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5G Positioning Using Code-Phase Timing Recovery

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Abstract—To facilitate 5G-based positioning applications, Release 16 of the 3GPP 5G standard has defined the Positioning Reference Signal (PRS), which can be used to measure Time of Arrival (TOA) for downlink positioning. However, Orthogonal Frequency Division Multiplexing (OFDM) signals are sensitive and vulnerable to synchronization errors. Moreover, the highly configurable 5G PRS in Release 16 calls for a unique allocation pattern on the subcarriers. Existing timing recovery methods that have been employed for reference signals, which are evenly inserted in the subcarrier symbols, may not perform well. To solve the timing recovery issue of the OFDM signal through 5G standard-compliant PRS, we propose a three-stage timing recovery scheme. We use the 5G PRS as pilot symbols to estimate the path time delay and complete receiver sampling clock synchronization. We propose a generalized path time delay estimation method that can correct timing errors larger than one sample. In addition, we incorporate a delay-locked loop (DLL) that can track the PRS code-phase when the phase errors are within one sample, which showcases the precise positioning possible with a standard-compliant 5G New Radio (NR) signal.

Index Terms—5G Positioning, OFDM, TOA, Timing Recovery, DLL

I. INTRODUCTION

The ITU-R has defined three main application areas for the enhanced capabilities of 5G, namely, enhanced mobile broadband (eMBB), ultra-reliable low latency communications (URLLC) and massive machine type communications (mMTC). According to IMT-2020, 5G standard boasts of being ten to hundred times better in terms of peak data rate, latency and connection density [1]. Due to the ubiquity of cellular communications, 5G also provides an added value in terms of localization services in some Global Navigation Satellite System (GNSS)-denied environments, e.g. indoors, urban canyon and tunnels. The fundamental positioning techniques of cellular networks include trilateration, triangulation, proximity, scene analysis and hybrid positioning. Among which, time of arrival (TOA) based trilateration is one of the most common methods utilized from 2G right up to 5G [2]. In estimating TOA, many popular algorithms have been suggested, namely, multiple signal classification (MUSIC) [3], estimation of signal parameters via rotational invariance techniques (ESPRIT) [4] and maximum likelihood (ML) estimation [5]. MUSIC and ESPRIT however, fail to offer reliable performance when the number of samples is small [6].

The Positioning Reference Signal (PRS) was first defined in LTE as a pilot signal to perform TOA measurements. By simulation, the technical report in [7] demonstrates that the PRS-based positioning errors for LTE Release 9 are a few tens

of meters, which can be improved with larger bandwidth. To improve TOA-based positioning with pilot signals, a proper timing recovery mechanism is needed on the receiver side. This is done in [8], which proposes a delay-locked loop (DLL) based timing recovery scheme that includes symbol synchronization and sampling clock synchronization. However, the proposed method requires that pilot symbols are evenly allocated in the subcarrier symbols. Another work [9] proposes a frequency-domain frame synchronization algorithm but the estimate obtained is biased. The authors in [10] extend the previous approach to a LTE PRS-based fine time synchronization algorithm. Simulation results show a few tens of meters positioning error. Real LTE signal experimentation with cell-specific reference signals (CRS) has been carried out in [11] with a positioning error around 5.09m. More recently, the authors in [12] propose a phase-locked loop (PLL)-aided DLL with CRS as pilot signals for timing recovery, with a resultant root mean square error (RMSE) of 3.17m. Meanwhile, the authors in [13] propose a CRS-based TOA/Direction of Arrival (DOA) integrated solution, which achieves sub-meter level positioning errors. However, the CRS has been removed from 5G standards, and the redefined 5G PRS has a different pattern compared to LTE PRS.

Taking on the legacy of LTE, 5G utilizes Orthogonal Frequency Division Multiplexing (OFDM) waveform with flexible numerology. By modeling the clock offset and channel, the simulation results of the DOA/TOA integrated solutions [14], [15] show positioning errors below 1m. Based on the 5G Synchronization Signals and Physical Broadcast Channel (PBCH) Block (SSB), [16] uses a TOA/Angle of Departure (AOD) Extended Kalman Filter (EKF) method and obtains positioning errors below 1m through simulations. A more comprehensive performance analysis for TOA-based positioning in 5G Ultra-Dense Networks can be found in [17]. The first commercial 5G network was launched in 2019 based on the Release 15 specifications, and it was not until Release 16 that 5G PRS was defined in 3GPP TS 38.211 and TS 38.214. In March 2020, corresponding positioning procedures and 5G New Radio (NR) downlink (DL) PRS configurations were finalized in TS 37.355. These specifications pave the way for further verifying the performance of 5G PRS based localization services.

In this paper, we investigate the use of the 5G PRS in TOA measurements. To achieve accurate TOA measurements, timing recovery is critical. We propose a generalized path time delay estimation method to adapt to the highly configurable

5G PRS as pilot symbols. In addition, the corresponding DLL is presented to track the phase errors that are within one sample. This paper is organized as follows. The 5G NR Cyclic Prefix OFDM (CP-OFDM) modulation/demodulation, channel propagation model and 5G standard-compliant signals adopted as pilot signals are introduced in Section II. A three-stage timing recovery scheme, for measuring TOA, is given in Section III. Simulation results are presented in Section IV. Finally, the conclusion is drawn in Section V.

II. CP-OFDM SYSTEM AND SIGNALS DESCRIPTION

A. CP-OFDM Modulation

In OFDM modulation, let N_{SC} denote the number of valid subcarriers, N denote the number of points of Fast Fourier Transform (FFT)/inverse FFT (iFFT) where $N_{SC} < N$. Let $\{X_{i,n} | n = 0, \dots, N-1\}$ denote N subcarrier symbols, where i denotes the OFDM symbol number in time domain and n represents the FFT/iFFT point in frequency domain. Thus, $X_{i,n} = 0$ when $n \geq N_{SC}$.

At the transmitter, the N_{SC} symbols are modulated into N iFFT point locations, the rest of the points being zero-padded. To avoid intersymbol interference (ISI) introduced by multipath, 5G NR adopts CP-OFDM modulation, which further adds guard interval T_g between adjacent symbol intervals. The continuous-time baseband signal output of the transmitter can be represented as

$$f(t) = \frac{1}{N} \sum_{i=0}^{\infty} \sum_{n=0}^{N-1} X_{i,n} e^{j2\pi \frac{n(t-T_g-iT_s)}{T_u}} \quad (1)$$

where $iT_s \leq t < (i+1)T_s$, T_u denotes the useful symbol interval, T_s denotes the symbol interval including guard interval, and $j = \sqrt{-1}$.

Over one useful symbol interval, there are N samples implemented by sampling $f(t)$ at time $t = n \times T$, where $n = 0, \dots, N-1$, and T denotes the sample interval. The guard interval T_g is always a multiple of T , therefore, over one guard interval, there exists N_{CP} samples by duplicating N_{CP} samples at the tail of N useful samples. Thus, over one symbol interval, there are $N_{CP} + N$ samples in total. Finally, the modulation and upconversion to the carrier frequency f_o of the complex-valued OFDM baseband signal is given by

$$s(t) = f(nT) e^{j(2\pi f_o t + \theta_o)} \quad (2)$$

$$f(nT) = \frac{1}{N} \sum_{i=0}^{\infty} \sum_{n=0}^{N-1} X_{i,n} e^{j2\pi \frac{k_o n}{N}} \quad (3)$$

where $k_o = -N_{CP}, \dots, N-1$, f_o denotes the carrier wave frequency, and θ_o is the initial phase at the transmitter.

B. CP-OFDM Demodulation

In this paper, we do not consider the carrier frequency offset issue, since the carrier frequency synchronization method in [18], for instance, can be easily combined with the timing recovery scheme. We also assume a multipath fading channel

including line-of-sight (LOS). The channel frequency response can be described as [8]

$$H_{i,n} = \sum_{l=0}^{L-1} h_l(iT_s) e^{-j2\pi \tau_l(n/T_u)} \quad (4)$$

where $h_l(t)$ denotes the path gain, τ_l denotes path time delay, L is the number of paths, and $\tau_0 = 0$.

If the receiver can successfully remove CP and set the FFT starting point within $[-CP, 0]$, the FFT window will cover samples only from the same symbol interval. The FFT window will include CP samples, if the starting point is in front of the first useful sample. Then, the received N samples over the i th symbol period for FFT are given by

$$r(nT) = \sum_{l=0}^{L-1} h_l(nT) s(nT - \tau_l) + n_e \quad (5)$$

where $n = 0, \dots, N-1$, and n_e is channel noise. Therefore, the n th symbol in the i th symbol period after OFDM demodulation (FFT) can be represented as

$$\begin{aligned} \hat{X}_{i,n} &= \sum_{i=0}^{\infty} \frac{1}{N} \sum_{k=0}^{N-1} \sum_{m=0}^{N-1} e^{j2\pi \frac{nm}{N}} X_{i,m} H_{i,n} + n_{i,n} \\ &= \sum_{i=0}^{\infty} e^{j2\pi \frac{nk'}{N}} X_{i,n} H_{i,n} + n_{i,n} \end{aligned} \quad (6)$$

where $k' \in [-N_{CP}, 0]$ is a real number, it denotes the offset of the first received useful sample with respect to the first transmitted useful sample during the i th symbol interval, and $n_{i,n}$ denotes channel noise. The Eq. (6) indicates that as long as the FFT window falls within the CP interval, the orthogonality between the different subcarriers is maintained. However, if offset $k' \neq 0$, a symbol timing error within this interval can result in a phase rotation $e^{j2\pi \frac{nk'}{N}}$ of every subcarrier symbol $X_{i,n}$. Eq. (6) assumes that there are no sampling clock errors, which implies, the actual receiver sampling time interval T' is equal to T , which is not always the case. To check the impact of the imperfect T' , define the sampling frequency offset $\beta = (T' - T)/T$. Without loss of generality, we also assume $k' = 0$, and the sampling time error of first useful sample in the $i = 0$ symbol interval is $\Delta T = T - T'$, giving us

$$\hat{X}_{i,k} = \frac{1}{N} \sum_{i=0}^{\infty} \sum_{n=0}^{N-1} X_{i,n} e^{\frac{j2\pi n \beta i T_s}{T_u}} f(\beta, n, k, N) \quad (7)$$

$$f(\beta, n, k, N) = \sum_{m=0}^{N-1} e^{j2\pi m \frac{n+n\beta-k}{N}}. \quad (8)$$

When $n\beta \ll 1$, if $n \neq k$, $f(\beta, n, k, N) \approx 0$, otherwise, $f(\beta, n, k, N) \approx 1$. Thus, we can rewrite (7) as

$$\hat{X}_{i,n} \approx \sum_{i=0}^{\infty} e^{\frac{j2\pi n \beta i T_s}{T_u}} X_{i,n} \quad (9)$$

where $n = 0, \dots, N-1$. The Eq. (9) indicates that the sampling frequency offset also results in phase rotation of symbols.

where k is a function of channel power delay profile $\hat{S}_k$ defined in [8]. Eq. (13) is for any symbol pattern described in Eq. (10). When $b = c = 0$, the pattern of pilot symbols and the corresponding estimate $\hat{e}_\theta$ is in line with [8]. The estimated timing error $\hat{e}_\theta$ is a negative real number. The integer part $\text{Int}\{\hat{e}_\theta\}$ is fed back to node A through node 2 to adjust the FFT starting point, and the readjusted 5G NR DL PRS symbol

Fig. 2: Normalized S-curve functions of different patterns

sequence is obtained subsequently. Finally, the fractional part $\text{Fra}\{\hat{e}_\theta\}$ is fed into the ‘Phase rotation’ module through node 1.

C. DLL Tracking

In the initial phase of stage three, the adjusted timing error of the PRS pilot symbol introduced by phase rotation is supposed to be within one sample interval. The normalized S-curve function over the i th symbol period is defined as

$$S_i(\hat{e}_\theta - e_\theta, \xi) = \frac{|R_l(i, \xi)|^2 - |R_e(i, \xi)|^2}{M^2 A^2} \quad (14)$$

where $A = |\Upsilon_m|^2$, M denotes the number of pilot symbols over the i th OFDM symbol interval, ξ ($0 < \xi \leq 0.5$) is the advanced (and retarded) interval which is normalized to the OFDM sample interval, $R_l(i, \xi)$ and $R_e(i, \xi)$ are the local generated late and early reference pilots described in [8]. As shown in Fig. 2, *Pattern A* and *Pattern B* denote two normalized S-curve functions with different pilot symbol patterns defined by Eq. (10). For *Pattern A*, $M = 432$, $a = 2$, $b = 160$, $c = 0$, $\xi = 0.5$, $A = 1$, and for *Pattern B*, $M = 72$, $a = 12$, $b = 160$, $c = 0$, $\xi = 0.5$, $A = 1$. The S-curve function indicates that the timing error could be represented by the range of near linearity. Therefore, the corresponding DLL discriminator output function can be defined as

$$D_i \triangleq -\frac{1}{k_s} \times S_i(\hat{e}_\theta - e_\theta, \xi) \quad (15)$$

where $k_d = -1/k_s$ and k_s is the slope of normalized S-curve function [12]. The system transfer functions of lowpass filter (LPF) and accumulator are shown in Fig. 3 where τ_1 and τ_2 are the filter parameters, and can be given by

$$\tau_1 = 1/\omega_n^2 \quad (16)$$

$$\tau_2 = 2\xi_f/\omega_n \quad (17)$$

where ξ_f is the damping factor, ω_n is the undamped natural frequency. The difference equations of LPF and Accumulator are given by

$$y_f(n) - y_f(n-1) = (2\xi_f\omega_n + \omega_n^2 T_f)x_f(n) - 2\xi_f\omega_n x_f(n-1) \quad (18)$$

$$\hat{e}_\theta(n+1) = \hat{e}_\theta(n) + T_f y_f(n) \quad (19)$$

where T_f is the filter interval, $y_f(n)$ is LPF output and $x_f(n)$ is LPF input (Discriminator output) in the n th filtering cycle. $\hat{e}_\theta$ represents the estimated timing error, which is normalized by sampling interval T .

Finally, the timing adjustment based on the estimated $\hat{e}_\theta$ is given by

$$\hat{X}_{i,n(m)} = \sum_{i=0}^{\infty} e^{-j2\pi \frac{n(m)}{N} (\hat{e}_\theta)} \tilde{X}_{i,n(m)} \quad (20)$$

where $\tilde{X}_{i,n(m)}$ is n th subcarrier symbol after OFDM demodulation over the i th OFDM symbol interval, while $\hat{X}_{i,n(m)}$ denotes the updated n th subcarrier symbol.

Fig. 3: The structure of DLL

IV. SIMULATION RESULTS

Experiments are conducted under multipath Rician fading channel (Fig. 4) and AWGN (30 dB) channel in Matlab, where ‘PathDelays’ denotes the discrete delays of five-path channels in seconds. For outdoor environments, the path delays after the first path are typically between $1e-7$ seconds and $1e-5$ seconds. ‘AveragePathGains’ denotes the corresponding path gains in dB, ‘KFactor’ is the linear ratio of specular power to diffuse power, for Rician fading, the K-factor is typically within the range [1, 10]. For 5G CP-OFDM signal, $N = 1024$, $N_{SC} = 864$, subcarrier spacing (SCS) is 30 kHz. For DLL, $\xi = 0.5$, $k_s = -3.24$, and, loop filter parameters, $\xi_f = 0.707$. The experiments use signals with 10 subframes (20 slots, 280 symbols), with one SSB placed in the first slot, and the PRS allocated in all slots except those occupied by the SSB.

```
ricChan = comm.RicianChannel( ...
    'SampleRate',      Settings.SCS*1000*Settings.nfft, ...
    'PathDelays',      (0:0.4:1.6)*1e-6, ...
    'AveragePathGains', [0 -3 -6 -9 -12], ...
    'KFactor',         6, ...
    'DirectPathDopplerShift', 400, ...
    'MaximumDopplerShift', 600, ...
    'RandomStream',    'mt19937ar with seed', ...
    'Seed',            100, ...
    'PathGainsOutputPort', true);
```

Fig. 4: Rician channel settings

A. Generalized Path Time Delay Estimate

The DL PRS has a diagonal pattern, among which, the following PRS columns (in time domain) are selected as pilot symbols: $a = 2, b = 160, c = 0, M = 432$ (see Fig. 5(a)). The overview of the frame structure including DL PRS pattern A is shown in Fig. 6, where PSS/SSS/PBCH/PBCH DM-RS are allocated only in 5-8 OFDM symbols, while the DL PRS is scattered in the selected OFDM symbols except for the SSB.

To evaluate the timing estimation performance of stage 1 and stage 2, where, only integer timing errors can be corrected, we look at Table I and Table II. Here, the coarse estimation (**Stage 1**) and generalized path time delay estimation (**Stage 2**) with different v are illustrated, with **GT** denoting the ground truth, and the unit being one sample. The results show that the proposed generalized path time delay estimation method in stage 2 can successfully correct timing errors larger than one sample, with fractional timing errors being handled by stage 3.

TABLE I: Timing estimation with AWGN channel

Stage 1	$+v$	Stage 2	GT
1	-15	-14	-14
0	-30	-30	-30
0	-45	-45	-45

TABLE II: Timing estimation with Rician plus AWGN channel

Stage 1	$+v$	Stage 2	GT
0	-15	-15	-15
11	-45	-34	-34
1	-60	-59	-59

Fig. 5: NR DL PRS patterns

Fig. 6: Frame structure

B. Timing Recovery with PRS Pattern A

With the same CP-OFDM signal frame structure depicted in Fig. 6 and the same pilot symbol pattern shown in Fig. 5(a), we further inject some fractional timing errors into OFDM symbols before CP-OFDM modulation at the transmitter. The

fractional part $\text{Fra}\{\theta_o\}$ of simulated timing error over the i th symbol period is given by

$$\text{Fra}\{\theta_o\} = -0.2 + 0.06 \sin(500\pi i T_s) \quad (21)$$

where T_s denotes the symbol interval. For $N = 1024$, and $SCS = 30$ kHz, 0.01 sample change represents 0.098m distance change between transmitter and receiver. Therefore, Eq. (21) represents maximum 1.176m ($0.06 * 2 * 9.8$) distance change over 2ms ($1/500$), which is large enough to represent typical land transportation.

As shown in Fig. 7, after correlation in stage one, the integer part of the timing error is 12 samples (ground truth is 4401 samples), and since $v = -50$, the adjusted integer error is -38. In stage two, the estimated integer part of timing error is -38, thus only the fractional part of timing error is left and fed into stage three.

In Fig. 8, we see the output of the three-stage timing error estimation, where ‘GT’ denotes ground truth of the fractional part of the timing error, and ‘Est’ denotes the DLL estimated timing error. Because the pilot symbol is not placed in every OFDM symbol column and the range of linearity of S-curve function is nearly linear, the result is a fluctuation in the estimation error with a mean of 0.052 samples, and an error standard deviation of 0.075 samples.

The quality of the transmission channel also contributes to the estimation error. For comparison, the timing estimation errors for the AWGN-only channel are shown in Fig. 9, with a mean estimation error of 0.043 samples, and an error standard deviation of 0.048 samples.

Fig. 7: PSS correlation (Rician multipath plus AWGN)

C. Timing Recovery with PRS Pattern B

In Section IV-B, a densely populated PRS is adopted, in order to further check the tracking ability of the DLL, the following PRS columns (in time domain) are selected as pilot symbols: $a = 12, b = 160, c = 0, M = 72$ (see Fig. 5(b)), the entire frame structure is illustrated in Fig. 10. The fractional part of simulated timing error over the i th symbol period is also given by Eq. (21).

As shown in Fig. 11, over 10 subframes (280 symbols), there are 16 PRS columns selected for the DLL code-phase

Fig. 8: DLL tracking results

Fig. 9: DLL tracking results (AWGN only)

tracking cycle. For the AWGN-only channel, the mean of estimation error is 0.039 samples, and the error standard deviation is 0.087 samples. For the Rician plus AWGN channel, the mean of the estimation error is 0.130 samples, and the error standard deviation is 0.148 samples.

The experiments demonstrate that regardless of the PRS pattern adopted within the specified standard, the timing recovery scheme detailed in this paper can track the timing error successfully, resulting in a good candidate for 5G localization.

Fig. 10: Frame structure (Pattern B)

Fig. 11: DLL tracking results (Pattern B)

V. CONCLUSION

In this paper, we propose a code-phase based CP-OFDM signal timing recovery scheme, which is suitable for 5G standard-compliant signals. The basis for a code phase-based implementation is because a DLL is used to track the leftover phase rotation of the PRS pilot symbols instead of tracking the carrier phase.

We use a generalized path time delay estimation method to adapt to the generalized pilot symbol pattern, and also include the design and implementation of a DLL in this paper. We evaluate the performance of the timing recovery scheme for various experimental parameters in MATLAB simulation, the results illustrate that the generalized path time delay estimation method can effectively recover the timing error of 5G NR DL PRS based pilot symbol, being able to reduce the timing error within one sample. In addition, the proposed DLL can track the CP-OFDM signal timing error within one sample, the mean of the tracking error being smaller than 0.130 samples, which represents less than 1.27m ranging error.

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