High Accuracy Uplink Timing Synchronization for 5G NR in Unlicensed Spectrum

Yuhong Wang, Yonghong Zeng, *Fellow, IEEE*, Sumei Sun, *Fellow, IEEE*, and Yang Nan

***Abstract*—Sub-microsecond high accuracy clock synchronization is essential in time critical applications such as factory automation. In 3GPP, 5G New Radio in unlicensed spectrum (NR-U) is being discussed to support time critical applications. In 5G NR-U uplink, synchronization is achieved by using block interleaved frequency division multiple access (B-IFDMA) reference signals (RS). Based on the B-IFDMA-structured RS, conventional methods can only obtain the frequency domain channel at non-evenly spaced frequency points. When this channel is transformed from frequency domain to time domain, the resultant channel impulse response (CIR) will contain additional "pseudo" paths. In multipath channel, such "pseudo" paths may cause aliasing effects, affecting accuracy of CIR-based timing synchronization algorithms. In this letter, we propose two methods to address the above problem, namely, the frequency-domain subspace (FDS) based algorithm, and FDS-aided algorithm. Simulation result shows that our proposed methods counteract the aliasing effects effectively and significantly outperform the conventional algorithms.**

***Index Terms*—Timing synchronization, 5G NR unlicesed specturm, B-IFDMA.**

I. Introduction

SUB-MICROSECOND high accuracy clock synchronization is essential to time critical applications such as factory automation [1]. A typical use case in factory automation is multi-robot cooperation, where multiple robots collaborate and operate synchronously with less than 1 μs jitter. A jitter constraint of 1 μs requires that clock offset estimation error between user equipment (UE) and base station (BS) shall be within $[-500\ ns, 500\ ns]$. The de facto standard for clock synchronization is IEEE 1588, also known as precision time protocol (PTP) [2]. According to PTP standard, BS serves as the reference clock, and continuously exchanges timestamp information with UEs to synchronize their clocks. When a BS (UE) receives a packet, the BS (UE) needs to record the instant when the packet is received. In order to accurately timestamp the instant of packet arrival, it is very important for BS (UE) to estimate the time of arrival with high accuracy.

To support 5G new radio (NR) in unlicensed spectrum (NR-U), we must guarantee that the maximum power spectrum density (PSD) is under control. Thus, block interleaved frequency division multiple access (B-IFDMA) has been considered for NR-U uplink. In B-IFDMA, total subcarriers are divided into several groups of subcarriers, with each group of subcarriers referred to as an interlace. Each interlace consists of multiple, evenly-spaced blocks of adjacent subcarriers. B-IFDMA enables signal transmission over narrow bandwidth at high transmission power while complying with the PSD limit. For uplink timing synchronization in NR-U, each UE will transmit reference signal (RS) on assigned one or two interlaces, while other interlaces are occupied by other UEs. Either time domain [3], [4] or frequency domain [5], [6], [7] timing synchronization algorithms have been proposed. The time domain algorithm in [3], [4] correlates the received signal with a sequence which is the inverse fast Fourier transform (IFFT) of the partially known RS subcarriers, and finds the instant of the maximum correlation as the estimation of time of arrival. However, the performance of this algorithm is not good due to multipath interference and other user's interference.

The frequency domain algorithm in [5], [6], [7] estimates the channel impulse response (CIR) by transforming estimated effective channel transfer function (CTF) into time domain via IFFT, and finds the position that has the maximal amplitude in the estimated CIR. However, as shown in [7], this method cannot achieve high accuracy with B-IFDMA-structured RS. The reason is that the method can only obtain the frequency domain channel at certain non-evenly spaced frequency points specified in the B-IFDMA-structured RS. When such a channel is transformed from frequency domain to time domain via IFFT, the resultant CIR will contain many "pseudo" paths. In multipath channel, "pseudo" paths may cause aliasing effects. Aliasing effect will degrade the accuracy of timing estimation, thus high accuracy timing synchronization cannot be achieved with B-IFDMA-structured RS [7].

In this letter, we propose two methods to address the above problem. The first one is the frequency-domain subspace (FDS) based algorithm. Subspace algorithm is a well-known method for frequency and delay estimation, which is shown to have high accuracy [8]. To further enhance the estimation accuracy of FDS based algorithm, we propose two FDS-aided algorithms: FDS-MAX and FDS-AVG algorithms. FDS-MAX and FDS-AVG use FDS based algorithm to obtain initial timing offset estimation, which is used as reference to select the position corresponding to actual timing offset in the estimated CIR. Simulation results show that our proposed methods counteract the aliasing effect effectively and significantly outperform the conventional CIR-based algorithm [7] and time domain algorithm [3]. The rest of this letter is organized as follows. The B-IFDMA system model is introduced in Section II. The FDS based algorithm and FDS-aided algorithms are presented in Section III. Simulation results

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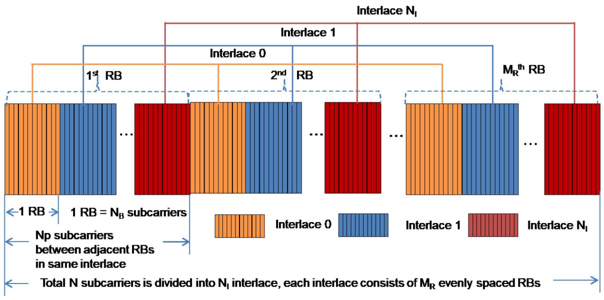

Fig. 1. Structure of interlace in B-IFDMA system model.

are given in Section IV. Finally, conclusions are drawn in Section V.

II. B-IFDMA System Model

A. B-IFDMA and Interlace Structure

In B-IFDMA system, total N subcarriers are used for transmission, with subcarrier spacing being Δf. N subcarriers are divided into N_I groups of subcarriers, with each group of subcarriers referred to as interlace. Each interlace consists of M_R evenly spaced blocks of adjacent subcarriers. We assume that each block consists of N_B adjacent subcarriers. In the following, a block of N_B adjacent subcarriers is referred as a resource block (RB). We assume that the space between adjacent RBs in one interlace is N_p. Fig. 1 illustrates the structure of interlace in B-IFDMA system.

Without loss of generality, we assume that one interlace has been allocated to a UE, and only one OFDM symbol is used to transmit the reference signal (RS). A total $M_R N_B$ known RS data symbols $\{D(m), m = 0, 1, \ldots, M_R N_B - 1\}$ from the UE are mapped to the subcarriers belonging to the allocated interlace according to

$$d_k = \begin{cases} D(i + jN_B), & k = k_0 + i + jN_p, \\ 0, & k \in \text{others} \end{cases} \tag{1}$$

where $i = 0, 1, \ldots, N_B - 1$, $j = 0, 1, \ldots, M_R - 1$, k_0 is a constant offset for the assigned interlace. The N data symbols $\{d_k, k = 0, .., N - 1\}$ are modulated onto N subcarriers via N_{FFT} point IFFT. To prevent inter-symbol interference (ISI), a guard interval consisting of N_g samples taken from the last N_g samples of IFFT output is put in front of the IFFT output.

B. Receiver Model

The signal is passed through a multipath fading channel with channel impulse response

$$h(\tau; t) = \sum_{l=0}^{L-1} h_l(t)\delta(\tau - \tau_l) \tag{2}$$

where L is the number of channel paths, $\{h_l(t)\}$ are the path gains, and $\{\tau_l\}$ are path delays. In the following, we assume path gains $\{h_l(t)\}$ are constant during the transmission of one OFDM symbol, and are denoted as $\{h_l\}$. The received signal is sampled with sampling interval $T_s = \frac{1}{\Delta f N_{\mathrm{FFT}}}$. The receiver then takes out every N_{FFT} samples to demodulate the data via fast Fourier transform (FFT). In the guard interval, there is a certain range that is not affected by previous symbol due to channel time dispersion [9]. As long as the FFT window starts in this range, the orthogonality between the different subcarriers will be maintained. Assume that the timing offset n_ϵ is within the ISI-free region of guard interval, that is $(-N_g + \lfloor \tau_{L-1}/T_s \rfloor) \le n_\epsilon \le 0$ (τ_{L-1} is the channel maximum delay, $\lfloor z \rfloor$ denotes taking largest integer less than z). For sake of analysis simplicity, we assume that $\{\tau_l\}$ are integer multiples of T_s. The FFT output of the OFDM symbol corresponding to the subcarriers of the allocated interlace can be written as,

$$Y(k) = e^{j2\pi(\frac{k}{N_{\mathrm{FFT}}})n_\epsilon} H(k)D(k) + w(k) \tag{3}$$

where, $k = k_0 + i + jN_p$, $i = 0, 1, \ldots, N_B - 1$, $j = 0, 1, \ldots, M_R - 1$ and

$$H(k) = \sum_{l=0}^{L-1} h_l e^{-j2\pi k \tau_l / T_u}, k = 0, \ldots, N_{\mathrm{FFT}} - 1 \tag{4}$$

is the CTF at frequency $f_k = k\Delta f$, and $T_u = 1/\Delta f$. $w(k)$ is the complex additive white Gaussian noise (AWGN). Dividing both end of (3) by $D(k)$, we get the estimated effective CTF $\check{H}(k)$ corresponding to the subcarriers of the allocated interlace

$$\check{H}(k) = \frac{Y(k)}{D(k)} = e^{j2\pi(\frac{k}{N_{\mathrm{FFT}}})n_\epsilon} H(k) + w(k)/D(k) \tag{5}$$

Note that $\check{H}(k)$ ($k = k_0 + i + jN_p$, $i = 0, 1, \ldots, N_B - 1$, $j = 0, 1, \ldots, M_R - 1$) is non-evenly spaced. $\check{H}(k)$ includes the effect of timing offset n_ϵ. When $n_\epsilon > 0$, the FFT window will contain samples from next OFDM symbol, leading to performance loss. Detailed analysis of performance loss due to $n_\epsilon > 0$ can be found in [10]. Based on $\check{H}(k)$, the joint ML estimation of n_ϵ, $\{h_l\}$ and $\{\tau_l\}$ can be written as,

$$[\hat{n}_\epsilon, \{\hat{h}_l\}, \{\hat{\tau}_l\}] = \arg \min_{[n_\epsilon, \{h_l\}, \{\tau_l\}]} \sum_{k \in \mathrm{P}} |\, \check{H}(k) - e^{j\frac{2\pi k n_\epsilon}{N_{\mathrm{FFT}}}} H(k) \,| \tag{6}$$

where $H(k)$ is defined in (4), which is a function of $\{h_l\}$ and $\{\tau_l\}$, P is the set of subcarrier indices where RS signals are located. Note that the implementation complexity of (6) is very huge, since a multiple dimension joint search for n_ϵ, $\{h_l\}$ and $\{\tau_l\}$ is needed. Obviously, complexity of ML algorithm is too high to be practically implemented. Thus we need to find low-complexity suboptimal algorithm.

III. Proposed FDS Based and FDS-Aided Algorithm for B-IFDMA System

A. FDS Based Algorithm

Substituting (4) into (5), we get

$$\check{H}(k) = \frac{Y(k)}{D(k)} = \sum_{l=0}^{L-1} h_l e^{\frac{-j2\pi k(\tau_l - n_\epsilon T_s)}{T_u}} + w(k)/D(k) \tag{7}$$

where, $\tau_l - n_\epsilon T_s$ is the actual relative arrival time for path l. For notation simplicity, let $\gamma_l = (\tau_l - n_\epsilon T_s)/T_u$, which is the scaled relative delay of path l. From (7), we see that the estimated effective CTF $\check{H}(k)$ is the superposition of L complex sinusoids with frequency γ_l. Note that the signal model here is different from the conventional model for subspace method [8]. The major difference is that $\check{H}(k)$ is only known at pre-allocated non-continuous subcarriers, while in conventional model the signals at all subcarriers are known. Thus, the conventional subspace method cannot be used here directly [8]. In the following, we propose a frequency domain subspace (FDS) based method to cater for the special signal model.

Define matrix $\mathbf{A}$ as $\mathbf{A} = [\mathbf{A}_0\ \mathbf{A}_1\ \cdots\ \mathbf{A}_{M_R-1}]$ with $\mathbf{A}_i = [\check{H}(k_0+jN_p)\ \check{H}(k_0+jN_p+1)\ \cdots \check{H}(k_0+jN_p+N_B-1)]^T$ being the frequency domain channel estimation of j-th RB ($j = 0, 1, \ldots, M_R - 1$). $\mathbf{A}$ is a matrix of size $N_B \times M_R$, which can be written as [8],

$$\mathbf{A} = \mathbf{BQ} + \mathbf{W} \tag{8}$$

where $\mathbf{B}$ is the steering matrix, and $\mathbf{W}$ is the noise sample matrix. $\mathbf{B}$ and $\mathbf{Q}$ can be written as,

$$\mathbf{B} = \begin{bmatrix} 1 & \cdots & 1 \\ e^{-j2\pi\gamma_0} & \cdots & e^{-j2\pi\gamma_{L-1}} \\ \vdots & \vdots & \vdots \\ e^{-j2\pi(N_B-1)\gamma_0} & \cdots & e^{-j2\pi(N_B-1)\gamma_{L-1}} \end{bmatrix} \tag{9}$$

$$\mathbf{Q} = \begin{bmatrix} h_0 e^{-j2\pi k_0\gamma_0} & \cdots & h_0 e^{-j2\pi\lambda\gamma_0} \\ \vdots & \ddots & \vdots \\ h_{L-1} e^{-j2\pi k_0\gamma_{L-1}} & \cdots & h_{L-1} e^{-j2\pi\lambda\gamma_{L-1}} \end{bmatrix} \tag{10}$$

where $\lambda = k_0 + (M_R - 1)N_p$. Assume that the number of paths L is smaller than M_R and N_B: $L < M_R$, $L < N_B$. Then the matrix $\mathbf{Q}$ is of full row rank. We can use singular value decomposition (SVD) to find the noise subspace and then estimate the delay of L_S ($L_S \leq L$) significant paths. Do the SVD of $\mathbf{A}$ such that

$$\mathbf{A} = \mathbf{U}\Sigma\mathbf{V}^H \tag{11}$$

where matrix $\mathbf{U}$ can be written as $\mathbf{U} = [\mathbf{U}_0\ \mathbf{U}_1\ \cdots \mathbf{U}_{N_B-1}]$. Let $\mathbf{\Phi} = [\mathbf{U}_{L_S}\ \mathbf{U}_{L_S+1} \cdots \mathbf{U}_{N_B-1}]$ be the matrix formed by last $N_B - L_S$ columns of matrix $\mathbf{U}$. Define $\mathbf{b}_m$ as steering vector,

$$\mathbf{b}_m = [1\ e^{-j2\pi m/N_{\mathrm{FFT}}}\ \cdots e^{-j2\pi m(N_B-1)/N_{\mathrm{FFT}}}]^H \tag{12}$$

Theoretically we can estimate the delay of L_S significant paths by finding L_S minimal values of $\mathbf{b}_m^H\mathbf{\Phi}\mathbf{\Phi}^H\mathbf{b}_m$. Since we are only interested in the delay of most significant path, we can set $L_S = 1$. Thus $\mathbf{\Phi} = [\mathbf{U}_1\ \mathbf{U}_2 \cdots \mathbf{U}_{N_B-1}]$. Then the relative arrival time of the most significant path can be estimated as:

$$\hat{\zeta} = \arg \min_{m\in[-\frac{M}{2},\frac{M}{2}]} \mathbf{b}_m^H\mathbf{\Phi}\mathbf{\Phi}^H\mathbf{b}_m \tag{13}$$

where M is the size of search range. Note that $L < M_R$ and $L < N_B$ is a condition for FDS-based algorithm to achieve good performance. If this condition is not satisfied, then performance of FDS based algorithm will be degraded to some extent. In the conventional subspace method, the signal is known at all subcarriers. Thus, the steering vector has a full size (N_{FFT}). In our case, the signal is only known at a few subcarriers, thus the steering vector is only of partial size (N_B). This reduces the resolution of estimation. To solve this problem, we will use other information to enhance the estimation accuracy.

B. FDS-Aided Max Path (FDS-MAX) Algorithm

Define $G(k)$ as,

$$G(k) = \begin{cases} \check{H}(k), & k = k_0 + i + jN_p, \\ 0, & k \in \text{others} \end{cases} \tag{14}$$

$G(k)$ can be written as $G(k) = \check{H}(k)U(k)$, with $U(k) = \sum_{l=0}^{M_R-1} V(k - lN_p)$ and $V(k)$ can be expressed as:

$$V(k) = \begin{cases} 1, & k_0 \leq k \leq k_0 + N_B - 1, \\ 0, & k \in \text{others} \end{cases} \tag{15}$$

Converting $G(k)$ into time domain through IFFT, we get $g(n) = \mathrm{IFFT}(G(k)) = \check{h}(n) \bigotimes u(n) = \sum_m \check{h}(m)u(\langle n-m\rangle_{N_{\mathrm{FFT}}})$, where $\bigotimes$ denotes cyclic convolution, $\langle z\rangle_{N_{\mathrm{FFT}}}$ denotes the remainder of z modulo N_{FFT}, $\check{h}(n) = \mathrm{IFFT}(\check{H}(k))$ is time shifted version (shift value is n_ϵ) of estimated CIR and $u(n) = \mathrm{IFFT}(U(k))$. $u(n)$ can be written as,

$$u(n) = \frac{e^{\frac{j2\pi nk_0}{N_{\mathrm{FFT}}}}(1 - e^{\frac{j2\pi nN_B}{N_{\mathrm{FFT}}}})(1 - e^{\frac{j2\pi nM_RN_p}{N_{\mathrm{FFT}}}})}{N_{\mathrm{FFT}}(1 - e^{\frac{j2\pi n}{N_{\mathrm{FFT}}}})(1 - e^{\frac{j2\pi nN_p}{N_{\mathrm{FFT}}}})} \tag{16}$$

$u(n)$ contains multiple peaks at positions $\alpha_j = \lfloor jN_{\mathrm{FFT}}/N_p + 0.5\rfloor$ ($j = 0, 1, \ldots, N_p - 1$). Between the peaks, the value of $u(n)$ is usually non-zero. Since $u(n)$ is not a Dirac delta function, after convolution of $\check{h}(n)$ with $u(n)$, the resultant $g(n)$ contains many "pseudo" paths, which is different from the true time domain channel $h(n)$. In multipath channel, "pseudo" paths may cause aliasing effect. As a result, the position corresponding to the timing offset will no longer have the maximal amplitude in the estimated CIR, and the position with maximal amplitude in estimated CIR is not the same as the true timing offset. Thus the accuracy of CIR-based algorithm [7] is affected.

With estimated $g(n)(n = 0, 1, \ldots, N_{\mathrm{FFT}} - 1)$, we define the estimated channel power delay profile (PDP) vector as $S = [s(0)\ s(1) \cdots s(N_{\mathrm{FFT}} - 1)]$, where, the n-th element of S is $s(n) = |g(n)|^2$. We divide the estimated channel PDP S into N_p segment. Segment j ($j = 0, 1, \ldots, N_p - 1$) is defined as $S^{(j)} = [s(\alpha_j)\ s(\alpha_j + 1)\ \cdots s(\alpha_j + N_j - 1)]$, $S^{(j)}$ consists of $N_j = \alpha_{j+1} - \alpha_j$ elements. Denote $S_{\max}(j)$ as the maximal value of $S^{(j)}$, and the position corresponding to maximal value $S_{\max}(j)$ as $\eta(j)$,

$$\eta(j) = \arg \max_{\alpha_j \leq k \leq \alpha_j + N_j - 1} s(k) \tag{17}$$

Assume that we have calculated $\hat{\zeta}$ according to (13). With the identified positions of N_p peaks $\eta(j)(j = 0, 1, \ldots, N_p - 1)$ in each segment, we can select the correct peak corresponding to the true relative arrival time with the aid of $\hat{\zeta}$. We determine the correct peak index J based on the criterion that the distance between $\hat{\zeta}$ and $\eta(J)$ is minimal:

$$J = \arg\min_j |\hat{\zeta} - \eta(j)| \tag{18}$$

$\eta(J)$ is the estimation of relative arrival time for FDS-MAX algorithm.

C. FDS-Aided Average Path (FDS-AVG) Algorithm

To further improve estimation accuracy, it is beneficial to refine the relative arrival time estimation by path averaging with de-noising. Denote the maximal value in segment

TABLE I
FLOP COUNT OF MAJOR FUNCTION BLOCKS

	FFT	IFFT	SVD	Minimal search
FLOP	112640	112640	19733	58995

J as $S_{\max}(J)$. By defining a threshold $\varrho_z = \rho S_{\max}\ (J)$ $(0 < \rho < 1)$, then de-noising can be expressed as,

$$Z_k = \begin{cases} s(k), \text{ if } s(k) \geq \varrho_z, \\ 0, \quad \text{otherwise} \end{cases} \tag{19}$$

The relative arrival time estimation for FDS-AVG algorithm can be written as,

$$\eta_{\text{avg}} = \frac{X_a}{X_b} \tag{20}$$

where, $X_a = \sum\limits_{k=\eta(J)-L_a}^{\eta(J)+L_a} kZ_k$, $X_b = \sum\limits_{k=\eta(J)-L_a}^{\eta(J)+L_a} Z_k$, $2L_a + 1$ is the window size for averaging.

D. Complexity of FDS Based Algorithm

We use the floating point operation (FLOP) count as the complexity metric [11]. FLOPs for FFT/IFFT, SVD and minimal search according to (13) are $5N_{\text{FFT}} log N_{\text{FFT}}$, $12N_B(M_R)^2 + 16(M_R)^3/3$ [11] and $M(M_R - 1)(8N_B - 1)$ respectively. The maximal search range in (13) is $[-N_{\text{FFT}}/2, N_{\text{FFT}}/2]$, however, in the condition that coarse synchronization has already been obtained, M can be reduced. For subcarrier spacing of 15 KHz and N_{FFT} = 2048, we can choose $M = 68$ which corresponds to search range from -1.1 μs to 1.1 μs. When $N_{\text{FFT}} = 2048$, $N_B = 12$, $M_R = 10$ and $M = 68$, FLOP of major function blocks is shown in Table I. From previous description, we can see that FDS-MAX algorithm needs 1 FFT, 1 IFFT, 1 SVD and 1 minimal search (according to (13)). The total required FLOP adds up to 304008 for FDS-MAX algorithm. CIR-based algorithm need 1 FFT and 1 IFFT, the total FLOP adds up to 225280. Compared with CIR-based algorithm, the complexity of FDS-MAX algorithm is increased by 34.95%.

IV. SIMULATION EVALUATION

We simulate uplink timing synchronization using B-IFDMA-structured RS in MATLAB platform. In our simulation, subcarrier spacing Δf is 30 KHz, $N_{\text{FFT}} = 2048$, $N_g = 144$, sampling interval $T_s = 16.275$ ns. Total 1200 ($N = 1200$) subcarriers are used for transmission and are divided into 10 interlaces, with each interlace consisting of 10 evenly-spaced RBs ($M_R = 10$). Each RB consists of 12 adjacent subcarriers ($N_B = 12$), and the space between adjacent RB in one interlace is 120 ($N_p = 120$). 10 UEs are considered, which transmit uplink RS signals simultaneously with the RS signal of each UE occupying one interlace and one OFDM symbol. The transmitted signal from each UE goes through independent propagation channel, and arrives at BS with random delay which is uniformly distributed in $[0, 16T_s]$. We adopt $\rho = 0.2$ and $L_a = 5$ in FDS-AVG algorithm. The search range for both FDS algorithm and CIR-based algorithm [7] is $M = 68$. Note that in the simulation we have quantized $\{\tau_l\}$ to be integer multiples of T_s.

A. Performance in Static Multipath Channel

Fig. 2 shows the Root Mean Square Error (RMSE) of uplink timing offset estimation error in two path ($L = 2$) static (non-fading) channel, with $h_0 = 0.8$, $h_1 = 0.6$, $\tau_0 = 0$,$\tau_1 = 16.275\ ns$. "CIR-based" denotes the CIR-based algorithm used in [7], "time domain algo" denotes the time domain algorithm used in [3], "ML" denotes the ML algorithm according to (6) (In the simulation, to save simulation time, we assume $\{h_l\}$ and $\{\tau_l\}$ are known at receiver, thus only need to search through n_ϵ). "CRLB" is Cramer Rao lower bound (CRLB) at AWGN channel for OFDM signal, which was derived in [4] and adapted to partial known RS subcarriers as follows:

$$var(\hat{\tau}) \geq \text{CRLB}(\hat{\tau}) = \frac{\sigma^2}{8\pi^2(\Delta f)^2 \sum\limits_{k \in \text{Q}} k^2 |\ d_k\ |^2} \tag{21}$$

where, σ^2 is the noise variance, and Q is the set of subcarrier indices where RS signals are located, however, the range of Q is from $-N/2$ to $N/2$. From Fig. 2 we can see that in multipath channel the accuracy of CIR-based algorithm is low. This is due to the aliasing effects caused by "pseudo" paths. As a result, the position corresponding to the timing offset will no longer has the maximal amplitude in the estimated CIR, and the position with maximal amplitude in estimated CIR is not same as the true timing offset. Thus the accuracy of CIR-based algorithm is degraded. From Fig. 2 we can see that FDS and FDS-MAX achieve much better performance than that of CIR-based algorithm and time domain algorithm. At RMSE = 10 ns, the performance gap between FDS-MAX and ML is 5.2 dB in the 2 path static channel.

B. Performance in Multipath Fading Channel

Fig. 3 shows the RMSE performance in Tapped Delay Line TDL-C [12] multipath fading channel. The moving speed is 3 Km/hr and the Root Mean Square (RMS) delay spread is 53 ns. The number of channel paths L is 14 in TDL-C channels. We can see that FDS based algorithm can counteract aliasing effect effectively, achieving much better performance than that of CIR-based algorithm and time domain algorithm. At SNR = 16 dB, the achieved accuracy of FDS-MAX is 17.5 ns. Under the same condition, the achieved accuracy is 173.7 ns and 159.4 ns for CIR-based algorithm and time domain algorithm, respectively. Compared with FDS-MAX, FDS-AVG can further improve the performance.

Note that in Fig. 2 RMSE of CIR-based algorithm increases slightly with SNR. This is because when SNR is high, due to aliasing effect the position with maximal amplitude in estimated CIR is not equal to the true timing offset. When SNR is low, the effect of noise occasionally causes the position with maximal amplitude in estimated CIR be equal to the true timing offset. In Fig. 3 such phenomenon is not observed. The reason is that in Fig. 2 the path amplitude does not change with time, while in Fig. 3 the path amplitude follows Rayleigh distribution and changes with time.

C. End-to-End Clock Synchronization Performance Evaluation

We assume that PTP is used as clock synchronization protocol. Based on timestamps exchanged between BS and UE, UE

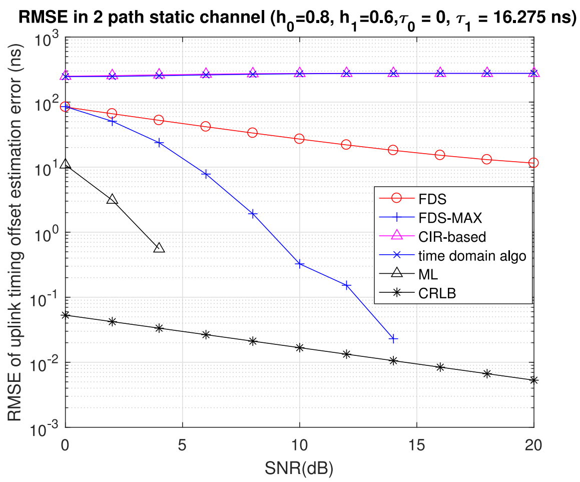

Fig. 2. RMSE performance in two path static channel ($h_0 = 0.8$, $h_1 = 0.6$,$\tau_0 = 0$,$\tau_1 = 16.275$ ns).

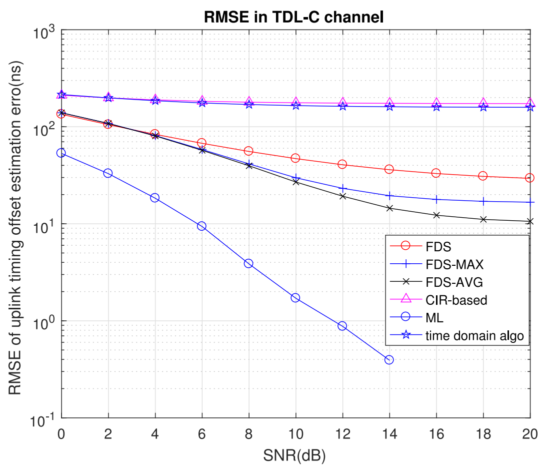

Fig. 3. RMSE performance in TDL-C channel.

TABLE II
CLOCK OFFSET ESTIMATION ERROR RANGE IN TDL-C CHANNEL (SNR=16 dB)

	error range at 99% probability	**error range at 99.999% probability**
FDS	[-273.2 ns, 256.2 ns]	[-344.4 ns, 314.6 ns]
FDS-MAX	[-263.2 ns, 261.7 ns]	[-395.1 ns, 324.3 ns]
FDS-AVG	[-261.9 ns, 261.5 ns]	[-391.0 ns, 314.9 ns]
CIR-based	[-363.5 ns, 366.7 ns]	[-445.5 ns, 514.2 ns]
time domain	[-343.2 ns, 368.9 ns]	[-443.0 ns, 524.9 ns]

can estimate the clock offset between UE and BS. The end-to-end clock offset estimation error $\epsilon_c = d_a - \epsilon_u/2 + \epsilon_d/2$, where, d_a is uplink/downlink delay asymmetry, ϵ_u and ϵ_d are uplink and downlink timing offset estimation error, respectively. Assume that the delay asymmetry is uniformly distributed in $[-160$ ns, 160 $ns]$ [13], and downlink timing offset estimation error is uniformly distributed in $[-260$ ns, 260 $ns]$ [13]. Table II shows clock offset estimation error range adopting different uplink timing offset estimation algorithm in TDL-C channel (SNR = 16 dB). We can see that clock offset estimation error of the proposed FDS, FDS-MAX FDS-AVG are within the required range $[-500$ ns, 500 $ns]$ with probability 99.999%, while clock offset estimation error of both CIR-based algorithm and time domain algorithm cannot meet $[-500$ ns, 500 $ns]$ requirement.

V. CONCLUSION

In this letter, we have proposed two efficient timing offset estimation methods for B-IFDMA-structured RS, namely, the FDS based algorithm, and FDS-aided algorithm (FDS-MAX and FDS-AVG). Our proposed algorithms can counteract the aliasing effects effectively and can achieve high accuracy in various propagation channels. At SNR = 16 dB, the RMSE of FDS-MAX algorithm is 8 times better than that of conventional CIR-based algorithm and time domain algorithm in TDL-C channels. At RMSE = 10 ns, the gap between our proposed FDS-MAX algorithm and ML algorithm is 5.2 dB in our considered two path static channel.

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