitude operation, $D_i(k) = P_i(k)P_i^*(k-1)$, $\bar{Z}_l(k)$ denotes the contribution to AWGN, and $\mathcal{D} = \{k | -N_p/2 + 1 \leq k \leq -1, 2 \leq k \leq N_p/2\}$. As benchmark, two representative conventional schemes are considered in this paper. One is a sequential IFO and SI detection scheme [27] and the other is a low-complexity joint detection method [23].

3.1. Reduced-complexity Sequential Detection (RCSD) Scheme

Recalling from (4) that $P_i(k) = P_i(-k)$, the property of $D_i(k) = D_i^*(-k+1)$ is valid for any i and $k \in [2, N_p/2]$. Using the symmetric feature of the PSSS, the sequential detection of the IFO and SI is proposed in [27]. A normalized PSSS by the channel magnitude is used to remove the effect of channel fading, taking the expression

$$\Omega_a(v) = \frac{1}{N_p/2 - 1} \sum_{k=2}^{N_p/2} \frac{\tilde{Y}_l(k+v)\tilde{Y}_l(-k+v+1)}{|\tilde{Y}_l(k+v)||\tilde{Y}_l(-k+v+1)|} \quad (9)$$

which approaches to one when signal-to-noise ratio (SNR) increases at a true IFO trial $v = \epsilon$. Based on this property, the IFO is detected by minimizing the quantity $|\Omega_a(v) - 1|$

$$\hat{\epsilon} = \arg \min_v |\Omega_a(v) - 1|, \quad (10)$$

which is performed independent of the PSSS. Based upon the estimation of ϵ , the SI is obtained via one dimensional search

$$\hat{n} = \arg \max_n |\Omega_b(n)|, \quad (11)$$

with

$$\Omega_b(n) = \sum_{k=2}^{N_p/2} \left\{ \frac{\tilde{Y}_l(k+\hat{\epsilon})}{|\tilde{Y}_l(k+\hat{\epsilon})|} + \frac{\tilde{Y}_l^*(-k+\hat{\epsilon}+1)}{|\tilde{Y}_l^*(-k+\hat{\epsilon}+1)|} \right\} D_n^*(k). \quad (12)$$

This decoupled strategy can avoid two-dimensional search, leading to substantial complexity reduction. Therefore, the RCSD scheme achieves good detection performance with low complexity when there is no residual STO, while the presence of residual STO has a negative effect on the detection accuracy.

3.2. Reduced-Complexity Joint Detection (RCJD) Scheme

In [23], a possible simplification to reduce computational complexity is proposed by partitioning the set $\mathcal{D}$ having $N_p - 2$ differential PSSS subcarriers into a number of subsets. The s -th subset with ZC root index u_n denoted by $\mathcal{D}_n^s$ is computed by

$$\mathcal{D}_n^s = \{k | |\Re\{D_n(k_n^s)D_n^*(k)\}| \geq T\}, \quad (13)$$

where $\Re\{\cdot\}$ is the real part of the enclosed quantity, k_n^s is the representative subcarrier of the s -th subset, and T is the pre-determined threshold that controls the correlation between differential PSSS subcarriers in each subset. Since $\mathcal{D}_n^s$'s are disjoint, the union of all subsets is equal to $\mathcal{D}$. Based on the subset partitioning, the corresponding metric is given by

$$\Omega_c(v, n) = \sum_{s=1}^{N_{b,n}} \left(\sum_{k \in \mathcal{D}_n^s} B_n(k) \tilde{Y}_l(k+v) \right) D_n^*(k_n^s), \quad (14)$$

where $N_{b,n}$ is the number of subsets corresponding to $D_n(k)$, and $B_n(k) \in \{1, -1\}$ is the conversion indicator that makes the phase difference between the representative PSSS subcarrier $D_n(k_n^s)$ and the other subcarriers $D_n(k)$'s in $k \in \mathcal{D}_n^s$ to be equal. By finding the maximum of $\Omega_c(v, n)$ with respect to v and n , the IFO and SI are jointly estimated by

$$(\hat{\epsilon}, \hat{n}) = \arg \max_{(v, n)} |\Omega_c(v, n)|. \quad (15)$$

A main drawback of the reduced-complexity joint detection (RCJD) scheme is that a loss of autocorrelation property of the original PSSS produces similar peaks for different IFO hypotheses and deteriorates the detection performance.

4. Proposed Detection Method

In this section, a robust IFO and SI estimation method is proposed for the LTE-D2D system. By using the conjugate property between two PSSS signals and symmetric feature of the SSSS, the IFO can be solely detected independent of the synchronization signals, facilitating joint estimation of the IFO and SI to be detected sequentially. As a performance measure, the probability of detection failure of the proposed detection scheme is theoretically calculated.

4.1. Sequential Detection Algorithm

To remove the effect of the channel fading and residual STO, a conjugate product from the received PSSS and SSSS sequences is defined as

$$\begin{aligned} R_l(k) &= Y_l^*(k)Y_{l+D_t}(k) \\ &\approx |C_l(k-\epsilon)|^2 P_i^*(k-\epsilon)S_j(k-\epsilon)e^{-jD_t\epsilon\rho} + W_l(k) \end{aligned} \quad (16)$$

with

$$\begin{aligned} W_l(k) &= C_{l+D_t}(k-\epsilon)S_j(k-\epsilon)Z_l^*(k)e^{-j2\pi(k-\epsilon)\tau/N}e^{j(l+D_t)\epsilon\rho} \\ &\quad + C_l^*(k-\epsilon)P_i^*(k-\epsilon)Z_{l+D_t}(k)e^{j2\pi(k-\epsilon)\tau/N}e^{-jl\epsilon\rho} + Z_{l+D_t}(k)Z_l^*(k) \end{aligned} \quad (17)$$

where D_t is the distance between the PSSS and SSSS sequences. Note that $D_t = 9$ indicates the extended CP mode, while $D_t = 10$ corresponds to the normal CP mode. For notational convenience, it is assumed in (16) that channel remains the same during one subframe period. However, the impact of time selectivity of the channel on the detection performance will be evaluated later by means of simulations. We observe from (16) that it requires knowledge of the transmitted two synchronization signals. To enable sequential estimation of the IFO and SI, the intrinsic property of the PSSS and SSSS sequences is exploited.

The sequential detection of the IFO and SI is based on observation of the correlation between the v -shifted version of $R_l(k)$ and local PSSS

$$\begin{aligned} \bar{R}_n(k+v) &= R_l(k+v)P_n(k)e^{jD_tv\rho} \\ &= |C_l(k+v-\epsilon)|^2 S_j(k+v-\epsilon)P_i^*(k+v-\epsilon)P_n(k)e^{jD_t(v-\epsilon)\rho} \\ &\quad + W_l(k)P_n(k)e^{jD_tv\rho}, \quad D_t = 9, 10. \end{aligned} \quad (18)$$

The conjugate symmetric property of the SSSS in the time domain ensures that $S_j(k)$ is binary in the frequency domain. To remove the dependency on the SSSS, an objective function is defined as

$$\Omega_s(v, n) = \sum_{k \in \mathcal{P}} [\bar{R}_n(k+v)]^2, \quad (19)$$

where $[\cdot]^2$ denotes the square operator. Using the fact that $P_0(k) = P_1^*(k)$, the real part of (19) can be expressed in terms of $P_0(k)$

$$\begin{aligned} \Re\{\Omega_s(v, n)\} &= \sum_{k \in \mathcal{P}} \Re\{[R_l(k+v)]^2\} \Re\{[\bar{P}_0(k)]^2\} \\ &\quad + (-1)^{n+1} \sum_{k \in \mathcal{P}} \Im\{[R_l(k+v)]^2\} \Im\{[\bar{P}_0(k)]^2\}, \quad n = 0, 1, \end{aligned} \quad (20)$$

where $\bar{P}_n(k) = P_n(k)e^{jD_tv\rho}$, $\Im\{\cdot\}$ is the imaginary part of the enclosed quantity, and

$$\begin{aligned} [R_l(k+v)]^2 &= |C_l(k+v-\epsilon)|^4 [S_j(k+v-\epsilon)]^2 [P_i^*(k+v-\epsilon)]^2 e^{-j2D_t\epsilon\rho} \\ &\quad + 2|C_l(k+v-\epsilon)|^2 S_j(k+v-\epsilon)P_i^*(k+v-\epsilon)e^{-jD_t\epsilon\rho} + [W_l(k)]^2. \end{aligned} \quad (21)$$

It is interesting to note that the uncertainty related with the PSSS is only the polarity in the second term on the right-hand side (RHS) of (20). Based on this observation, the complex conjugate relation between two PSSS sequences is exploited to estimate the IFO without relying on the transmitted PSSS information. For this purpose, we propose a novel metric as

$$\Omega_f(v) = \sum_{k \in \mathcal{P}} \Re \left\{ [R_l(k+v)]^2 \right\} \Re \left\{ [\bar{P}_0(k)]^2 \right\} + \left| (-1)^{n+1} \sum_{k \in \mathcal{P}} \Im \left\{ [R_l(k+v)]^2 \right\} \Im \left\{ [\bar{P}_0(k)]^2 \right\} \right|, \quad n = 0, 1, \quad (22)$$

which indicates that $\Omega_f(v)$ is constant to the index n .

Under the hypothesis that $v = \epsilon$ and $n = i$, using $[S_j(k+v-\epsilon)]^2 = E_X$, (22) can be expressed as

$$\Omega_f(v) = E_X \sum_{k \in \mathcal{P}} |C_l(k)|^4 \Re \left\{ [P_i^*(k)]^2 \right\} \Re \left\{ [P_0(k)]^2 \right\} + \sum_{k \in \mathcal{P}} \Re \left\{ \bar{W}_l(k) \right\} \Re \left\{ [\bar{P}_0(k)]^2 \right\} + \left| E_X \sum_{k \in \mathcal{P}} |C_l(k)|^4 \Im \left\{ [P_i^*(k)]^2 \right\} \Im \left\{ [P_0(k)]^2 \right\} + \sum_{k \in \mathcal{P}} \Im \left\{ \bar{W}_l(k) \right\} \Im \left\{ [\bar{P}_0(k)]^2 \right\} \right|, \quad (23)$$

where $\bar{W}_l(k) = 2|C_l(k+v-\epsilon)|^2 S_j(k+v-\epsilon) P_i^*(k+v-\epsilon) e^{-jD_l \epsilon \rho} + [W_l(k)]^2$ and E_X denotes the symbol energy. It is worthy mentioning that $\Omega_f(v)$ is independent of the transmitted PSSS and SSSS under hypothesis that the signal is synchronized. This observation means that the detection performance for the normal and extended CP modes is approximately the same under the assumption $C_l(k) \approx C_{l+D_l}(k)$, which is valid for the situation where the mobility of the D2D UE is limited [30]. When the signal is not synchronized, i.e., $v \neq \epsilon$, it is obvious from (21) that $[S_j(k+v-\epsilon)]^2$ becomes E_X , while $[P_i^*(k+v-\epsilon)P_n(k)]^2$ is a random sequence subject to the amount of ϵ . Under this hypothesis, $\Omega_f(v)$ takes expression

$$\Omega_f(v) = E_X \sum_{k \in \mathcal{P}} |C_l(k+v-\epsilon)|^4 \Re \left\{ [\bar{P}_i^*(k+v-\epsilon)]^2 \right\} \Re \left\{ [\bar{P}_0(k)]^2 \right\} + \sum_{k \in \mathcal{P}} \Re \left\{ \bar{W}_l(k) \right\} \Re \left\{ [\bar{P}_0(k)]^2 \right\} + \left| E_X \sum_{k \in \mathcal{P}} |C_l(k+v-\epsilon)|^4 \Im \left\{ [\bar{P}_i^*(k+v-\epsilon)]^2 \right\} \Im \left\{ [\bar{P}_0(k)]^2 \right\} + \sum_{k \in \mathcal{P}} \Im \left\{ \bar{W}_l(k) \right\} \Im \left\{ [\bar{P}_0(k)]^2 \right\} \right|. \quad (24)$$

Due to the good autocorrelation and cross-correlation properties of $P_n(k)$'s [28], the central limit theorem (CLT) can be invoked such that (24) is regarded as a zero-mean Gaussian random variable (RV). Based on this observation, the IFO can be estimated as follows

$$\hat{\epsilon} = \arg \max_v \left| \Omega_f(v) \right|. \quad (25)$$

After successful detection of the IFO, the SI has to be acquired, which can be performed by

$$\Re \left\{ \Omega_s(\hat{\epsilon}, 0) \right\} \stackrel{u_0}{\geq} \Re \left\{ \Omega_s(\hat{\epsilon}, 1) \right\}. \quad (26)$$

Using (20), (26) can be in an equivalent form

$$\hat{n} = 1 - u \left(\Re \left\{ \Omega_s(\hat{\epsilon}, 0) \right\} - \Re \left\{ \Omega_s(\hat{\epsilon}, 1) \right\} \right) \quad (27)$$

where $u(t)$ is the step function that takes $u(t) = 1$ for $t > 0$ and $u(t) = 0$ otherwise. Substituting (20) to (27), we have

$$\hat{n} = 1 - u \left(\sum_{k \in \mathcal{P}} \Im \left\{ [R_l(k+\hat{\epsilon})]^2 \right\} \Im \left\{ [\bar{P}_0(k)]^2 \right\} \right), \quad (28)$$

which can be implemented without additional computational operations because the enclosed quantity $\sum_{k \in \mathcal{P}} \Re \left\{ [R_l(k + \hat{\epsilon})]^2 \right\} \Re \left\{ [\bar{P}_0(k)]^2 \right\}$ is pre-calculated in (22).

Once (25) and (27) have been completed, the SSSS is exploited to get the sidelink ID. Upon successful detection of the IFO and SI, the sidelink ID detection can be sequentially performed as [17]

$$\hat{g} = \arg \max_g \left| \sum_{k \in \mathcal{P}} \frac{Y_{l+D_t}(k) S_g^*(k)}{\hat{H}_l(k)} \right| \quad (29)$$

where $\hat{H}_l(k)$ is the estimated channel from the PSSS symbol.

4.2. Detection Performance

Denote $P_f = \text{Prob}\{\hat{\epsilon} \neq \epsilon\}$ be the probability that the IFO is falsely detected. The probability of detection failure P_f of (25) in the AWGN channel is derived. Under the hypothesis that the local PSSS with ZC root index u_n is synchronized with the v -th shifted version of the received PSSS with ZC root index u_i , equivalently, when $v = \epsilon$ and $n = i$ (hypothesis H_1), (23) is treated as a complex Gaussian RV based on the CLT. In this case, the mean is $\mu = N_p E_X^3$ and the variance of (23) is calculated by

$$\sigma_1^2 = N_p E_X^2 \sigma_W^2 (4E_X^2 + 3\sigma_W^2) \quad (30)$$

where $\sigma_W^2 = 2E_X \sigma_Z^2 + \sigma_Z^4$ is the variance of (17). When $v \neq \epsilon$ or $n \neq i$ (hypothesis H_0), namely, the signal is not synchronized, $\Omega_f(v)$ in (24) has Gaussian distribution with zero mean and variance

$$\sigma_0^2 = N_p E_X^6 + N_p E_X^2 \sigma_W^2 (4E_X^2 + 3\sigma_W^2). \quad (31)$$

Under hypothesis H_1 , the probability density function (PDF) of $|\Omega_f(v)|$ can be given by

$$p_1(y) = \frac{2y}{\sigma_1^2} e^{-\frac{y^2 + \mu^2}{\sigma_1^2}} I_0 \left(\frac{2y\mu}{\sigma_1^2} \right) \quad (32)$$

where $I_0(\cdot)$ is the modified Bessel function of first kind and zero order. Under hypothesis H_0 , $|\Omega_f(v)|$ has the Rayleigh distribution

$$p_0(y) = \frac{2y}{\sigma_0^2} e^{-\frac{y^2}{\sigma_0^2}}. \quad (33)$$

Since IFOs are equally likely, the probability that the IFO is correctly detected can be given by

$$P_d = \int_0^\infty p_1(y) \left[\int_0^y p_0(x) dx \right]^{2M} dy. \quad (34)$$

Substituting (32) and (33) to (34) yields

$$P_d = \int_0^\infty \frac{2y}{\sigma_1^2} e^{-\frac{y^2 + \mu^2}{\sigma_1^2}} I_0 \left(\frac{2y\mu}{\sigma_1^2} \right) \left[1 - e^{-\frac{y^2}{\sigma_0^2}} \right]^{2M} dy. \quad (35)$$

Using the binomial expansion, (35) can be further derived as

$$P_d = e^{-\frac{\mu^2}{\sigma_1^2}} \sum_{m=0}^{2M} (-1)^m \binom{2M}{m} \int_0^\infty \frac{2y}{\sigma_1^2} e^{-y^2(1/\sigma_1^2 + m/\sigma_0^2)} I_0 \left(\frac{2y\mu}{\sigma_1^2} \right) dy \quad (36)$$

where the integral has a closed-form solution [31]. Let $\gamma_z = E_X/\sigma_Z^2$ be the SNR. After some mathematical manipulations, the probability of detection failure P_f is obtained as

$$P_f = 1 - \sum_{m=0}^{2M} (-1)^m \binom{2M}{m} \frac{1}{1 + m\sigma_1^2/\sigma_0^2} e^{-\frac{m}{\sigma_0^2/\mu^2 + m\sigma_1^2/\mu^2}}. \quad (37)$$

with

$$\frac{\sigma_1^2}{\mu^2} = \frac{\bar{\gamma}_z(4 + 3\bar{\gamma}_z)}{N_p}, \quad (38)$$

and

$$\frac{\sigma_0^2}{\mu^2} = \frac{1 + \bar{\gamma}_z(4 + 3\bar{\gamma}_z)}{N_p}, \quad (39)$$

where $\bar{\gamma}_z = 2/(\gamma_z) + 1/(\gamma_z)^2$ and σ_1^2/σ_0^2 is obtained on dividing (38) by (39).

5. Simulation Results and Discussions

The performance of the proposed sequential detector is evaluated using the system parameters designed for the LTE-D2D system with 5 MHz bandwidth and extended CP. The conducted simulation results is based on the OFDM simulator implemented using Matlab, which uses the 3GPP LTE-D2D specifications. The carrier center frequency is 2 GHz, the subcarrier spacing is $\Delta f = 15$ kHz, the sampling time instant is $T_s = 1/7.68 \mu\text{s}$, the FFT size is $N = 512$, and the GI length is $N_g = 128$. For data transmission, QPSK modulation is used. Specifically, the number of guard subcarriers is 212 and the number of occupied subcarriers is 300, leading to 4.5 MHz occupied bandwidth [1]. We adopt the extended pedestrian A (EPA), extended vehicular A (EVA), extended typical urban (ETU), and extended delay spread (EDS) channel models [32]. In the simulations, the channel is generated using Jakes' model to simulate different fading rates and a different channel realization is generated for each subframe. The channel coefficients are time varying according to the Doppler frequency shift. We consider three Doppler frequencies to represent low, medium, and high speeds of 5 Hz, 70 Hz, and 300 Hz, respectively. Table 1 summarizes the channel profiles used in our experiments. For initial synchronization, we assume the D2D UE may have reference clock uncertainty of ± 10 ppm, which amounts to putting $M = 3$. For all simulations, the IFO for each iteration is distributed uniformly within $[-M, M]$ and the STO is randomly drawn from $[0, \tau_{max}]$ in samples, where τ_{max} is a maximum STO. As for the RCJD scheme, we choose $T = 0.99$ in (13) so that $N_{b,1} = N_{b,2} = 13$ is obtained in (14) [23]. Unless mentioned otherwise, we let $M = 3$, $N_{b,1} = N_{b,2} = 13$, and consider the extended CP type.

The accuracy of the proposed IFO estimation scheme is analytically measured in term of the detection error probability. As a performance measure, the overall probability of detection failure denoted by $P_{o1} = \text{Prob}\{(\hat{\epsilon}, \hat{n}) \neq (\epsilon, i)\}$ and the computational complexity of the presented detection methods are evaluated. As a global performance indicator, the probability that the IFO, SI, and sidelink ID are erroneously detected, denoted by $P_{o2} = \text{Prob}\{(\hat{\epsilon}, \hat{n}, \hat{g}) \neq (\epsilon, i, j)\}$, is also evaluated.

Table 1. Channel profiles.

Model	Channel Delay Spread (μs)	Doppler Frequency (Hz)
EPA	0.41	5
EVA	2.51	70
ETU	5.0	300
EDS	28.58	5

5.1. Complexity Analysis

To compare the complexity of the presented detection schemes, the mathematical operations are converted to the number of real floating point operations (flops). For this purpose, one complex multiplication, one complex addition, one complex magnitude, and one complex square operations are counted as six, two, three, and four flops, respectively [33]. For fair comparison, we assume that the quantities $D_n(k)$'s and $[\bar{P}_n(k)]^2$'s are prior known at the D2D UE. To calculate v -shifted differential correlations $\tilde{Y}_l(k+v)$ and $\tilde{Y}_l(-k+v+1)$ for $2M+1$ trials, it has to perform $N_p + 2M - 1$ complex multiplications. To get the quantity $\Omega_a(v)$ in the RCSD scheme, $9N_p - 19$ flops are required. Finally, (10) requires $6(N_p + 2M - 1) + (2M + 1)(9N_p - 15)$ flops. Since some operations are pre-calculated in (9), $2(5N_p - 9)$ flops are required to get (11). In the case of the RCJD scheme, (14) requires $N_{b,n}$ complex multiplications and $N_p - 3$ complex additions. Accordingly, the total number of flops is $6(N_p + 2M - 1) + (2M + 1)(6N_p - 9 + 6\sum_{n=1}^2 N_{b,n})$. Concerning the proposed sequential detection method, $10(N_p + 2M)$ flops are necessary to compute v -shifted correlation $[R_l(k+v)]^2$ for all v 's. The proposed scheme computes (22) with $4N_p$ flops. Since the quantity in (28) has been already calculated during the IFO-matching process in (22), no extra flop is needed. Considering $(2M + 1)$ hypotheses, $10(N_p + 2M) + 4(2M + 1)N_p$ flops are eventually needed. Figure 2 plots the number of flops used in the joint detectors versus M . Obviously, we can see that the proposed detection scheme has the lowest complexity in comparison to the benchmark schemes, irrespective of the value of M . It is observed that the complexity of the proposed method is considerably reduced, as the local oscillator's stability is getting worse. For $M = 3$, the number of flops of the proposed approach can be decreased by 49.7% and 40.1%, compared to the RCSD and RCJD methods, respectively.

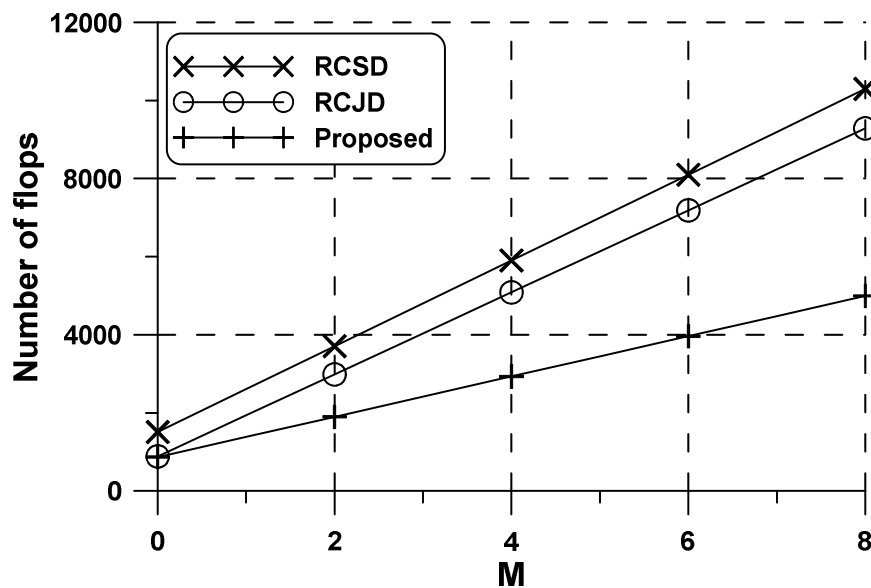


Figure 2. Number of flops of the conventional and proposed joint detectors versus M .

5.2. Performance Evaluation

Figure 3 presents the probability of failure P_f in detecting IFO in the AWGN channel. The theoretical analysis of the proposed scheme is obtained using (37) and the simulated results of the conventional schemes are plotted as benchmark. It is evident that the simulated curves match well with the theoretical curves in the case of the proposed method. More importantly, it can be seen that the probability of detection failure of the existing methods is significantly influenced by the presence of residual STO, while the performance of the proposed detection scheme is independent of τ_{max} . This robustness is obtained by using the temporal correlation from the received PSSS and SSSS sequences instead of differential correlation. Consequently, the influence of residual STO on the detection performance is eliminated in the process of performing (16). However, the performance

degradation of the proposed detection scheme over the conventional methods is observed for low SNR values, which is the penalty for robustness to residual STO. This is attributed to the fact that squaring temporal correlation as described in (19) results in SNR loss. Because of its severe sensitivity to symbol timing errors, the benchmark methods do not work properly in the presence of residual STO especially as the SNR increases.

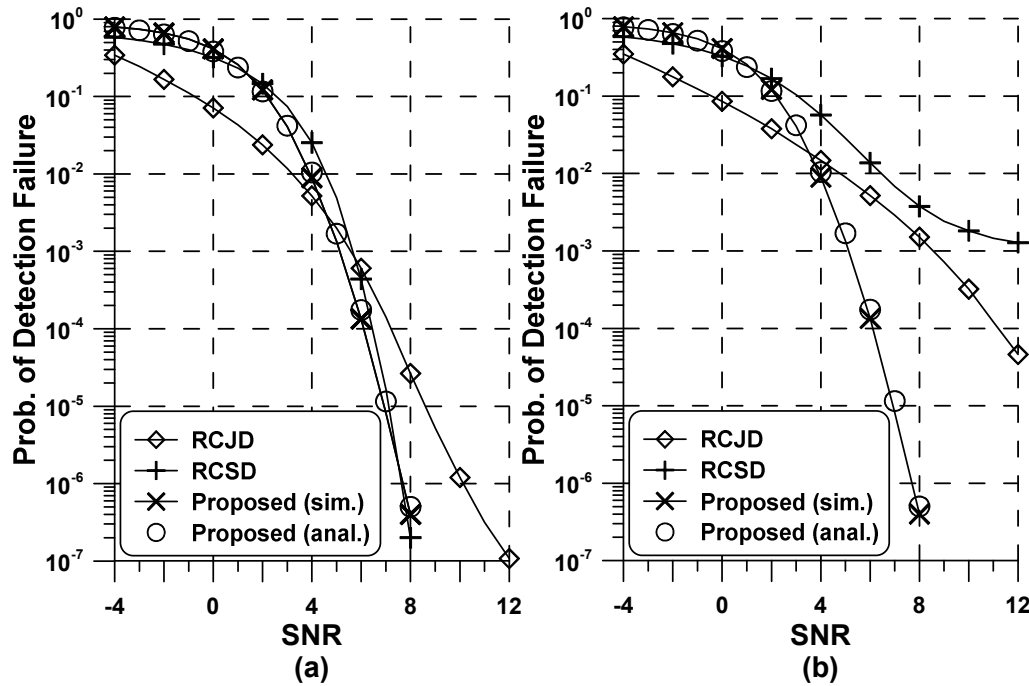


Figure 3. Performance of the proposed and conventional detectors in the additive white Gaussian noise (AWGN): (a) $\tau_{max} = 0$ (b) $\tau_{max} = 10$.

Figures 4 and 5 show the overall probability of failure P_{o2} in jointly detecting IFO, SI, and sidelink ID in the various channel models characterized in Table 1. From the presented results, it can be seen that the detection accuracy of all detection schemes is affected by the amount of the frequency selectivity of the channel. This phenomenon is more pronounced for the conventional detection schemes, which is explained by the fact that the existing detection methods exploit the differential correlation under the assumption $C_l(k) \approx C_l(k-1)$. This approximation proves to be valid for weakly frequency-selective channels as AWGN and EPA models, but is less appropriate for high frequency-selective channels as ETU and EDS. As reported in [34,35], a typical SNR of LTE signal is around 5 dB in most cell areas and over 20 dB SNR is achieved in the cell center, while a low SNR of 0 dB or below can be observed near the cell edge. When there is frequency selectivity and residual STO, the performance of the proposed sequential method is still superior to that of the conventional method since the SSSS detection can be performed only after the PSSS being successfully identified. Specifically, the SNR regions of which the proposed scheme shows better performance are above 5 dB and 1 dB in the EPA and EDS channel models, respectively. Another interesting observation from Figure 5 is that a performance degradation in the proposed scheme occurs due to the high Doppler frequency especially in the ETU channel, however the impact of Doppler frequency on its detection performance is not critical. Since the frequency selectivity and STO are inevitable in D2D applications, it is concluded that the proposed detection scheme achieves good overall detection performance in a typical LTE SNR range with the complexity not higher than existing detection methods.

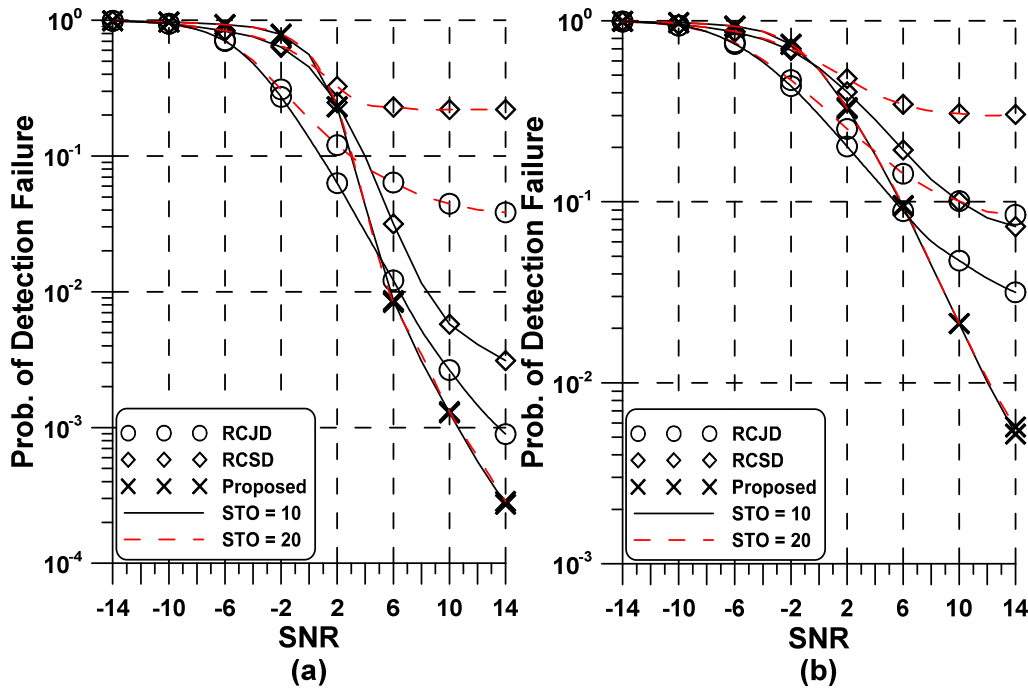


Figure 4. Performance of the proposed and conventional detectors versus SNR: (a) EPA (b) EVA.

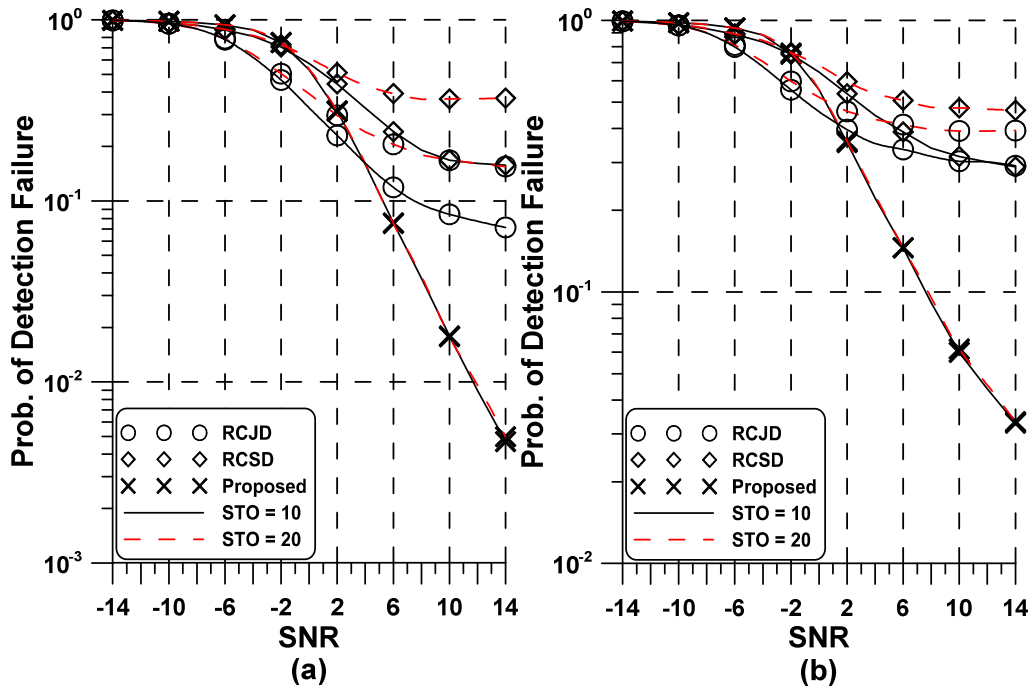


Figure 5. Performance of the proposed and conventional detectors versus signal-to-noise ratio (SNR): (a) extended typical urban (ETU) (b) extended delay spread (EDS).

Figures 6 and 7 depict P_{o1} of the joint detection schemes using the extended CP and normal CP, respectively, when they have about the same computational complexity and $\tau_{max} = 20$ is set. Since the complexity of the proposed method is about half that of existing methods as shown in Figure 2, we average the proposed metric over two pairs of the synchronization signals in (25) so that the proposed IFO and SI detection scheme has approximately the same complexity as the conventional RCJD scheme when $M = 3$. It is immediately seen that the performance of the proposed detection method is almost the same independent of the CP type. Similar to the results in Figures 4 and 5, the proposed and conventional schemes have experienced some performance degradations as well,

however the performance of the conventional scheme is worse because it is affected by the channel variation in the frequency domain. As expected, it is evident that the performance of the proposed sequential detection scheme using two symbol averaging is dramatically improved with increase in SNR, compared to that of the RCJD scheme. The superiority of the proposed scheme over existing methods is still observed in the case when residual STO is present. More specifically, the SNR gain of the proposed scheme over the RCJD scheme is more than 5 dB for a target estimation error of 0.1% or less, regardless of the channel model. Such observation allows us to conclude that the proposed joint detection method is better suited for situations where there is an unknown STO or high frequency selectivity than the conventional detection methods.

6. Conclusions

To ensure a reliable synchronization in LTE-D2D, it is necessary to perform joint estimation of the IFO and SI with reduced complexity and high performance. To address this issue, a computationally effective joint detection method was proposed in the LTE-D2D system. Exploiting the correlation between the received PSSS and SSSS sequences, the proposed detection scheme was designed to remove the impact of symbol timing error on the synchronization performance. The inherent feature of the PSSS and SSSS was used to perform sequential detection without any information on the transmitted synchronization signals. Simulation results have illustrated that the proposed sequential detection scheme is robust to symbol timing error in multipath frequency-selective fading channel and its performance is superior to the conventional methods.

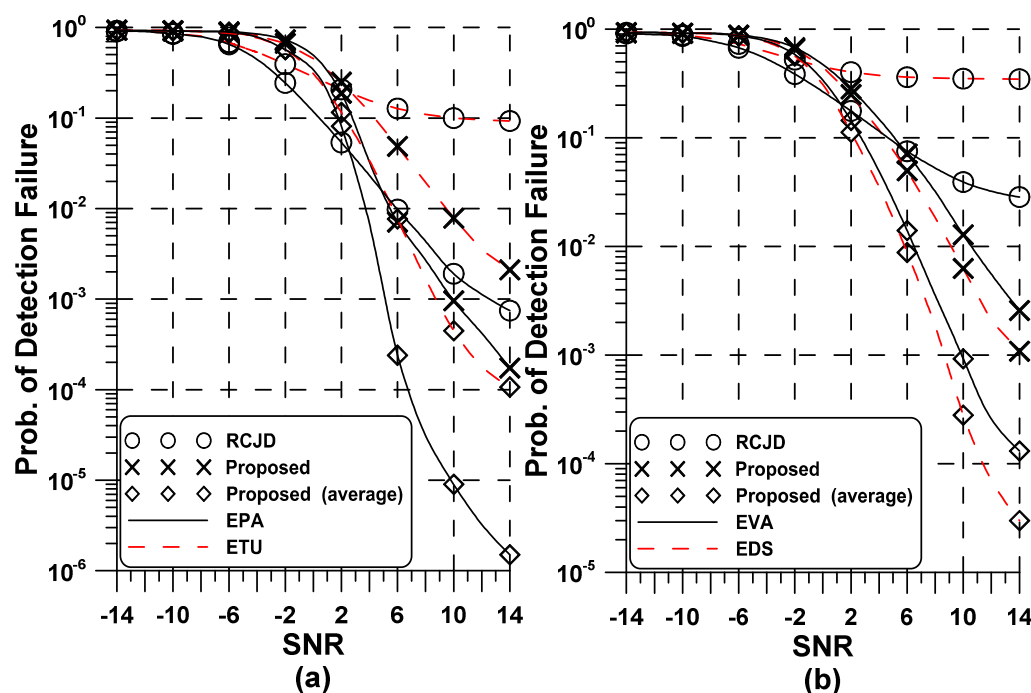


Figure 6. Performance of the proposed and reduced-complexity joint detection (RCJD) schemes using extended cyclic prefix (CP) versus SNR when the same complexity is considered.

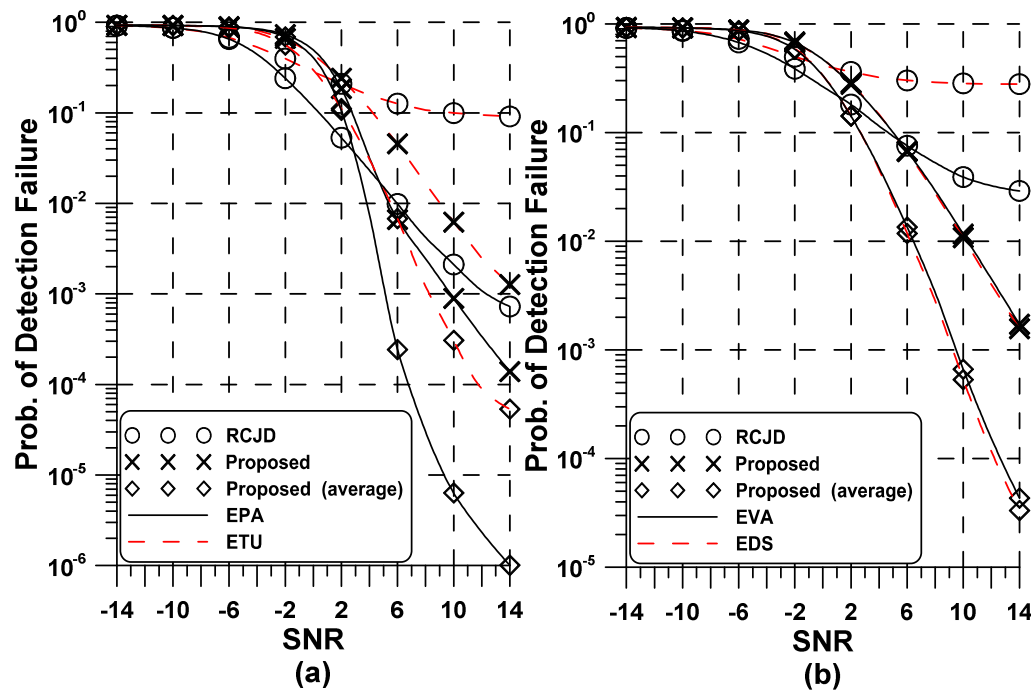


Figure 7. Performance of the proposed and RCJD schemes using normal CP versus SNR when the same complexity is considered.

Author Contributions: Y.-A.J. and Y.-H.Y. designed the proposed estimation algorithm. Y.-A.J. supplemented the proposed algorithm. H.-K.S. complemented the problems of the proposed scheme and suggested solutions. Y.-H.Y. gave feedbacks about an enhanced algorithm and all simulation results. All authors have read and agreed to the published version of the manuscript.

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