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Nonlinearity-Based Ranging Technique in SC-FDE Communication System With Oversampled Signals

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ABSTRACT Ground-based positioning systems are necessary as the conventional satellite-based systems suffer from weak received signals. In this paper, we proposed a precise ranging method in single-carrier frequency domain equalization (SC-FDE) system using the amplitude nonlinearity of oversampled signals. A new pattern of the preamble and the unique word was designed for SC-FDE system, where the preamble can be exploited in correlation-based timing algorithm to obtain impulse-like timing metric. Combining with the coarse timing procedure, we proposed a fine ranging method relying on the oversampled signals in SC-FDE receiver employing Qth-power nonlinearity, and then analyzed its estimation mean and variance. The extensive simulations were conducted to validate the proposed method with distinct modulation schemes, rolling-off factors, block lengths, and nonlinearity factors. The results show that the proposed ranging method can achieve unbiased estimate and its root mean square errors will reach the order of centimeter at medium-to-high signal-to-noise ratio region in flat-fading channels, whereas the observed performance degradation in frequency selective channel can be mitigated by using equalized oversampled signals.

INDEX TERMS Ranging, SC-FDE, oversampled signals, timing synchronization.

I. INTRODUCTION

Location schemes have been adopted in the available wireless networks to implement the promising location based services [1]. Different wireless networks can benefit from location information, such as cellular networks [2], wireless sensor networks [3], underwater acoustic sensor network [4], and ad-hoc networks [5], as overviewed in [6] and [1]. The localization accuracy and efficiency demand is ever increasing. Existing localization methods include global positioning system (GPS), beacon (or anchor) nodes, and proximity-based localization [7]. When conventional satellite-based systems have limitations where the GPS signal is weak or unavailable, ground-based positioning systems can be adopted in the context of wireless emergency services [8], tactical military operations [9], and various other applications offering location based services [10].

The ranging-based localization method infers location information taking advantage of a distance-dependent parameter using radio signals. The methods of received signal strength (RSS) [11], time of arrival (TOA) [1], time difference of arrival (TDOA) [12], frequency difference of arrival (FDOA) [13], and angle of arrival (AOA) [14], [15] are the most popular types of measurements. The indoor location systems for the IEEE 802.11 WLAN standard are based on received signal strength indicator (RSSI) mapping and fingerprinting, which suffer from limited accuracy, time consuming and vehicle movement [4]. The estimation of the TDOAs in a location system is based on the differences of the TOA or the differences in the arrival times of the signals at multiple receivers [16]. Localization schemes based on TOA or TDOA offer high precision, but this comes at the cost of a very complex process of accurate time synchronization among all users compared to round-trip delay (RTD). There are several methods to obtain distance information from a delay estimate, such as TOA, RTD, or any other form of so-called multiway TOA ranging schemes. In this paper, we use the propagation delay-based ranging.

Ultra-wide bandwidth–based systems are commonly used in the localization community with chirp spread spectrum or direct sequence spread spectrum [17]. Multicarrier transmission has become popular and widely used within the

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last two decades for broadband communication to combat time-dispersive fading channels [18]. The orthogonal frequency division multiplexing (OFDM) system makes use of a cyclic prefix (CP) and a one tap per subchannel equalizer [19]. The single-carrier frequency domain equalization (SC-FDE) techniques also have been popularly applied to many modern wireless communication systems, such as IEEE 802.11.ad and IEEE 802.16 Wimax. The data format of an SC-FDE system is generally composed of data blocks, each preceded with a CP such that low-complexity and accurate frequency-domain channel equalization can be effectively applied. Major advantages of SC-FDE over OFDM are its lower peak-to-average power ratio (PAPR) and less sensitivity to carrier-frequency offset (CFO) [20]. Traditional pseudonoise (PN) ranging methods suffer from limitation of the chip rate of the PN code [21], whereas if the TDOA estimation algorithms are directly adopted in the OFDM based networks, the precision would be limited by the sampling interval. The properties of OFDM in time and frequency domains can be employed to estimate the parameters of the transmission time delay [22]. To the best of the authors' knowledge, there is no existing work that presents ranging method for SC-FDE signals. In this paper, we investigate the high-precision ranging method in SC-FDE system. While this paper does not specifically address the problem of accurate localization in multipath environments, the proposed approach enhances the ranging resolution and can be combined with any of the known techniques to provide an enhanced localization performance.

The main contributions of this paper are as follows.

- We propose a new pattern of preamble and unique words (UWs) for SC-FDE system. In receiver, the preamble is exploited in timing synchronization using cross-correlations and autocorrelations, which exhibit a desirable impulse-like timing metric.
- We propose a novel fine ranging method relying on oversampled signals in SC-FDE receiver employing Qth-power nonlinearity, and analyze its estimation mean and variance for second power nonlinearity. The corresponding Cramér-Rao lower bound (CRLB) is also given as a performance benchmark.
- Extensive simulations are also conducted to validate the proposed ranging method with distinct modulation schemes, rolling-off factors, block lengths, and nonlinearity factors Q. Results show that, the proposed method can achieve unbiased estimate and its root mean square errors (RMSEs) reach the order of centimeter at medium-to-high signal-to-noise ratio (SNR) region.

The rest of this paper is organized as follows. Section II overviews the related work in the literature. Section III introduces the SC-FDE signal models, where we propose a new preamble pattern for SC-FDE and design the corresponding demodulation architecture in receiver. In Section IV, we propose a ranging method based on oversampled SC-FDE signals and analyze the performance. The extensive simulations will be carried out to validate our novel ranging method in Section V. Finally, conclusions are drawn in Section VI.

II. RELATED WORK

A TDOA location scheme for the OFDM based wireless metropolitan area networks (WMANs) was proposed in [12], where the TDOA estimation algorithm enhances the location performance by utilizing the information in the time and frequency domains obtained from the received location OFDM signals. The cross-correlation function between the local sequence and the received sampling sequence is employed as the cost function. Some properties of the OFDM signals in both time and frequency domains are utilized in the super-resolution algorithms, e.g. the maximum likelihood algorithm and the multiple-signal classification (MUSIC) algorithm [23]. However, the super-resolution estimations are embarrassed with high complexity and some needs for priori knowledge, albeit their impressive precisions.

Jung *et al.* [24] proposed a two-phase ranging method based on the decentralized timing-offset estimation in an OFDMA system with a half-duplex amplify-and-forward relay stations. For robotic swarms with high relative mobility, Staudinger *et al.* [17] designed the subcarrier-interleaved OFDMA using a decentralized TDMA reservation scheme, jointly with round-trip delay ranging at high update rates. Dai *et al.* [25] used the property of the OFDM signal itself in the frequency domain with respect to transmission delay, resulting in the accurate time of arrival estimation in the order of centimeters, close to the CRLB in the range of medium-to-high SNR. In [22], OFDM coarse estimation and fine estimation in time and frequency domains were employed respectively, to estimate the parameters of the transmission time delay. It can be used in applications with large range and high accuracy for space. It was then enhanced in [26] using the classical MUSIC algorithm [27] in the frequency domain to estimate the fractional part of the round-trip delay, including the transmission delay and the inherent delay of the repeater. In the flying ad hoc network (FANET) framework, a ranging algorithm using OFDM frequency-domain signals was investigated in [28] jointly with timing metric computation based on the correlation of the OFDM preambles, where the derived estimation variance approached the CRB.

III. SC-FDE SIGNAL MODEL

The time-domain adaptive equalizer are mainly one or more transversal filters for which the number of adaptive tap coefficients is on the order of the number of data symbols spanned by the multipath, with at least several hundred multiplication operations per data symbol [29]. However, the SC-FDE technique has similar performance, efficiency, and low signal processing complexity advantages as OFDM [30]. In this section, we commence with a novel preamble pattern designed for SC-FDE system. Then, we describe the demodulator architecture and present a correlation-based timing algorithm based on single preamble sequence.

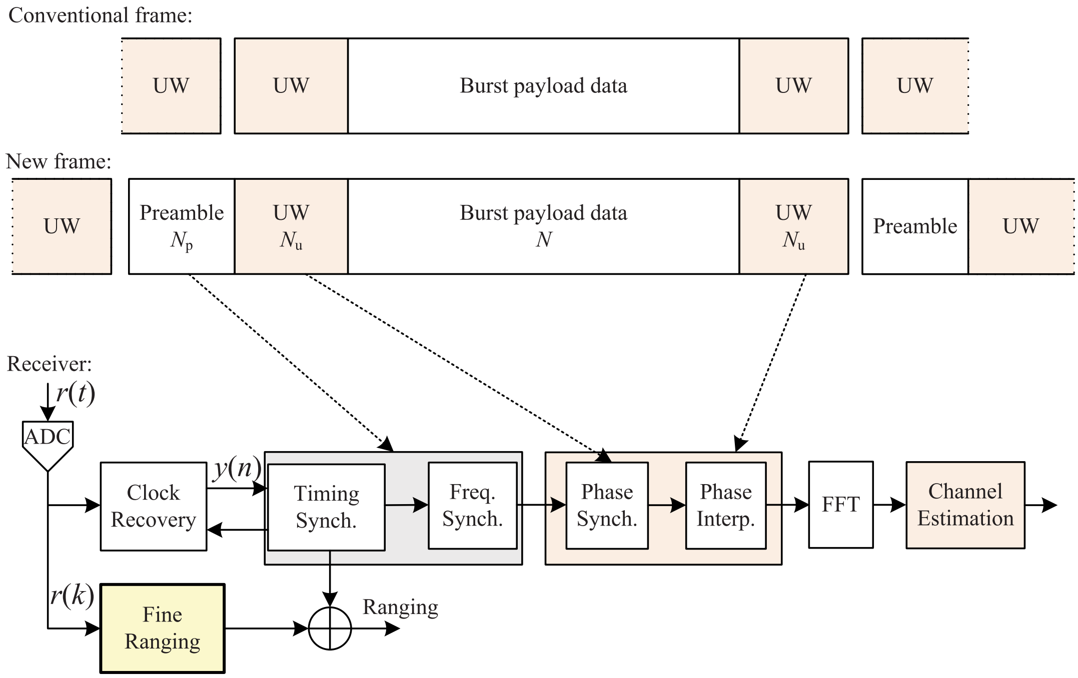

FIGURE 1. Receiver architecture of SC-FDE with new preamble pattern.

A. PREAMBLE PATTERN FOR SC-FDE

The data to be transmitted in SC-FDE system is divided into symbol blocks with identical length. One of its major differences with conventional SC systems is the insertion of CP at the transmitter side as well as the CP removal at the receiver side. The transmitted training symbols, known as UW, should have equal or nearly equal magnitude for all frequencies in frequency domain. Two or more back-to-back unique words, inserted within a data sequence, are used for synchronization and equalizer training purposes.

The UW following the data segment serves as a cyclic prefix for the succeeding UW. In general, the SC-FDE pilot blocks should be at least twice as long as the expected maximum channel impulse response length [31]. Frequency domain interpolation is necessary for the equalizer design of data block. The m-th block consists of N symbols under different modulation formats such as M-ary phase-shift keying (MPSK) or M-ary quadrature amplitude modulation (MQAM) and is given as

$$s_m = [s_{m,0}, s_{m,1}, \ldots, s_{m,N-1}]^T, \tag{1}$$

where the $s_{m,n}$ are the complex valued transmitted symbols with mean power 1.

The predefined UW sequence is inserted and served as a reference sequence to ensure the estimation accuracy. The transmitted signal block after adding UW symbols is represented as

$$s_m = [s_{m,-N_u}, \ldots, s_{m,-1}, s_{m,0}, s_{m,1}, \ldots, s_{m,N-1}, \ldots, s_{m,N+N_u-1}]^T, \tag{2}$$

where N_u is the symbol length of UW. The first of these back-to-back unique words acts as a cyclic prefix, and absorbs any inter symbol interference from previous data. The second and subsequent words form a periodic sequence that have the ideal periodic autocorrelation property.

To obtain an impulsive timing metric at receiver, we design a new simple preamble for SC-FDE as shown in Fig. 1, having the following form

$$s_m = [s_{m,-N_p-N_u}, \ldots, s_{m,-1}, s_{m,0}, s_{m,1}, \ldots, s_{m,N-1}, \ldots, s_{m,N+N_u-1}]^T, \tag{3}$$

where an extra unique word with symbol length N_p is added as preamble sequence. The UW also possesses the constant amplitude zero auto-correlation (CAZAC) property [32]. Therefore, the overhead fraction is $(2N_u+N_p)/(N+2N_u+N_p)$.

In both transmitter and receiver, a root raised cosine (RRC) filter is used to form a raised cosine filter. The received equivalent base-band signal can be written as

$$r(t) = \sum_{n=-\infty}^{\infty} s_n g(t - nT - \tau T) + w(t), \tag{4}$$

where s_n are transmitted symbols, T is the symbol duration, τT is an unknown round-trip time delay, and $w(t)$ is the additive white and Gaussian noise (AWGN) with power density N_0, and $g(t)$ is the raised cosine filter with frequency spectrum $G(f)$ [33], which is a function of roll-off factor α and bandwidth B. The oversampled equivalent digital base-band signal after matching filter can be written as

$$r(k) = \sum_{n=-\infty}^{\infty} s_n g(kT/L - nT - \tau T) + w(kT/L), \quad (5)$$

where L is the oversampling factor, $\tau T = \tau_I T + \tau_F T$, and the $\tau_I T$, $\tau_F T$ herein represent respectively the integer and fractional-valued unknown symbol timing offset (STO).

B. SC-FDE RECEIVER ARCHITECTURE

Many synchronization methods have been developed to estimate timing offsets for OFDM systems [34], [35]. One of the best known pilot-aided timing algorithms is Schmidl's method exploiting a preamble containing two identical halves in the time-domain [34]. Similarly, the counterpart in SC-FDE takes the conjugate product between two successively received UW blocks, whereas the timing metric plateau inherent in this method causes a large variance of the timing estimate at low SNRs [36]. Compared to OFDM systems, the synchronization techniques for SC-FDE systems are much less examined in the literature. The frequency-domain estimation was investigated using the conjugate product of two consecutive pilot sub-blocks in [37], and a blind synchronization was designed by employing decision-feedback structure to obtain the reference signal in [20].

Based on the receiver architecture of the SC-FDE system in Fig. 1, we present the following synchronization and estimation procedures for SC-FDE receivers:

1) Sample the received SC-FDE signals $r(t)$ with oversampling factor L, and the baud rate symbols $y(n)$ can be obtained after clock recovery;
2) Find the start position d of preamble sequence in $y(n)$ by timing synchronization, and estimate the frequency offsets with local preambles sequence $p(n) \in \{s_{m,-N_p-N_u}, \ldots, s_{m,-N_u-1}\}$.
3) Compute the range between terminal and target as $D = C(\hat{\tau}_I + \hat{\tau}_F)$, where C is the speed of light, using the integer-valued time delay estimation $\hat{\tau}_I$ and the fine estimation $\hat{\tau}_F$ we would provide in Section IV.
4) Estimate phase offsets of UW sequences and get the phase offsets of burst payload data using interpolation between the first UW and the second one in the time-domain.
5) Estimate the radio channel based on UW sequences and then equalize the burst payload data using known frequency-domain equalization methods [38], [39].

C. TIMING BASED ON CORRELATIONS OF PREAMBLE

We herein propose the timing algorithm employed in step-2 mentioned above, with integer-valued time delay estimation $\hat{\tau}_I = \arg\max\{R(d)\}$. The timing metric can be written in (6), as shown at the bottom of this page, where d_0 is designed parameter for the correlations, $1 \leq d_0 \leq d_T$, d_T is the maximum length of correlations. The correlations used for computing timing metric include the cross-correlation between $y(n)$ and local preambles sequence $p(n)$, $R_N(d) = y(n + d)p^*(d)$, and the autocorrelation of $R_N(d)$. The denominator term in (6) denotes normalizing this correlation in numerator term by received total preamble power of m-th block.

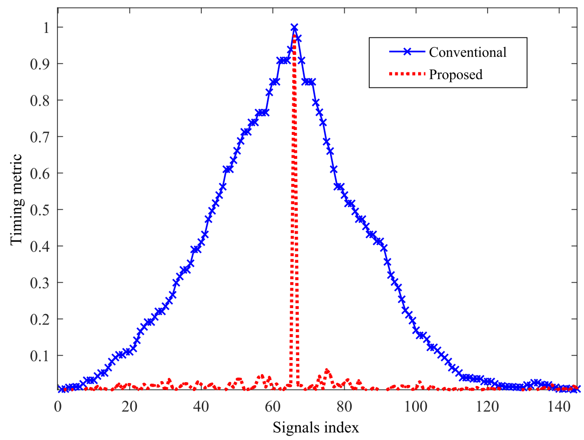

FIGURE 2. Timing metrics of proposed timing algorithm with preamble pattern (3) and conventional method [34] for with QPSK mapping, $N_p = 64$, and $\alpha = 0.35$.

This timing metric (6) is a function of the length of this inserted new preamble N_p. Following the proposed timing algorithm, we present the timing metric by simulation, as shown in Fig. 2, for the new preamble pattern in (3) with QPSK mapping, $N_p = 64$, $\alpha = 0.35$. Compared with the conventional method using correlations between two successive UW blocks [34], the proposed method exhibits impulsive peak without ambiguity among the received preamble sequence. Furthermore, the autocorrelation between $R_N(d)$ and $R_N^*(d - d_0)$ will be tolerant to frequency offset.

In order to detect the impulsive peaks of timing metric in implementation, we can set the detection threshold for $\hat{\tau}_I$ according to the required detection and false alarm probability. Finally, we arrive at the estimation of integer-valued timing offset, and thereby the coarse range estimation between mobile station and target, $D_I = C \times \tau_I$. In the following text,

$$R(d) = \frac{\frac{1}{d_T} \sum_{d_0=1}^{d_T} \frac{N_p}{N_p - d_0} \left| \sum_{n=d_0}^{N_p-1} y^*(n + d - d_0) y(n + d) p(d - d_0) p^*(d) \right|^2}{\left(\sum_{n=1}^{N_p} |y(n + d)|^2 \right)^2} \quad (6)$$

we devote to study the fine estimation of D_F using payload segment to improve the ranging resolution.

IV. FINE RANGING WITH OVERSAMPLED SC-FDE SIGNALS

In this section, we propose a novel fine ranging method based on Qth-power nonlinearity of amplitude of oversampled SC-FDE signals. Then, we derive the estimation performance.

A. RANGING BASED ON NONLINEARITY

The oversampled signals $r(k)$ (5) in SC-FDE receiver go through Qth-power nonlinearity and then we obtain

$$x(k) = \left| \sum_{n=-\infty}^{\infty} s_n g(kT/L - nT - \tau_F T) + w(kT/L) \right|^Q, \tag{7}$$

where $\otimes$ denotes convolution operation, and $g(t) = g_T(t) \otimes g_R(t)$. These samples of Qth-power input signals contain a spectral component at $1/T$. After computing the complex Fourier coefficient at the symbol rate, we can determine this spectral component,

$$X(m) = \sum_{k=mLN}^{(m+1)LN-1} x(k) e^{-j2\pi k/L} \tag{8}$$

where the normalized phase $\hat{\tau}_F = -\arg\{X(m)/(2\pi)\}$ of this coefficient is an unbiased estimate. Therefore, we propose the range estimation (9), as shown at the bottom of this page, where the oversampling rate L used for ranging should be set by keeping spectral component $1/T$ in (7), usually with $L = 4$.

B. PERFORMANCE ANALYSIS

In the following text, we analyze the statistics of the fine range estimation (9). To simplify the analysis, the statistics for Qth-power amplitude of oversampled signals are presented with secondary power.

1) ESTIMATION MEAN

In the case of $|\arg\{X\}| \ll \pi$, we have an approximation by linearizing the arg-operation,

$$\begin{aligned} E[\hat{D}_F] &= \frac{-C}{2\pi} \arg\{E[X(m)]\} \\ &= \frac{-C}{2\pi} \arg\left\{ \sum_{k=0}^{LN-1} E[x(k)] e^{-j2\pi k/L} \right\}, \end{aligned} \tag{10}$$

where the expectation $E[X(m)]$ is computed with respect to the joint distribution of transmitted symbols s_n and the noise term $w(k)$, which are independent of each other. To compute the expectation, we omit the cross term in (10) and get the remaining terms in (11), as shown at the bottom of this page. Then, we update the range estimation (10) as

$$\begin{aligned} E[\hat{D}_F] &= \frac{-C}{2\pi} \arg\left\{ \frac{LN}{T} P_0(1/T) e^{-j2\pi\tau_F} \right\}, \\ E[X(m)] &= \frac{LN}{T} P_0(1/T) e^{-j2\pi\tau_F}, \end{aligned} \tag{12}$$

where $P_0(f)$ can be written as the function of Fourier transform of the matched-filtered $G(f)$. In addition, $P_n(f)$ can be expressed as (14).

$$P_0(f) = \mathcal{F}\left\{ |g(t)|^2 \right\} |_{f=1/T} = G(f) \otimes G(f), \tag{13}$$

$$\begin{aligned} P_n(f) &= \mathcal{F}\left\{ g(t) g^*(t-nT) \right\} | \\ &= G(f) \otimes G^*(-f) e^{-j2\pi f nT}, \end{aligned} \tag{14}$$

where $\mathcal{F}\{\cdot\}$ denotes Fourier transformation.

Therefore, we have the unbiased estimate of $E[\hat{D}_F] = C \times \tau_F$ under the condition $\arg\{P_0(1/T)\} = 0$, which holds true for transceivers using RRC shaping filters.

2) ESTIMATION VARIANCE

Next, we will determine the variance of ranging variable $\hat{D}_F$, and then we have

$$\begin{aligned} \mathrm{Var}[\hat{D}_F] &= E[\hat{D}_F^2] - (E[\hat{D}_F])^2 \\ &= \frac{C^2}{(2\pi)^2} E\left[(\arg X(m))^2 \right]. \end{aligned} \tag{15}$$

Bearing in mind that the imaginary part of X has zero mean, $E[\Im\{X(m)\}] = 0$, and the real part has expectation as $E[\Re\{X(m)\}] = E[X(m)]$, we make an approximation for high SNR and small τ_F,

$$\begin{aligned} \mathrm{Var}[\hat{D}_F] &= \frac{C^2}{(2\pi)^2} E\left[\frac{(\Im\{X(m)\})^2}{(\Re\{X(m)\})^2} \right] \\ &\approx \frac{C^2}{(2\pi)^2} \frac{E\left[(\Im\{X(m)\})^2 \right]}{(E[\Re\{X(m)\}])^2}. \end{aligned} \tag{16}$$

The variance in (16) is given in (17), as shown at the bottom of the next page. After some direct manipulation, for

$$\hat{D}_F = -\frac{C}{2\pi} \arg\left\{ \sum_{k=mLN}^{(m+1)LN-1} \left| \sum_{n=-\infty}^{\infty} s_n g(kT/L - nT - \tau_F T) + w(kT/L) \right|^Q e^{-j2\pi k/L} \right\}, \tag{9}$$

$$\begin{aligned} E[x(k)] &= E\left[\left| \sum_{n=-\infty}^{\infty} s_n g(kT/L - nT - \tau_F T) \right|^2 \right] + E\left[|w(kT/L)|^2 \right] \\ &= \sum_{n=-\infty}^{\infty} |g(kT/L - nT - \tau_F T)|^2 + \sigma^2 \end{aligned} \tag{11}$$

transceivers using RRC shaping filters, we have

$$\mathrm{Var}[\hat{D}_F] = \frac{C^2}{(2\pi)^2 L} \frac{\sum_m (\Im\{P_m(1/T)\})^2}{(P_0(1/T))^2} = 0, \quad \text{with } \Im\{P_m(1/T)\} = 0. \tag{18}$$

Therefore, the proposed range estimate has zero variance for high SNRs. We also use CRB to compare analytically the performance of proposed range estimation method for different Q in the next section. The CRB states the variance of range estimates and it is lower bounded as,

$$\mathrm{Var}[\hat{D}_F - D_F] \geq \frac{C^2}{2NE_s/N_0} \frac{\int_{-\infty}^{\infty} G(f)df}{\int_{-\infty}^{\infty} 4\pi^2 f^2 G(f)df} \doteq CRB. \tag{19}$$

Therefore, for the raised-cosine pulse, the CRB can be derived as,

$$CRB = \frac{C^2T^2}{2NE_s/N_0} \frac{1}{3(\pi^2-8)\alpha^2+\pi^2}. \tag{20}$$

V. NUMERICAL RESULTS AND DISCUSSIONS

In the following section, we investigate the performance of proposed ranging method for SC-FDE system in terms of estimation bias and RMSE performance. An SC-FDE system with BPSK, QPSK, 16PSK and 16QAM modulations, nonlinearity factors of $Q = 0.5, 1, 1.5, 2$, rolling off factors $\alpha = 0.1, 0.2, 0.25, 0.35$, data block length of $N = 256, 512, 1024$, UW length of $N_u = 64$, channel bandwidth 10 MHz has been considered with extensive simulations in flat-fading channel to validate our proposed ranging method.

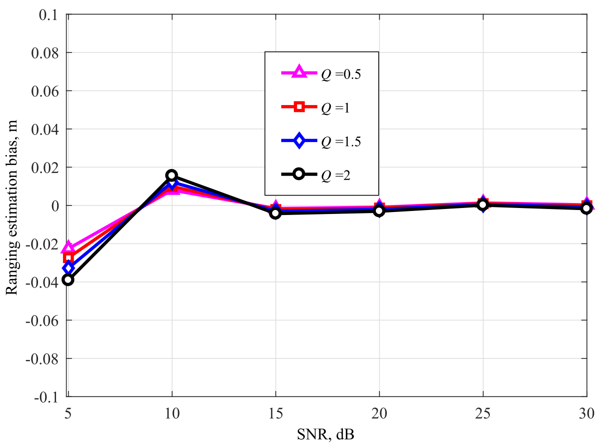

FIGURE 3. Bias of proposed ranging method with $Q = 0.5, 1, 1.5$, and 2, QPSK modulation, $N = 1024$, and $\alpha = 0.35$.

Figure 3 presents the estimation bias performance of proposed ranging method under different $Q = 0.5, 1, 1.5, 2$. We can find that, for medium to high SNRs, all the estimates will be unbiased ones. Although the smaller Q will outperform the higher ones at low SNR region, we note that their differences are only 2 cm at SNR of 5 dB, which can be omitted.

In order to evaluate estimation performance, RMSE is adopted as the comparison measure with CRB as a benchmark.

$$\sigma_{D,CRB} = \frac{CT}{\sqrt{2NE_s/N_0}} \frac{1}{\sqrt{3(\pi^2-8)\alpha^2+\pi^2}} \tag{21}$$

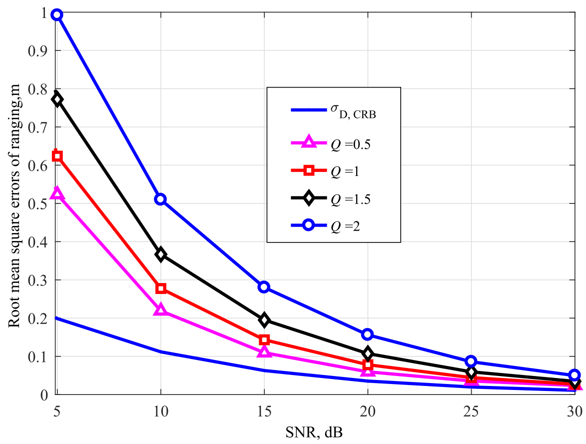

FIGURE 4. RMSE of proposed ranging method with $Q = 0.5, 1, 1.5$, and 2, QPSK modulation, $N = 1024$, and $\alpha = 0.35$.

Figure 4 depicts RMSEs of proposed ranging method under different SNRs and Q factors, and N and α are fixed at 1024 and 0.35, respectively. It can be find that, the increasing Q will introduce a degradation for estimation jitter especially at low SNRs. However, for high SNRs, the proposed ranging method with $Q = 0.5, 1, 1.5, 2$ will perform close to CRB. For $Q = 1, 2$, the ranging jitter will reach below 1 cm at SNR $>$ 18 dB, 24 dB, respectively.

In addition, similar simulations are also conducted to compare the RMSE performance of BPSK, 16PSK and 16QAM modulations, under the conditions of Q = 0.5, 1, 1.5, 2, as illustrated in Fig. 5 and Fig. 6. For lower-order modulation BPSK, the RMSE performance will get better than that of QPSK, whereas for the higher-order modulation 16QAM it will degrade. For $Q = 2$, the ranging jitter of BPSK and 16QAM will also reach below 1 cm at SNR $>$ 24 dB. However, for a decreasing $Q = 0.5, 1$, the ranging jitters of 16QAM will encounter a floor of 0.2 m, 0.15 m, respectively, at SNR = 30 dB, whereas there is little gap between that of BPSK and 16PSK. Therefore, the proposed ranging method performs preferably for MPSK and small Q factors.

$$E\left[(\Im\{X(m)\})^2\right] = E\left[\left(\Im\left\{\sum_{k=0}^{LN-1}\left|\sum_{n=-\infty}^{\infty} s_n g(kT/L - nT)\right|^2 e^{-j2\pi k/L}\right\}\right)^2\right]. \tag{17}$$

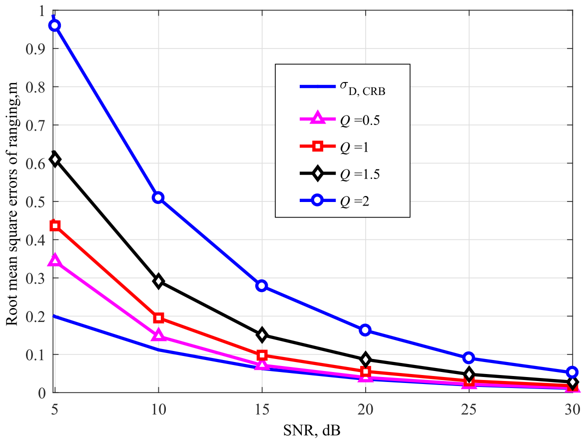

FIGURE 5. RMSE of proposed ranging method with $Q = 0.5$, 1, 1.5, and 2, $N = 1024$, BPSK modulation.

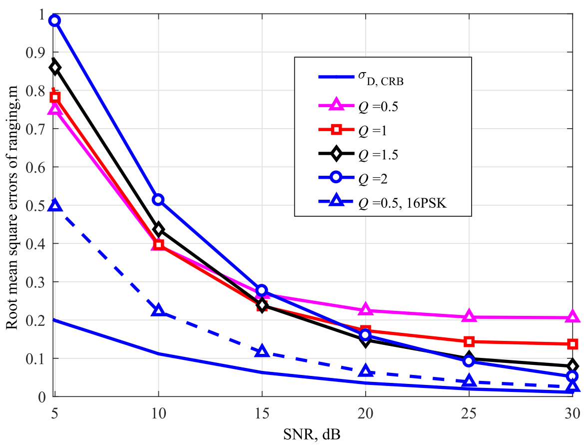

FIGURE 6. RMSE of proposed ranging method with $Q = 0.5$, 1, 1.5, and 2, $N = 1024$, 16QAM, and 16PSK.

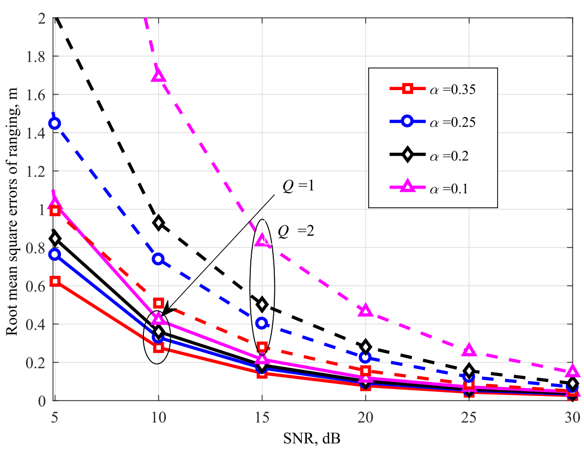

FIGURE 7. RMSE of proposed ranging method with $\alpha = 0.1$, 0.2, 0.25, and 0.35, $Q = 1$, and 2, QPSK modulation, and $N = 1024$.

Next, range estimation performances for $Q = 1, 2$, versus SNR as well as $\alpha = 0.1, 0.2, 0.25, 0.35$, are provided in Fig. 7. We find that, the increasing rolling factors will improve the estimation jitters for both $Q = 1$ and 2. Furthermore, for

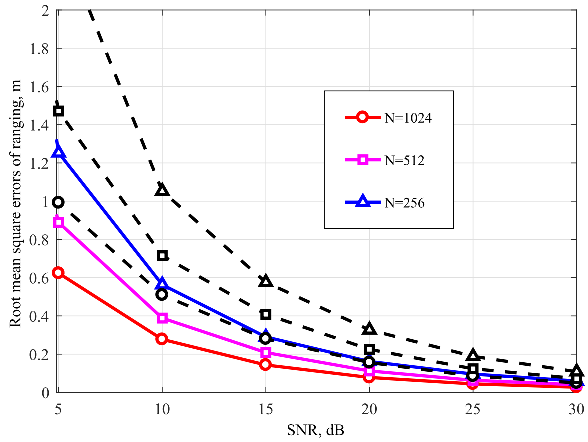

FIGURE 8. RMSE of proposed ranging method versus $N = 1024$, 512, and 256, $\alpha = 0.35$, $Q = 1$, and 2, and QPSK modulation.

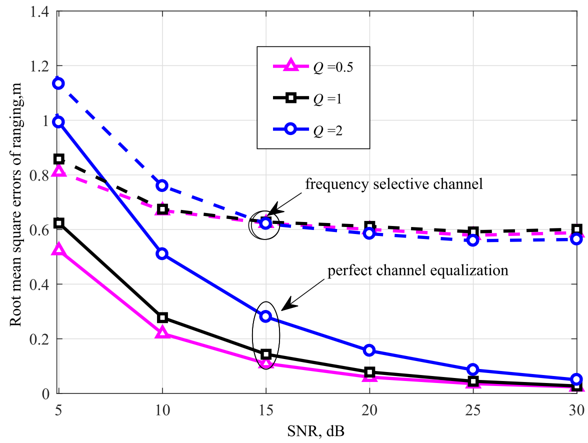

FIGURE 9. RMSE of proposed ranging method under frequency selective channel with $N = 1024$, $\alpha = 0.35$, $Q = 0.5$, 1, and 2, and QPSK modulation.

$Q = 2$, the enhancements for increasing α are steeper than those of $Q = 1$. The proposed ranging method with $Q = 1$ performs well enough with little potential improvement.

Comparisons of estimation performance versus different length of data blocks N are illustrated in Fig. 8, for $\alpha = 0.35$, $Q = 1, 2$, and QPSK modulation. Results show that the proposed ranging method will perform better with more signals per section. For $N = 256, 512, 1024$, the ranging jitters under $Q = 1$ will get below 1 cm at SNR > 18 dB, 15 dB, 13 dB, respectively, whereas the jitters under $Q = 2$ will get below 1 cm at SNR > 25 dB, 21 dB, 18 dB, respectively.

Finally, we examine the RMSE performance of proposed ranging method in frequency selective channel for $\alpha = 0.35$, $N = 1024$, $Q = 0.5, 1, 2$, and QPSK modulation, as shown in Fig. 9. The Rician fading channel is modelled as 10 paths with path delays τ_k of $k = 0, 2, 4, \ldots, 20$ samples, exponentially decaying path gains of $e^{-\tau_k/3}$ and K factor of 20 dB [35]. We find performance degradation in this frequency selective channel with a floor of 0.6 m for $Q = 0.5, 1, 2$. Accurate channel estimate and powerful frequency-domain equalization will improve the RMSE performance, such as the iterative block decision feedback equalizer (IB-DFE) [31], [40]. It is noted that the proposed method will achieve desired performance as it is preceded by perfect channel equalization for oversampled signals in SC-FDE receivers.

As for computational complexity, the coarse range estimation needs no additional computation taking advantage of the timing synchronization module (6), whereas the proposed fine ranging method (9) can be simplified as (22) with only multiplication and addition for $L = 4$ and $Q = 2$.

$$\begin{aligned}\hat{D}_F &= -\frac{C}{2\pi}\arg\left\{\sum_{k=0}^{N-1}\sum_{l=0}^{L-1}|r(Lk+l)|^Q e^{-j2\pi l/L}\right\}\\ &= -\frac{C}{2\pi}\arg\left\{\sum_{k=0}^{N-1}|r(4k)|^2 - j|r(4k+1)|^2\right.\\ &\qquad \left. - |r(4k+2)|^2 + j|r(4k+3)|^2\right\}. \end{aligned} \tag{22}$$

For each SC-FDE block consisting of $4N$ samples, (22) requires $8N + 1$ real multiplications, $2(N - 1)$ additions and 1 phase arg operation, which impose low computational complexity in SC-FDE receivers.

VI. CONCLUSION

To obtain high-precision ranging estimation in SC-FDE system which exhibits both low PAPR and low-complexity equalization, we propose a novel ranging estimation method employing a generalized nonlinearity of the amplitude of oversampled signals. After reviewing the related work in OFDM system, we present a new pattern of preamble and UW for correlation-based timing synchronization. Then, we depict a receiver architecture and a timing algorithm with impulse-like timing metric. A novel fine ranging method is proposed and analyzed with unbiased estimate and zero variance for high SNRs. Simulation results demonstrate that the proposed method under different conditions can reach the order of centimeter at medium-to-high SNR region. Considering computation complexity, payload data section length of $N = 1024$ and Qth-power nonlinearity of $Q = 2$ are recommended in practical implementation.

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