A novel fine Doppler frequency acquisition algorithm for BDS-3 B1C signal based on an adaptive filter

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Abstract
Fine Doppler frequency estimation has an important role in accelerating the convergence of the tracking loop in a global navigation satellite system (GNSS) receiver to achieve short time to first fix. BDS-3 started broadcasting a civil B1C signal to provide open services for global users, which is beneficial for GNSS-based applications. Therefore, a fine Doppler frequency acquisition algorithm based on an adaptive filter is proposed, whose purpose is to acquire the BDS-3 B1C signal Doppler frequency accurately after the completion of coarse acquisition. The proposed algorithm is based on a first-order complex-coefficients adaptive filter. The adaptive filter depends on the proposed adaptation algorithm to track the input BDS-3 B1C signal. An accurate Doppler frequency estimate is extracted. Simulation results show the proposed algorithm has high acquisition sensitivity, high acquisition accuracy, short acquisition time, and few hardware resources consumption, and also works well under many different coarse acquisition strategies. Overall, the proposed algorithm is better than the generic second-order frequency locked loop. Consequently, the proposed algorithm has high practical value.

1 INTRODUCTION

An acquisition engine is one of the most important digital baseband blocks in a global navigation satellite system (GNSS) receiver. It provides coarse estimates of the code phase and Doppler frequency of input signals [1]. The acquisition engine generally outputs these rough estimates to the tracking stage to assist the tracking loop to track the input signal [2]. However, rough Doppler frequency estimates cause the tracking loop to spend a long time on convergence, which cannot satisfy the requirements of a fast convergence rate and short time to first fix (TTF) [3]. To improve Doppler frequency estimation accuracy, many algorithms are proposed to address this problem [4–16]. These algorithms are generally divided into two categories: the one-step algorithm and the two-step algorithm.

The one-step algorithm includes three methods. The first increases the number of frequency bins in the sequential searching process of Doppler frequency [4], but it requires a long acquisition time. The second extends the length of Fast Fourier Transform (FFT) by padding zeros to improve frequency resolution [5], which unfortunately requires a large-size FFT. The last one depends on the non-linear relation involving three FFT samples calculated in the coarse acquisition to obtain an accurate Doppler frequency estimate [6, 7], but it is suitable only for the high signal-to-noise ratio (SNR) situation.

In the two-step algorithm, the coarse estimates of code phase and Doppler frequency are found by the first step. Based on these coarse estimates, the second step first demodulates primary code and then obtains the fine estimate around the coarse estimate of Doppler frequency. There are four typical methods. The first carries out a sequential searching process of Doppler frequency, again with small frequency steps [8], but it still requires a long acquisition time. The second conducts a parallel frequency search again by one more FFT [9], which leads to an increase in hardware resources consumption. The third is based on the frequency locked loop (FLL) [10]. Although the FLL consumes less hardware resources, it takes a long time to converge. The last is based on the Gram-Schmidt
orthogonal method [11, 12], which, however, needs a lot of iterations and has a heavy computational burden.

The Chinese third-generation BeiDou navigation satellite system (BDS-3) started broadcasting a civil B1C signal to provide open services for global users, which is beneficial for GNSS-based applications. The BDS-3 B1C signal has a long primary code that contains a total of 10,230 chips in 10 ms, and then acquisition of the BDS-3 B1C signal requires lots of correlation operations, which implies high computational complexity [17].

To acquire the BDS-3 B1C signal by limited resources, some acquisition algorithms such as partial correlation method are proposed. These algorithms also provide a coarse Doppler frequency estimate [18, 19]. Hence, a fine Doppler frequency acquisition algorithm is also necessary for the BDS-3 B1C signal.

To overcome those drawbacks of the fine Doppler frequency acquisition methods and fully use the BDS-3 B1C signal, an adaptive filter–based fine Doppler frequency acquisition algorithm for BDS-3 B1C signal is proposed. The method is based on a first-order complex-coefficients adaptive filter. The adaptive filter depends on the proposed adaptation algorithm to track the BDS-3 B1C signal after removal of the primary code. Finally, the fine Doppler frequency estimate is extracted. The proposed algorithm also belongs to the two-step algorithm category.

The rest of this work is organised as follows. First, the BDS-3 B1C signal is briefly introduced in Section 2. Then, coarse acquisition principles of the BDS-3 B1C signal are described in Section 3. Section 4 presents the proposed fine Doppler frequency acquisition algorithm in detail. Simulation results and performance evaluation are given in Section 5. Finally, some conclusions are drawn in Section 6 to summarise the work.

2 | BDS-3 B1C SIGNAL

The BDS-3 B1C signal is a modern GNSS signal that includes two components: data and pilot. The data component modulates data bits on the primary code, and the pilot component modulates secondary code on the primary code. The primary code contains 10,230 chips in 10 ms, and the secondary code contains 1800 chips in 18 s.

As Table 1 shows, the pilot component of the BDS-3 B1C signal adopts the modulation mode of QMBOC(6,1,4/33) that is composed of two parts: BOC(1,1) and BOC(6,1). On the one hand, the power of BOC(1,1) accounts for 29/44 of the total power of the BDS-3 B1C signal. On the other hand, the subcarrier frequency of the BOC(1,1) is 1.023 MHz and is less than that of the BOC(6,1), which indicates that the acquisition engine designed for the BOC(1,1) generally has lower complexity than that designed for the BOC(6,1). Therefore, only the BOC(1,1) of the pilot component is used to acquire the BDS-3 B1C signal.

3 | COARSE ACQUISITION PRINCIPLES

This work presents the proposed fine Doppler frequency acquisition algorithm. To illustrate the proposed algorithm, without a loss of generality, the short-time coherent integration is used plus an FFT method to carry out the coarse acquisition of a BDS-3 B1C signal [20–23], as shown in Figure 1. Many other methods are also suitable for the coarse acquisition of the BDS-3 B1C signal, such as sequential search [5], parallel code phase search [24], and parallel Doppler frequency search [9]. All of these methods may be combined with the proposed method. Moreover, this work adopts the Binary Phase Shift Keying-like method [25] to remove the side peaks of the BOC(1,1) autocorrelation function (ACF).

In general, at the output of the radio-frequency front end, the discrete-time BOC(1,1) signal of the BDS-3 B1C pilot component is given by:

\[
r[n] = \sqrt{2P} \cdot s[n - \tau] \cdot c[n - \tau] \cdot \text{sign}[\sin(2\pi f_c n T_s)] \cdot \sin[2\pi (f_{IF} + f_d) n T_s + \varphi_0] + \eta[n]
\]

where \(P\) represents the signal power, \(s[n]\) is the secondary code, \(c[n]\) is the primary code, \(\tau\) represents the code phase, \(f_c = 1.023\) MHz, \(f_{IF}\) represents the intermediate frequency, \(f_d\) is the Doppler frequency, \(T_s\) is the sampling interval, \(\varphi_0\) is the initial phase, and \(\eta[n]\) is additive white Gaussian noise with a two-side noise power spectrum density of \(N_0/2\) W/Hz.

The local generated complex exponential carrier \(\exp[-j2\pi(f_{IF} + f_d)nT_s]\) subsequently mixes with \(r[n]\), and then the result after the mixing passes through the low-pass filter (LPF). For the convenience of expression, the LPF is assumed to be an ideal finite impulse response (FIR) filter. Afterwards, the local generated primary code \(c[n - \tilde{\tau}]\) correlates with the result at the output of the FIR filter, and short-time coherent integration is performed. As a result, the \(l\)th short-time coherent integration result is given by:

\[
G(\tilde{\tau}, l) = \sum_{n=-\infty}^{\infty} LPF\{r[n] \cdot \exp[-j2\pi(f_{IF} + f_d)nT_s]\} \cdot c[n - \tilde{\tau}] \\
= \frac{\sqrt{2P} \cdot R(\Delta \tau)}{2} \cdot \frac{\sin(\pi f_d T_s)}{\sin(\pi f_d T_s)} \cdot s[l] \\
\cdot \exp\{j\pi f_d (2lN + N - 1) T_s + \varphi_0\} + \xi[l]
\]
FIGURE 1 Block diagram of coarse acquisition algorithm based on short-time coherent integration plus an FFT for BDS-3 BIC signal. BDS, BeiDou navigation satellite system

where $T_c$ is the short coherent integration time and $T_c’ = NT_c$. Also in Equation (2), $\Delta \tau = \tau - \hat{\tau}$. $R(\Delta \tau)$ denotes the primary code ACF and $2^m$ is the noise term. The total coherent integration time is $T_p$ and $T_p’ = L T_c’$.

Afterwards, an $N_{FFT}$-points zero-padding FFT is used to conduct parallel Doppler frequency search, and $N_{FFT} = 2^m$ ($m \in N^+$). To avoid the scalloping loss, $N_{FFT} \geq 2L$, and the number of zero is $N_{FFT} - L$. As a result, the output of the $i$th FFT operation is given by:

$$F_i(\hat{\tau}, \hat{f}_d) = \sum_{l=1}^{(i+1) L - 1} G(\hat{\tau}, l) \cdot \exp \left( -j \frac{2 \pi}{N_{FFT}} kl \right)$$

$$= \sqrt{2 \pi} \cdot R(\Delta \tau) \cdot \sin \left( \pi f_d T_c \right) \cdot \sin \left[ \pi \Delta f_d T_p \right]$$

$$\cdot \exp (j \varphi) + \xi[l]$$

$$= \Lambda_0 \cdot s[l] \cdot \exp (j \varphi) + \xi[l]$$

(3)

where $\Delta f_d$ represents the estimation deviation of Doppler frequency, $\varphi = \phi_0 + \pi f_d (N - 1) T_s + \pi \Delta f_d (2L + L - 1) T_c$. According to Equation (3), the coarse Doppler frequency estimate is $\hat{f}_d$, and $\hat{f}_d = k / N_{FFT} \cdot T_c$. Then, $\Delta f_d = \tilde{f}_d - \hat{f}_d$, and $\Delta f_d \in \left[ -\frac{1}{2N_{FFT}T_c}, \frac{1}{2N_{FFT}T_c} \right]$ when the SNR of the input BDS-3 BIC signal satisfies the acquisition sensitivity. Also in Equation (3), $\xi[l]$ is the noise term. According to the central limit theorem, $\xi[l]$ is a complex Gaussian random variable with zero mean and variance $2\sigma^2$. Furthermore, the real and imaginary part of the $\xi[l]$ are independent Gaussian variables with zero mean and variance $\sigma^2$.

At this moment, the SNR in $dB$ is given by:

$$SNR = 10 \cdot \log \left( \frac{\Lambda_0 \cdot 2}{2 \sigma^2} \right)$$

(4)

It is known that non-coherent integration is an effective way to extend total integration time to $N_{FFT}$. Based on Equation (3), the result after $I$-times non-coherent integration is given by:

$$S(\hat{\tau}, \hat{f}_d) = \sum_{i=0}^{I-1} \left| F_i(\hat{\tau}, \hat{f}_d) \right|^2$$

(5)

When $R(\Delta \tau) \rightarrow 0$ or the satellite is out of view, there is hypothesis $H_0$, and $S(\hat{\tau}, \hat{f}_d) = \sum_{l=0}^{I-1} \left| \xi[l] \right|^2$, which indicate the $S(\hat{\tau}, \hat{f}_d)$ follows the central chi-square ($\chi^2$) distribution with 2$I$ degrees of freedom. When $R(\Delta \tau) \rightarrow 1$, there is hypothesis $H_1$, and $S(\hat{\tau}, \hat{f}_d)$ follows non-central $\chi^2$ distribution with 2$I$ degrees of freedom and non-centrality parameter $\lambda$. The power of input signal is generally assumed to be constant during $N_{FFT}$, therefore, $\lambda = 2I \cdot SNR \cdot \sigma^2$.

In terms of false-alarm probability $P_{fa}$ and detection probability $P_d$, the critical SNR is given by:

$$SNR_{Th} = F_n^{-1} \left( 1 - P_{fa}, 2I \right), 1 - P_d, 2I \right)$$

(6)

where $F_n^{-1} (\cdot)$ is the inverse function of the cumulative distribution function of central $\chi^2$ distribution, and $F_n^{-1} (\cdot)$ is the inverse function of the cumulative distribution function of non-central $\chi^2$ distribution. According to Equations (3), (4), and (6), critical carrier-to-noise ratio ($C/N_0$) in dB-Hz is given by:

$$C/N_0_{Th} = SNR_{Th} - 10 \cdot \log(T_p)$$

(7)

$C/N_0_{Th}$ represents the required minimum value of the $C$/$N_0$ of the received signal to satisfy the receiver operation characteristics.

4 | PROPOSED ALGORITHM

The adaptive notch filter technique is widely used for single-frequency sinusoidal signal enhancement and extraction, and single-frequency interference mitigation [26–30], because the adaptive notch filter is able to estimate the carrier frequency of the input signal accurately, which implies the adaptive notch filter can be used for the fine Doppler frequency estimation. Therefore, based on the adaptive notch filter, a low-complexity adaptive filter is proposed to conduct fine Doppler frequency acquisition.

4.1 | Adaptive filter

The transfer function of the first-order complex-coefficients zero-pole constrained notch filter is generally given by:
\[ H(z) = \frac{1 - \exp(j2\pi f) \cdot z^{-1}}{1 - \gamma \cdot \exp(j2\pi f) \cdot z^{-1}} \]  

(8)

where \( f \) represents the central frequency of the notch filter, and \( \gamma (0 < \gamma < 1) \) is the zero-pole constraint factor and determines the 3-dB attenuation bandwidth of the notch filter. According to Borio et al. [31], Equation (8) is composed of two parts. The first part is called the autoregression (AR) block, corresponding to \( H_{AR}(z) = 1/[1 - \gamma \cdot \exp(j2\pi f) \cdot z^{-1}] \). The second part is called the moving-average (MA) block, corresponding to \( H_{MA}(z) = 1 - \exp(j2\pi f) \cdot z^{-1} \). The direct-form structure of the notch filter is shown as Figure 2, where \( u[n] \) and \( y[n] \) are the \( n \)th complex input and output samples, respectively.

As Figure 2 shows, the cost function is given by:

\[
J(f) = E[|y[n]|^2] = E[|x[n]|^2] + E[|x_{n-1}|^2] - E[x_n \cdot x_{n-1}] \exp(j2\pi f) - E[x_n \cdot x_{n-1}^*] \exp(-j2\pi f)
\]

(9)

where \(|y[n]|^2\) at the right up corner represents conjugate operation, and \( x_n \) represents the \( x[n] \). To find the parameter \( f \) that minimises the cost function and equals the carrier frequency of the input signal, the cost function is differentiated as:

\[
\frac{dJ(f)}{df} = (j2\pi) E[x_n \cdot x_{n-1}^*] \exp(-j2\pi f) - (j2\pi) E[x_{n-1}^* \cdot x_n] \exp(j2\pi f)
\]

(10)

We can see from Equation (10) that the second derivative of the cost function \( J(f) \) exists.

Furthermore, according to the Newton algorithm, \( d_k \) is defined as the descent direction at the \( k \)th iteration, and then the adaptation equation can be generally written as:

\[
f_{k+1} = f_k + ad_k
\]

(11)

where \( a \) is the step size. Then, the precise linear search method is applied to determine the value of the \( a \), so the function about \( a \) is given by:

\[
\phi(a) = J(f_{k+1}) - J(f_k + ad_k)
\]

(12)

The derivative of the function \( \phi(a) \) is deduced as:

\[
\frac{d\phi(a)}{da} = (j2\pi d_k) E[x_n \cdot x_{n-1}^*] \exp(-j2\pi f_{k+1}) - (j2\pi d_k) E[x_{n-1}^* \cdot x_n] \exp(j2\pi f_{k+1})
\]

(13)

Let \( \frac{d\phi(a)}{da} = 0 \), and then it can be obtained as:

\[
\exp(j2\pi f_{k+1}) = \frac{E[x_n x_{n-1}^*]}{E[x_{n-1}^* x_n]} = \exp[j2\pi F(x_n x_{n-1}^*)]
\]

(14)

Substituting Equation (14) into Equation (10), it can be obtained as:

\[
\frac{dJ(f_{k+1})}{df} = (j2\pi) \{ E[x_n \cdot x_{n-1}^*] - E[x_{n-1}^* \cdot x_n] \exp(j2\pi f_{k+1}) \} \cdot \exp(-j2\pi f_{k+1}) = 0
\]

(15)

Considering Equations (14) and (15), if Equation (14) is adopted as the adaptation algorithm, the central frequency of the first-order complex-coefficients adaptive notch filter will converge to the carrier frequency of the input signal by only one iteration in theory, which implies a fast convergence rate.

However, Equation (14) means heavy computation, and \( \exp(j2\pi f_{k+1}) \) is the expected value. Therefore, we further simplify Equation (14), and the adaptation algorithm is given by:

\[
\exp(j2\pi f_{k+1}) = \frac{x_n x_{n-1}^*}{x_{n-1}^* x_n} = \exp[j2\pi F(x_n x_{n-1}^*)]
\]

(16)

where \( F(x_n x_{n-1}^*) \) represents the function of extracting the carrier frequency of the input signal \( x_n \), and \( F(0) = 0 \). Function \( F(x_n x_{n-1}^*) \) can be implemented by a frequency discrimination function such as a four-quadrant arctangent function \( \arctan^2(\cdot) \).

Equation (16) is further smoothed by:

\[
\exp(j2\pi f_{k+1}) = \lambda_k \exp(j2\pi f_k) + (1 - \lambda_k) \exp(j2\pi F(x_n x_{n-1}^*))
\]

(17)

where \( \lambda_k (0 \leq \lambda_k < 1) \) is the forgetting factor, \( k \geq 0 \), and \( f_0 = 0 \).

To ensure a fast convergence rate and avoid intense oscillation, \( \lambda_k \) is adjusted by:
where \( 0 < \lambda_0 < 1 \), and \( 0 < \beta \leq \lambda_0 \).

Observing Equation (17), the adaptation algorithm is concerned only with \( x[n] \), the output of the AR block shown as Figure 2. Consequently, based on this adaptation algorithm, the AR block of the first-order complex-coefficients notch filter can be taken as an adaptive filter to track the carrier frequency of input signals, shown in Figure 3. The AR block has lower complexity than the notch filter.

### 4.2 Fine acquisition principles

According to the coarse code phase estimate \( \hat{\tau} \), local code NCO generates the product of primary code and subcarrier \( c[n - \hat{\tau}] \cdot \text{sign}[\sin(2\pi f_{NCO} n T_s)] \) to correlate with the data samples after mixing. Coherent integration is subsequently carried out. The coherent integration time is \( T_{\text{cob}} \), and \( T_{\text{cob}} = N_p T_s \). As a result, the i\textsuperscript{th} coherent integration result is deduced as:

\[
E_i(\hat{\tau}, \hat{f}_d) = \sum_{n=0}^{N_p-1} r[n] \cdot \exp(-j2\pi f_{NCO} n T_s) \cdot c[n - \hat{\tau}]
\]

\[
\cdot \text{sign}[\sin(2\pi f_{NCO} n T_s)]
\]

\[
= \frac{\sqrt{2p} \cdot R(\Delta f \cdot \sin(\pi f_d T_{\text{cob}}))}{2 \sin(\pi f_d T_{\text{cob}})} \cdot s[\hat{i}]
\]

\[
\cdot \exp\{j[\pi f_d (2iN_p + N_p - 1) T_s + \phi_0]\}
\]

\[
+ \xi_p[\hat{i}]
\]

\[
= A_p \cdot s[\hat{i}] \cdot \exp\{j[\pi f_d (2iN_p + N_p - 1) T_s + \phi_0]\} + \xi_p[\hat{i}]
\]

\[
(20)
\]

where \( \xi_p[\hat{i}] \) is the present noise term.

The secondary code \( s[\hat{i}] \) in Equation (20) is a pseudorandom noise sequence; therefore, the \( E_i(\hat{\tau}, \hat{f}_d) \) is a direct spread spectrum signal. Because the adaptive filter merely tracks the single-frequency signal, the fine Doppler frequency acquisition algorithm has to overcome the secondary code chip-sign transition, and thus it is given by:

\[
T_d = V \cdot T_u = V \cdot M T_{\text{cob}}
\]

\[
(21)
\]

where \( T_d \) is the secondary code chip duration, and \( T_d = 10 \, \text{ms} \).
Also, \( V \in N^+ \), \( T_u \) is the update cycle of the adaptive filter.

Referring to Ward [10], the frequency lock detection method of the adaptive filter is given by Equation (22). Also in Equation (22), \( \alpha = 0.05 \), and \( X = \text{ceil}(M/2) + 1 \). In addition, the frequency lock detection threshold is \( L_{TB} \), and \( L_{TB} = 0.97 \). When \( L_{TB}[k] \geq L_{TB} \), the adaptive filter has entered the lock state. On the contrary, the adaptive filter has exited the lock state when \( L_{TB}[k] < L_{TB} \). To improve the accuracy of the
Doppler frequency estimate, the frequency lock detection block outputs the fine Doppler frequency estimate \( \hat{f}_{dp} \) only when the adaptive filter enters the lock state and lasts at least \( UT_w \). Otherwise, the frequency lock detection block outputs nothing. Hence, the \( \hat{f}_{dp} \) is given by:

\[
L_F[k] = \begin{cases} 
0, & L_F[k-1], \\
(1-\alpha)L_F[k-1] + \alpha \cdot \cos \left( 2\pi \frac{\Delta f_d[k]}{2} + \frac{\Delta f_d[k-1]}{2} \right), & \text{locked}
\end{cases}
\]

Moreover, the total time from when the adaptive filter started working until the frequency lock detection block first outputs the fine Doppler frequency estimate is fine Doppler frequency acquisition time \( T_{acq} \), which is given by:

\[
f_{dp} = \begin{cases} 
\frac{1}{UM} \sum_{UM} f_{NCO} - f_{IF}, & \text{locked} \\
\inf, & \text{unlocked}
\end{cases}
\]

Based on Equations (3), (23), (25), and (26), acquisition accuracy is defined as:

\[
D = \frac{1}{1000 \cdot P_{acq}} \sum_{i=1}^{1000} \left( f_d[0] - \hat{f}_{dp,i} \right) \cdot \noninf(\hat{f}_{dp,i}) \quad \text{for all } i
\]

where \( D \) is the mean absolute deviation between fine Doppler frequency estimate \( \hat{f}_{dp} \) and real input Doppler frequency \( f_d \). The smaller the \( D \), the higher the acquisition accuracy. Eventually, acquisition time based on Equations (24) and (25) is defined as:

\[
T_{acq} = \frac{1}{1000 \cdot P_{acq}} \sum_{i=1}^{1000} T_{acq,i} \quad \text{for all } i
\]

where \( T_{acq} \) represents the mean acquisition time.

Furthermore, the proposed algorithm is compared with a generic second-order (2-ord) FLL shown in Figure 4, because the 2-ord FLL is usually used for fine Doppler frequency acquisition [10]. Based on the Laplace domain, the transfer function of the 2-ord loop filter of the 2-ord FLL is given by:

\[
F(s) = \frac{k_1}{s} + \frac{k_2}{s^2}
\]

where coefficients \( k_1 \) and \( k_2 \) are concerned with equivalent noise bandwidth \( B_F \) of the 2-ord FLL, as Equation (30) shows.

\[
\left\{ \begin{array}{l}
k_1 = \frac{8}{3} B_F \\
k_2 = \frac{1}{2} k_1^2
\end{array} \right.
\]

Referring to Ma et al. [32], \( B_F = 10 \) Hz. More detailed information about FLL can be found in Kaplan [33].
Referring to Ward [10], the frequency lock detection method of the 2-ord FLL is given by Equation (31). Also in Equation (31), \( \alpha = 0.01 \), and \( L_{Th} = 0.8 \):

\[
L_F[k] = \begin{cases} 
0, & k = 0 \\
\alpha \cdot \cos \left( 2\pi \frac{\hat{f}_e[k] + \hat{f}_e[k-1]}{2} \right) + (1 - \alpha) \cdot L_F[k-1], & k \geq 1 
\end{cases}
\]  

In general, the error of the Doppler frequency estimation after coarse acquisition is a few hundred hertz [34]. For instance, \( \Delta f_d \in [-250, 250] \) in hertz. In terms of the proposed fine Doppler frequency acquisition algorithm, \( T_{coh} = 1 \) ms, \( M = 10 \), \( U = 10 \), \( \gamma = 0.95 \), \( \lambda_0 = 0.95 \), and \( \beta = 0.74 \). As far as the 2-ord FLL, the predetection integration time (PIT) is 1 ms. Taking \( \Delta f_d = 250 \) Hz and \( C/N_0 = 45 \) dB-Hz as an example, the fine Doppler frequency acquisition results given by the proposed algorithm and the
2-ord FLL after a Monte Carlo simulation are shown as Figure 5.

As Figure 5 illustrates, the convergence rate of the proposed algorithm is much faster than that of the 2-ord FLL. Before obtaining a fine Doppler frequency estimate \( \hat{f}_{dp} \), there should be a delay of 1070 iterations for the 2-ord FLL, because the 2-ord FLL tends to converge steadily after approximately 1070 iterations when the PIT is 1 ms. As a consequence, in terms of the 2-ord FLL, \( \hat{f}_{dp} = 252.23 \text{ Hz} \) and \( T_{acq} = 1170 \text{ ms} \). For the proposed algorithm, \( \hat{f}_{dp} = 249.33 \text{ Hz} \) and \( T_{acq} = 248 \text{ ms} \).

Similar to this example, Monte Carlo simulations based on different \( \Delta f_d \) are performed, and the results are shown in Figures 6–8. As Figure 6 shows, when considering \( P_{acq} \geq 90\% \), the proposed algorithm has an acquisition sensitivity of approximately 30 dB-Hz, and the 2-ord FLL has an acquisition sensitivity of approximately 25 dB-Hz. As Figure 7 illustrates, the mean absolute deviation \( D \) of \( \hat{f}_{dp} \) is significantly lower for the proposed algorithm than for the 2-ord FLL, especially at higher carrier-to-noise ratios (C/N₀). Finally, as Figure 8 shows, the mean acquisition time \( T_{acq} \) is also significantly reduced for the proposed algorithm, especially at lower C/N₀ ratios.
sensitivity of approximately 34 dB/Hz. As Figure 7 shows, although the acquisition accuracy of the proposed algorithm is slightly lower than that of the 2-ord FLL when $C/N_0 \geq 40 \, dB-Hz$, the acquisition accuracy of the proposed algorithm gradually becomes increasingly higher than that of the 2-ord FLL when $C/N_0 < 40 \, dB-Hz$. As Figure 8 shows, the acquisition time of the proposed algorithm is much less than that of the 2-ord FLL. It should be noted that $D = 10 \, Hz$ and $T_{a,c} = 2000 \, ms$ actually means the non-existence of acquisition accuracy and time, respectively, because the $P_{acq}$ is small and gradually tends towards zero when $C/N_0 < 34 \, dB-Hz$.

In addition, $P_{fa} = 10^{-5}$ and $P_d = 90\%$, and thus the critical carrier-to-noise ratio $C/N_0$ correspond to Equations (6) and (7), as shown as Figure 9. Every combination in Figure 9 between coherent integration time $T_p$ and the number of non-coherent integration $I$ represents a kind of strategy of the coarse acquisition. Hence, the proposed algorithm is able to work well under many different coarse acquisition strategies.

$$\text{FIGURE 9} \quad \text{Figure of critical carrier-to-noise ratio } C/N_0 \text{ based on different combinations between coherent integration time } T_p \text{ and the number of non-coherent integration } I.$$

6 | CONCLUSIONS

An adaptive filter-based fine Doppler frequency acquisition algorithm is proposed for a BDS-3 B1C signal. The proposed algorithm is a kind of two-step algorithm. Based on coarse acquisition estimates of code phase and Doppler frequency, the proposed algorithm tracks the input BDS-3 B1C signal and gives the final fine Doppler frequency estimate.

First, the proposed algorithm has an acquisition sensitivity of 30 dB-Hz, and it is approximately 4 dB better than that of the 2-ord FLL when considering that the acquisition success rate is greater than 90%, that is $P_{acq} \geq 90\%$. Thus, the proposed algorithm has high acquisition sensitivity and works well under many different coarse acquisition strategies that represent different combinations between a coherent integration time and the amount of non-coherent integration.

Second, the mean absolute deviation of the fine Doppler frequency estimate is less than 2 Hz and the acquisition success rate is approximately 100%, when $C/N_0 \geq 40 \, dB-Hz$. Even if $C/N_0 = 30 \, dB-Hz$, the mean absolute deviation of the fine Doppler frequency estimate is only approximately 6 Hz and the acquisition success rate is approximately 90%. Although the acquisition accuracy of the proposed algorithm is slightly lower than that of the 2-ord FLL when $C/N_0 \geq 40 \, dB-Hz$, it gradually becomes increasingly higher than that of the 2-ord FLL when $C/N_0 < 40 \, dB-Hz$. Hence, the proposed algorithm has high acquisition accuracy.

Third, the mean acquisition time of the proposed algorithm is around 250 ms when $C/N_0 \geq 35 \, dB-Hz$, and it is less than 500 ms even if $C/N_0 = 30 \, dB-Hz$. The mean acquisition time of the proposed algorithm is much less than that of the 2-ord FLL, because the convergence rate of the proposed algorithm is much faster than that of the 2-ord FLL. Overall, the proposed algorithm is better than the 2-ord FLL.

This work deduces all related equations concerned with the proposed algorithm, including the adaptation algorithm, acquisition principles, and evaluation methods. Only one specific set of parameters is used to verify the performance of the proposed algorithm. There is no doubt that different sets of parameters can be obtained based on those equations to bring about a different performance and then satisfy different requirements and applications, which is an important contribution of this work.

Finally, the proposed algorithm can be used to acquire other GNSS signals accurately by the appropriate configuration of related parameters.

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