An innovative FPGA-based low-complexity and multi-constellations compatible GNSS acquisition scheme

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Abstract
There is a strong demand for the multi-constellations compatible Global Navigation Satellite System (GNSS) acquisition scheme, since it is able to acquire signals from different constellations to increase availability of satellites. However, the presence of multiple modulation modes and diverse Pseudo-Random-Noise (PRN) code lengths makes the design challenging. Moreover, existing schemes consume a lot of hardware resources. Hence, we present an innovative Field Programmable Gate Array (FPGA)-based low-complexity and multi-constellation compatible GNSS acquisition scheme to provide a solution for the above-mentioned challenges. This scheme is based on the proposed Improved Serial-Parallel Matched Filter structure that not only requires less hardware resources than conventional structures but also performs all operations in a pipeline to simplify the implementation in FPGA. Additionally, the maximum likelihood criterion is used to obtain decision statistics, which ensures the compatibility between BPSK and sBOC (1,1) signals. Furthermore, a three-step method to acquire signals with a long PRN code is proposed. This also guarantees the compatibility among signals with different PRN code lengths. Finally, a new aided acquisition method to accelerate the acquisition process of signals with a long PRN code is proposed. Experimental results show this scheme is capable of acquiring multi-constellations civil GNSS signals, and therefore it has high practical value.

1 INTRODUCTION
Global Navigation Satellite System (GNSS) receivers greatly benefit from the development of the GNSS [1], and now the GNSS includes four independent global constellations: Global Positioning System (GPS), Global Navigation Satellite System (GLONASS), European GNSS (Galileo) and BeiDou Navigation Satellite System (BDS). In fact, these constellations keep enough available navigation satellites in view. When these constellations are combined, navigation solution availability is significantly improved even in a hard situation like urban canyons [2]. It is apparent that Multi-Constellation Compatible (MCC) GNSS receivers have higher reliability in operation than conventional GPS receivers, and therefore the MCC GNSS receiver is the trend in the future [3,4].

However, the presence of modern GNSS signals whose modulation modes and Pseudo-Random-Noise (PRN) code lengths are different from legacy signals makes difficulties for the compatible design. Consequently, modern GNSS signals require a new architecture for MCC GNSS receivers [5]. Additionally, in an GNSS receiver, acquisition is a crucial step of baseband signal processing to coarsely estimate the code phase and Doppler frequency of incoming signals [6], and therefore the MCC GNSS acquisition scheme attracts much attention. Hence, an innovative FPGA-based low-complexity MCC GNSS acquisition scheme is proposed to meet the demand.

At present, the short-time correlation and the FFT scheme are popular [7], and there are mainly three types of acquisition engine architectures in terms of hardware-based MCC GNSS receivers. As reference [8] shows, the first type is based on numerous parallel correlator channels to search code phases simultaneously, and every correlator channel contains one accumulator to carry out coherent integration, and therefore, lots of multiplexers and coherent RAM blocks are used to buffer short-time coherent results before FFT operation, which means a heavy consumption of hardware resources.

As references [9-11] show, the second type is based on the Serial-Parallel Matched Filter (SPMF) structure, but it also demands a large-size coherent RAM to buffer short-time
coherent integration results. The third type is based on the Partial Matched Filter (PMF) structure [12,13], yet it requires many complex multi-inputs adders and a high-speed FFT.

In order to avoid these drawbacks of the above-mentioned three types of architectures, we propose the improved SPMF (ISPMF) structure that is different from the conventional SPMF and PMF structures. Compared with these two structures, the ISPMF not only has the lowest complexity, but also performs all operations in a pipeline so as to simplify the implementation in FPGA.

Additionally, the binary offset carrier (BOC) modulation was proposed in the literature [14] for the modernization of the GPS, and it achieves the desired objective of splitting the spectrum of the code. However, the BOC modulation, due to the sub-carrier, leads to the multi-peak problem about the auto-correlation function (ACF), which may cause the degradation of acquisition precision. Since the sine-BOC (1,1) modulation is widely used for new civil GNSS signals such as GPS L1C, Galileo E1, and BDS-3 B1C, unambiguous acquisition of sBOC(1,1) signals becomes a challenging issue for an MCC GNSS acquisition scheme. Many methods were proposed in the past decades [15–17]: BPSK-like [18], SCPC [19], ASPeCT [20], GRASS [21], PUDLL [22], SRSA [16], and SEA [17]. On the one hand, these methods all require extra specific system designs or reconstructed auxiliary signals, which increase the complexity of the acquisition scheme. On the other hand, the BOC modulation is different from the Binary Phase Shift Keying (BPSK) modulation that is used for legacy signals such as GPS L1 C/A and BDS B1I, and these extra specific system designs or reconstructed auxiliary signals are not suitable for BPSK signals, which makes the compatible design concerned with BPSK and sBOC(1,1) signals challenging. Thus, we propose applying the maximum likelihood criterion for obtaining final acquisition decision statistics in this scheme to ensure low complexity and the compatibility between BPSK and sBOC(1,1) signals. Since we establish an appropriate linear representations about several random variables on the ACF of sBOC(1,1) signals, it is possible to do the analyses and Monte Carlo simulations to prove that this criterion is effective in achieving the unambiguous acquisition of the sBOC(1,1) signals.

Also, the compatibility among PRN codes with different lengths is another challenging issue for an MCC GNSS acquisition scheme. Present civil GNSS signals have diverse PRN code lengths such as 1023, 2046, 4092 and 10,230. It is noted that different PRN code lengths require different quantities of correlation operation for coherent integration, which is also closely concerned with the number of correlators. The study [14] uses large number of correlators to satisfy the requirement of the longest PRN code, which unfortunately leads to lots of redundancy and low utilization in terms of correlators when acquiring signals with a short PRN code. Furthermore, numerous correlators mean heavy consumption of hardware resources. Hence, based on the low-complexity ISPMF structure, we propose a three-step method to acquire signals with a long PRN code such as the BDS B1C, which also guarantees the compatibility with signals with different PRN code lengths.

Moreover, low complexity of an acquisition scheme generally means high cost of processing time especially for signals with a long PRN code, and therefore many aided acquisition methods are widely used to speed up the acquisition process of signals with a long PRN code [23]. However, these existing methods are merely based on two types of auxiliary information about code phase and Doppler frequency respectively. In order to accelerate the acquisition process further, we propose a new aided acquisition method that is based on not only two types of auxiliary information mentioned above, but also the proposed third type of auxiliary information about the signal strength, which reduces the redundant number of non-coherent integration.

The rest of this article is organized as follows. First, the signal model and basic acquisition principles are introduced. Then, based on the maximum likelihood criterion, the acquisition decision strategy is designed and it is proved to be effective on unambiguous acquisition of sBOC(1,1) signals. Subsequently, the proposed ISPMF structure is given, and then the innovative MCC GNSS acquisition scheme and the proposed three-step method are described in detail. Furthermore, the new aided acquisition method is explained clearly. Finally, based on real GNSS signals data, the effectiveness and performance of the proposed acquisition scheme and aided acquisition method are verified by experiments, and some conclusions are summarized.

2 | SIGNAL MODEL

Generally, the signal received by a GNSS receiver antenna can be described as [24]

\[
r(t) = \sum_{i=1}^{M} A_i d_i(t - \tau_i)c_i(t - \tau_i)\cos[2\pi(f_L + f_{d_i})t + \phi_i] + \eta(t)
\]

where \(r(t)\) is the sum of \(M\) signals, the subscript \(i\) represents the \(i\)th signal, \(A_i\) is the amplitude, \(d_i\) represents the data message, \(c_i\) is PRN code or the product of PRN code, secondary code and sub-carrier, \(\tau_i\) represents the path delay, \(f_{d_i}\) is the Doppler frequency, \(f_L\) denotes the carrier center frequency, \(\phi_i\) is the initial phase, and \(\eta(t)\) is an additive white Gaussian noise with two-sides noise power spectrum density of \(N_0/2\ W/Hz\).

After \(r(t)\) has been exported by the antenna to later stages, \(r(t)\) is filtered, amplified, down converted to the intermediate frequency (IF), filtered again, sampled, and quantized by the radio-frequency (RF) front end. Neglecting the quantization effect, the signal at the output of the RF front end is given by

\[
r[n] = \sum_{i=1}^{M} A_i d_i[n - \tau_i]c_i[n - \tau_i]\cos[2\pi(f_{IF} + f_{d_i})nT_s + \phi_i] + \eta[n]
\]

(2)
where \([m] = \tau[nT]\) is a discrete time sequence with a sampling frequency \(f_s = 1/T_s\), and \(f_{IF}\) is the IF of the receiver. The \(\tau[n]\) is the input of the acquisition scheme.

3 | ACQUISITION PRINCIPLES

Signal acquisition is a procedure that conducts a 2D search over the time-frequency space for every PRN number as Figure 1 shows. Short-time coherent integration function \(F(\cdot)\) is obtained after the correlation operation for \(k\)th signal \((k\in[1, M])\), and it can be written as

\[
F_k(\tau,\hat{f}_d, p) = \sum_{n=pm}^{pm+m-1} r[n]c_k[n-\tau] \exp[j2\pi(f_{IF} + \hat{f}_d)nT_s]
\]

\[
= A_k R_k(\Delta \tau) \sin(\pi \Delta f_d T_c) \frac{\sin(\pi \Delta f_d T_c)}{2} \exp[j\pi \Delta f_d (2pm + m - 1)T_s - j\phi_c] + \eta_p
\]

(3)

where \(T_c = mT_s\) is the coherent-integration time of the \(p\)th short-time coherent integration, \(\exp[j2\pi(f_{IF} + \hat{f}_d)nT_s]\) is the local complex carrier, \(c_k[n-\tau]\) is the local PRN code replica for BPSK signals or the product of PRN code, secondary code and sub-carrier in terms of sBOC(1,1) signals, \(\tau\) and \(\hat{f}_d\) are the estimates of code phase and Doppler frequency respectively, \(R_k(\Delta \tau)\) is the corresponding ACF, and \(\eta_c\) is the present noise term. Then, it is derived that \(\Delta \tau = \tau - \tau_k\) and \(\Delta f_d = \hat{f}_d - f_{d,k}\). According to central limit theorem, \(\eta_p\) is a Gaussian random variable.

Subsequently, \(N\) points FFT operation is performed on \(P\) consecutive short-time coherent-integration results followed by \(P\) zeros, since \(N = 2P\), which implies the effort of improving frequency resolution and reducing scalloping loss [6], and then the \(n\)th \((n \in [0, N - 1])\) output of the \(l\) times FFT operation is given by:

\[
G_k(\tau,\hat{f}_d, n, l) = \sum_{p=0}^{p-1} F_k(\tau,\hat{f}_d, p \pm \frac{1}{2}) \exp\left(-j2\pi n p \frac{\tau}{N}\right)
\]

\[
= A_k R_k(\Delta \tau) \sin(\pi \Delta f_d T_c) \frac{\sin(\pi \Delta f_d T_c)}{2} \exp[j\pi \Delta f_d (2pm + m - 1)T_s - j\phi_c] + \eta_p
\]

(4)

where \(\phi_{n,l}\) is the residual carrier phase, \(\eta_{n,l}\) is noise term. According to the central limit theorem, \(\eta_{n,l}\) is also complex Gaussian random variable with zero mean and variance \(2\sigma^2\), and the real and imaginary parts of the \(\eta_{n,l}\) are independent and with zero mean and equal variance \(\sigma^2\). In addition, taking Equation (4) into consideration, the searching range of frequency is \([-1/2T_c, 1/2T_c]\), and the frequency resolution is \(1/NT_c\).

In general, non-coherent integration is an effective way to extend the integration time, and non-coherent integration function \(S(\cdot)\) can be derived as:

\[
S_k(\tau,\hat{f}_d, n) = \sum_{l=0}^{L-1} |G_k(\tau,\hat{f}_d, n, l)|^2
\]

(5)

where \(L\) is the number of non-coherent integration, and \(l \in [0, L-1]\).

According to Equation (4), there is a hypothesis \(H_0\) when the local PRN code replica does not align with the PRN code of the received \(k\)th signal or the searched satellite is absent, that is, \(R_k(\Delta \tau) \rightarrow 0\) [25], and Equation (4) can be rewritten as

\[
G_k(\tau,\hat{f}_d, n, l) = \eta_{n,l}
\]

(6)

At this point, the random variable \(S_k(\tau,\hat{f}_d, n)\) follows the central chi-square \((\chi^2)\) distribution with \(2L\) degrees of freedom.

On the contrary, there is another hypothesis \(H_1\) when the local PRN code replica aligns with PRN code of the received \(k\)th signal, that is, \(R_k(\Delta \tau) \rightarrow 1\), and Equation (4) can be rewritten as

\[
G_k(\tau,\hat{f}_d, n, l) = \hat{A}_k R_k(\Delta \tau) \exp(j\phi_{n,l}) + \eta_{n,l}
\]

(7)

where \(\hat{A}_k = \frac{A_k \sin(\pi \Delta f_d T_c)}{2 \sin(\pi \Delta f_d T_c)} \frac{\sin[(\Delta f_d T_c - \hat{\Delta})_{\tau}]}{\sin[\pi (\Delta f_d T_c - \hat{\Delta})_{\tau}]}\). When \(R_k(\Delta \tau) \rightarrow 1\), that is \(\Delta \tau = 0\), the signal-to-noise ratio (SNR) at the output of FFT operation can be expressed as \(\text{SNR} = \frac{\hat{A}_k^2}{(2\sigma^2)}\). It is worth noting that the influence of \(\phi_{n,l}\) on \(S_k(\tau,\hat{f}_d, n)\) is negligible when \(\text{SNR}\) satisfies the acquisition threshold, and therefore we assume \(\phi_{n,l} = \phi_c\). At this moment, \(S_k(\tau,\hat{f}_d, n)\) follows the non-central \(\chi^2\) distribution with \(2L\) degrees of
SNR = F_{\eta}^{-1}\left[F_{\eta}^{-1}(1 - P_{fa}, L), 1 - P_d, L\right] (8)

where $F_{\eta}($) is the cumulative distribution function of central $\chi^2$ distribution, $F_{\eta}($) is the cumulative distribution function of non-central $\chi^2$ distribution, and $F_{\eta}^{-1}(\cdot)$ and $F_{\eta}^{-1}(\cdot)$ are the inverses of function $F_{\eta}(\cdot)$ and $F_{\eta}(\cdot)$, respectively. This SNR represents the minimum detectable SNR value of received signals as $P_{fa} = 10^{-5}$ and $P_d = 0.9$, and therefore this SNR is called as the ‘critical SNR’ hereinafter.

# 4 ACQUISITION DECISION

The maximum likelihood criterion is applied for obtaining the final acquisition decision statistic in this acquisition scheme to ensure the compatibility between the BPSK and sBOC(1,1) signals.

According to previous acquisition principles, every bin in the time-frequency space corresponds to a non-coherent integration value $S_k(\tilde{r}, \tilde{f}_d, n)$, and then the maximum and mean values among all non-coherent integration values can be obtained and denoted as $S_{k, max}(\tilde{r}, \tilde{f}_d, n)$ and $S_{k, mean}(\tilde{r}, \tilde{f}_d, n)$, respectively. Afterwards, the Peak-Mean Ratio (PMR) is derived as

$$PMR = \frac{S_{k, max}(\tilde{r}, \tilde{f}_d, n) - S_{k, mean}(\tilde{r}, \tilde{f}_d, n)}{S_{k, mean}(\tilde{r}, \tilde{f}_d, n)} (9)$$

The PMR is compared with the predefined threshold $Tb$ to make an acquisition decision about the presence or absence of the searching satellite. If the searched satellite is in view, the bin mapped to the $S_{k, max}(\tilde{r}, \tilde{f}_d, n)$ will be taken as the correct bin, and the values corresponding to the correct bin will be taken as the final rough estimates of code phase and Doppler frequency respectively.

It is obvious that the maximum likelihood criterion works well for BPSK signals to lock the ACF main peak, since the ACF of BPSK signals only contains one peak. However, there are no other articles to claim that the maximum likelihood criterion also works well for the unambiguous acquisition of sBOC(1,1) signals until now. As a consequence, we prove this criterion is able to consistently achieve the unambiguous acquisition of sBOC(1,1) signals.

The sub-carrier of the sBOC(1,1) modulation is expressed as $\text{sign} \left[ \sin(2\pi f_c t) \right]$, where $\text{sign} [.]$ is the sign function, and $f_c = 1.023$ MHz. As Figure 2 shows, the ACF of sBOC(1,1) signals has multiple peaks, including one main peak and two side peaks, which constitutes the so-called multi-peaks problem.

![Auto-correlation function of sBOC(1,1) signals](image)

According to Equation (7), for sBOC(1,1) signals, there are three random variables $G(k, \tilde{r}_m, \tilde{f}_d, n, l)$, $G(k, \tilde{r}_1, \tilde{f}_d, n, l)$, and $G(k, \tilde{r}_2, \tilde{f}_d, n, l)$, corresponding to the main peak, side peak 1, and side peak 2 respectively, and their ACF terms are $R(\Delta \tau_{m})$, $R(\Delta \tau_{1})$, and $R(\Delta \tau_{2})$ respectively. In addition, $G(k, \tilde{r}_m, \tilde{f}_d, n, l)$ correlates with $G(k, \tilde{r}_1, \tilde{f}_d, n, l)$ and $G(k, \tilde{r}_2, \tilde{f}_d, n, l)$, and $G(k, \tilde{r}_1, \tilde{f}_d, n, l)$ is independent of $G(k, \tilde{r}_2, \tilde{f}_d, n, l)$. The coefficients of association among these random variables are given by Equation (10)

$$P_i = \text{Pr}\{S_k(\tilde{r}_m, \tilde{f}_d, n)\} = 1 \quad \text{max}\{S_k(\tilde{r}_1, \tilde{f}_d, n), S_k(\tilde{r}_2, \tilde{f}_d, n)\} \quad H_1, SNR, H_1, H_2, H_3\}
\begin{align*}
\rho_{G_{\tilde{r}_m, \tilde{r}_1}} & = \rho_{G_{\tilde{r}_m, \tilde{r}_2}} = -1/2 \\
\rho_{G_{\tilde{r}_1, \tilde{r}_2}} & = 0
\end{align*}
\quad (10)$$

Also, we construct six independent Gaussian random variables: $u_r, u_Q, v_r, v_Q, w_r$, and $w_Q$, and they are denoted as $u_r \sim N(0, \sigma^2/4)$, $u_Q \sim N(0, \sigma^2/4)$, $v_r \sim N(0, \sigma^2/4)$, $v_Q \sim N(0, \sigma^2/4)$, $w_r \sim N(0, \sigma^2/2)$, and $w_Q \sim N(0, \sigma^2/2)$, respectively. Then, another six Gaussian random variables are constructed as

\begin{align*}
X_I & = u_I + v_I + w_I \\
X_Q & = u_Q + v_Q + w_Q \\
Y_I & = u_I - v_I - w_I \\
Y_Q & = u_Q - v_Q - w_Q \\
Z_I & = v_I - u_I - w_I \\
Z_Q & = u_Q - v_Q - w_Q
\end{align*}
\quad (11)
where these random variables are denoted as $X_t \sim N(0, \sigma^2)$, $X_\omega \sim N(0, \sigma^2)$, $Y_t \sim N(0, \sigma^2)$, $Y_\omega \sim N(0, \sigma^2)$, $Z_t \sim N(0, \sigma^2)$, and $Z_\omega \sim N(0, \sigma^2)$, respectively. Subsequently, six coefficients of association are given as

$$
\begin{align*}
\rho_{X_1, Y_1} &= \rho_{X_Q, Y_Q} = -1/2 \\
\rho_{X_1, Z_1} &= \rho_{X_Q, Z_Q} = -1/2 \\
\rho_{Y_1, Z_1} &= \rho_{Y_Q, Z_Q} = 0
\end{align*}
$$

As a consequence, $G_k(\tau_m, \hat{f}_d, n, l)$, $G_k(\tau_1, \hat{f}_d, n, l)$, and $G_k(\tau_2, \hat{f}_d, n, l)$ can be equivalently expressed as

$$
\begin{align*}
G_k(\tau_m, \hat{f}_d, n, l) &= A \cdot R(\Delta \tau_m) + X_I + jX_Q \\
G_k(\tau_1, \hat{f}_d, n, l) &= A \cdot R(\Delta \tau_1) + Y_I + jY_Q \\
G_k(\tau_2, \hat{f}_d, n, l) &= A \cdot R(\Delta \tau_2) + Z_I + jZ_Q
\end{align*}
$$

where $A = \sqrt{2 \cdot SNR \cdot \sigma^2}$. Since the influence of $\phi_n$ on $S_k(\tau_m, \hat{f}_d, n)$ is negligible when $SNR$ satisfies the acquisition threshold. It is assumed that $\phi_n = \pi/4$, and then $S_k(\tau_m, \hat{f}_d, n)$ is given by

$$
\begin{align*}
S_k(\tau_m, \hat{f}_d, n) &= \sum_{l=0}^{L-1} |G_k(\tau_m, \hat{f}_d, n, l)|^2 \\
S_k(\tau_1, \hat{f}_d, n) &= \sum_{l=0}^{L-1} |G_k(\tau_1, \hat{f}_d, n, l)|^2 \\
S_k(\tau_2, \hat{f}_d, n) &= \sum_{l=0}^{L-1} |G_k(\tau_2, \hat{f}_d, n, l)|^2
\end{align*}
$$

When zero-IF data rate is 4.092 Msp and PRN code chip rate is 1.023 Msp, according to the practical acquisition process, $\Delta \tau_m$ is uniformly distributed in the interval $[-1/8, 1/8]$ chip. In order to simplify the analysis, we mainly consider three situations shown in Table 1, and the occurrence probabilities of these three situations are set to $Pr(H_{1,i}) = Pr(H_{2,i}) = Pr(H_{3,i}) = 1/3$.

Then the probability of locking the ACF main peak while making acquisition decision for sBOC(1,1) signals is approximately given by:

$$
P = \int_{-\frac{1}{2}}^{\frac{1}{2}} P_i d\Delta \tau_m \approx \sum_{i=1}^{3} P_i \times Pr(H_{1,i}) = \frac{1}{3} \times \sum_{i=1}^{3} P_i
$$

where $P_i$ is expressed as Equation (16).

Based on Equations (13)–(16), the value of the $P$ can be computed via Monte Carlo simulation, and the result is shown as Figure 3.

Although the probability $P$ is based on the critical SNR, it is also greater than 94% when the number of non-coherent integration $L$ is less than 100. Even in worst situation, the probability $P_i$ is still greater than 86%, and with the growth of the SNR, $P$ and $P_i$ increase further. It is noted that only three key samples are considered for every situation while ignoring few other samples, but the probability $P$ obtained here is still a good approximation of the real probability due to the specific and closely correlating relation among all of these samples.

### Table 1 Key parameters about three situations

| Situation | Ideal | Worse | Worst |
|-----------|-------|-------|-------|
| Index $i$ | 1     | 2     | 3     |
| Hypothesis | $H_{1,i}$ | $H_{2,i}$ | $H_{3,i}$ |
| $\Delta \tau_m$ | 0     | -1/16 | -1/8  |
| $R(\Delta \tau_1)$ | 1     | 13/16 | 10/16 |
| $R(\Delta \tau_2)$ | -1/2  | -7/16 | -6/16 |
| $R(\Delta \tau_3)$ | -1/2  | -5/16 | -2/16 |
| Probability | $Pr(H_{1,i})$ | $Pr(H_{2,i})$ | $Pr(H_{3,i})$ |

![Critical SNR](image)

**Figure 3** Probability of locking the ACF main peak for sBOC(1,1) signals

## 5 ISPMF Structure

First, we introduce the conventional PMF structure that is commonly used for acquisition schemes. However, Figure 4 shows the conventional PMF structure has two palpable drawbacks. (1) The $m$-inputs complex adder is much more complex than a two-inputs complex adder with respect to the same input width, and the complexity of the $m$-inputs complex adder increases with the growth of $m$. (2) All $P$ parallel $m$-inputs complex adders export $P$ complex values simultaneously at one clock, which leads to the extreme demand for a high-speed FFT.

Similarly, the conventional SPMF structure is also commonly used for acquisition schemes. As Figure 5 shows, the coherent integration RAM is used to store short-time coherent integration results, and it contains at least $2mP$ memory cells. In addition, every candidate code phase corresponds to $P$ short-time coherent integration results, and only $m$ candidate code phases can be inspected at a single time. As a consequence, this search needs to be repeated $P$ times to inspect all $mP$ candidate code phases, which makes the control
In brief, the conventional SPMF structure also has two palpable drawbacks: (1) large-size coherent integration RAM and (2) repeat searches.

In order to avoid above-mentioned drawbacks and obtain a low-complexity 2D searching structure, we propose the ISPMF structure that is shown in Figure 6. Although the ISPMF structure is transformed from the conventional SPMF structure, it is able to output $P$ short-time coherent integration results continuously corresponding to a candidate code phase by means of a special playback order of the zero-IF data, and all candidate code phases are inspected consecutively.

As Figure 6 shows, the ‘SRL$_P$

$SRL_P = [V_{j1}, V_{j2}, ... , V_{jp}]$

where $V_{jp}$ ($p \in [1,P]$, and $p \in Z$) is the $p$th element that is stored in the $p$th ‘Reg’ of the ‘SRL$_P$’. And after initialization, $V_{jp}$ is derived as

$V_{jp} = V_{j-1,p} + D[(p-1)m + i + j - 1]$

where $j \in [1,m]$, and $V_{0,p} = 0$.

On the one hand, the ISPMF performs all operations in a pipeline. On the other hand, the ‘Reg’ can be implemented by Look-Up Tables (LUTs) or slice registers in FPGA, and therefore the synthesis of all SRLs is easier than that of coherent integration RAM with respect to high-speed acquisition. Additionally, $2mP$ RAM cells occupy two times the bits space than $mP$ registers with respect to same data bits width. As a consequence, the ISPMF structure has lowest complexity when compared with the conventional PMF and SPMF structures, which is shown as Table 2.

It is worth noting that the ISPMF is different from the folding matched filter structure that is based on the time division multiplexing strategy, since the correlation operations and the playback of the zero-IF data all run at the same frequency in terms of the ISPMF.

6 | MCC GNSS ACQUISITION SCHEME

First of all, the MCC GNSS acquisition scheme is based on the above-mentioned acquisition principles, acquisition decision strategy, and the ISPMF structure, which ensures the MCC GNSS acquisition scheme has low complexity.

The MCC GNSS acquisition scheme mainly focuses on acquiring signals including GPS L1 C/A, GPS L1C, BDS B1I, BDS-3 B1C, Galileo E1, and GLONASS L1, as these signals are from four main global constellations and are most widely used. The characteristic parameters of these signals are shown in Table 3. It is noted that only the BOC(1,1) component is used for acquisition in terms of Galileo E1B, GPS L1Cp, and BDS-3 B1Cp, because the power of BOC(1,1) components is much greater than that of their BOC(6,1) components. Therefore, acquiring BOC(1,1) components has high cost performance in terms of common services. In fact, these signals have different modulation modes, chip rates, and PRN code lengths. Consequently, the similarities and differences among these signals should be fully considered in this design.
TABLE 2  Quantity comparison of main components and operations among three structures. An m-income adder consists of m–1 2-input adders in implementation, and the storage space in bit occupied by mP Coherent RAM cells is as large as that occupied by mP Regs with respect to same precision.

| Structure | 2-input adder | m-income adder | Reg | Coherent RAM cells | N-points FFT | Search repetitions |
|-----------|---------------|----------------|-----|---------------------|--------------|-------------------|
| PMF       | 0             | P              | mP  | 0                   | 1            | 1                 |
| SPMF      | m             | 0              | m   | ≥2mP               | 1            | P                 |
| ISPMF     | m             | 0              | mP  | 0                   | 1            | 1                 |

Abbreviation: ISPMF, Improved Serial–Parallel Matched Filter; PMF, Parallel Matched Filter; SPMF, Serial–Parallel Matched Filter.

TABLE 3  Characteristic parameters of typical multi-constellations GNSS signals in this study.

| GNSS signal | PRN code | Length | Period | Secondary code chip width | Modulation mode | Carrier frequency |
|-------------|----------|--------|--------|---------------------------|-----------------|------------------|
| GPS L1 C/A  |          | 1023   | 1 ms   | -                         | BPSK-R(1)       | 1575.42 MHz      |
| L1C         |          | 10230  | 10 ms  | 10 ms                     | L1Cp: TBOC(6,1,4/33) | 1575.42 MHz      |
| GLONASS L1  |          | 511    | 1 ms   | -                         | BPSK-R(1)       | ~ 1602 MHz       |
| Galileo E1  |          | 4092   | 4 ms   | 4 ms                      | E1B: CBOC(6,1,1/11) | 1575.42 MHz      |
| BDS B1I     |          | 2046   | 1 ms   | --                       | BPSK-R(2)       | 1561.098 MHz     |
| BDS B1C     |          | 10230  | 10 ms  | 10 ms                     | B1Cp: QMBOC(6,1,4/33) | 1575.42 MHz      |

On the one hand, GPS L1 C/A and GLONASS L1 both adopt the BPSK-R(1) modulation mode, and BDS B1I adopts the BPSK-R(2) modulation mode, which means most of the power of these three signals is within the range of ±2.046 MHz relative to respective central frequencies. On the other hand, the spectrum of BOC(1,1) signals is split into two symmetrical components with no remaining power at the central frequency, and two main lobes are centred at ±1.023 MHz relative to the central frequency. Thus a minimum bandwidth of 4.092 MHz is needed to keep most of the power for BOC(1,1) signals [27]. As a consequence, the zero-IF data rate is set to 4.092 Msps so as to reduce the complexity of the acquisition scheme.

As we all know, the data bit transition and secondary-code chip transition limit the extension of coherent integration, and the longer the coherent integration time, the higher the complexity of the acquisition scheme. Thus, only 1 ms coherent integration is performed before non-coherent integration, and there are 4092 complex zero-IF data samples within 1 ms.

Additionally, the acquisition speed is mainly in direct proportion to the number of correlators and the implementation frequency in FPGA, and the consumption of hardware resources is also highly sensitive to the number of correlators and the size of the FFT. In order to make a reasonable trade-off between the low complexity, the high processing speed, and wide dynamic searching range of frequency, the parameters of the ISPMF are set as m = 128 and P = 32. ‘Non-IF data RAM’ is used to store zero-IF data, and then the high-speed playback of the zero-IF data is possible to accelerate the acquisition process.

Furthermore, an accumulator is inserted between the last SRL of the ISPMF and the complex FFT to halve the data rate. Then a complex 32-points zero-padding FFT is adopted so as to reduce the scalloping loss, and therefore the whole acquisition block can run at the same frequency \( f_{\text{FPGA}} \), which is conducive to the synthesis of the acquisition scheme. Moreover, the expected Doppler frequency span is ±5 KHz for common stationary ground-based applications. Consequently, for a given candidate code phase, only 21 frequency bins that form a set are required at the output of the FFT, and the frequency resolution is 500 Hz. Then, the ‘non-coherent integration RAM’ should contain 85,932 (i.e. \( 4092 \times 21 \)) memory cells.
Finally, the proposed innovative FPGA-based and low-complexity MCC GNSS acquisition scheme is shown as Figure 7.

7 | THREE-STEP METHOD

Taking signals such as GPS L1C, Galileo E1, BDS-3 B1C into consideration, their PRN code periods are all greater than 1 ms. However, in order to ensure less consumption of hardware resources, only 1 ms coherent integration can be conducted as previously discussed. Hence, we acquire these signals in three steps, which is called the three-steps method.

Step 1 We first divide the long PRN code with period of \( n \) ms into \( n \) short sections, and every short section has a duration of 1 ms.

Step 2 These \( n \) short sections are non-coherently integrated, which is similar with the acquisition process that is performed for the signal whose PRN code period is 1 ms to achieve \( n \)-times non-coherent integration.

Step 3 Only \( 1/n \) code phases are inspected after Step 2. Thus Step 2 needs to be repeated \( n \) times. Each time Step 2 inspects different \( 1/n \) code phases and outputs a \( PMR \). The maximum among these \( n \) \( PMRs \) is taken as the final acquisition decision statistic.

Indeed, the above-mentioned three-step method comes at the cost of the degradation of the ACF performance. When taking the BDS-3 B1C signal with PRN of 1 as an example, the comparison of the ACF performance is shown in Figure 8. Although the short section is significantly inferior to the respective long PRN code in terms of the ACF performance, it is only slightly inferior to the PRN code of BDS B1I signals. As a consequence, the degradation of the ACF performance is acceptable.

In addition, the acquisition time is another key factor to evaluate the performance of the acquisition scheme [28]. As previously discussed, the acquisition time is mainly concerned with crucial procedures during acquisition such as correlation operation, coherent integration, and non-coherent integration. It is noted that the time to load a new PRN code or to modify the carrier, and the latency in the processing are not taken into account, and then the acquisition time in second is approximately derived as:

\[
T_{acq} = \begin{cases} 
[LN_e + (m-1)]P/f_{FPGA} & n = 1 \\
[N_e + (m-1)]PL/f_{FPGA} & n > 1 
\end{cases} \tag{19}
\]

where \( n \) is the number of short sections, \( L \) is the number of non-coherent integration, \( m = 128 \), and \( P = 32 \). Also in Equation (19), \( N_e \) is the number of required candidate code phases in 1 ms. Without aided acquisition methods, \( N_e = 4092 \). On the other hand, with aided acquisition methods, \( N_e = 80 \).

8 | NEW AIDED ACQUISITION METHOD

Signal 1 with a short PRN code such as BDS B1I spend less time on acquisition than signal 2 with a long PRN code such as BDS-3 B1C, since the shorter the PRN code, the fewer the candidate code phases. Furthermore, signals transmitted from the same GNSS satellite shall be coherently derived from a common on-board reference frequency source. A typical coherent relation between two different signals from the same GNSS satellite is shown as Figure 9. In fact, the time interval between the beginning of receiving data for signal 1 and the beginning of receiving data for signal 2 is an integral multiple of 1 ms, or zero. Then the code phase of signal 2 should be:

\[
\tau_2 = \tau_1 + 4092 \cdot k + \Delta \tau \tag{20}
\]

where \( k = 0, 1, 2, \ldots, n-1 \), \( \tau_1 \) is the estimate of code phase of signal 1, \( \Delta \tau \) represents the error caused by several factors like estimation accuracy of the \( \tau_1 \), Doppler shift, receiver clock offset, etc. The Doppler frequency of signal 2 should be

\[
f_{d,2} = f_{d,1} \cdot f_{f,2} / f_{f,1} + \Delta f_d \tag{21}
\]

where \( f_{d,1} \) is the estimate of Doppler frequency of the signal 1, \( f_{f,1} \) and \( f_{f,2} \) are carrier center frequencies of signal 1 and signal 2, respectively, and \( \Delta f_d \) represents the error including the influence of the ionosphere and the estimation accuracy of the \( f_{d,1} \).

Equations (20) and (21) represent two types of commonly used auxiliary information about the code phase and Doppler frequency. In order to accelerate the acquisition process of signal 2 further, the third type of auxiliary information about the signal strength is proposed here to reduce the redundant number of non-coherent integration.

On the one hand, after the successful acquisition of signal 1, the \( PMR_2 \) can be obtained, and the \( PMR_1 \) can be regarded as the rough estimate of the \( SNR_1 \). On the other hand, the minimum received power difference between two different GNSS signals at the output of the same GNSS antenna is prior information. Consequently, according to the \( PMR_1 \) and the power difference between signals 1 and 2, the \( SNR_2 \) can be coarsely obtained, and then the required number of non-coherent integration for signal 2 can be determined. In this way, when the power difference is within the range of \( \pm 3 \) dB, the intended number of non-coherent integration for signal 2 is empirically given by:

\[
L_2 = \begin{cases} 
1, & PMR_1 \geq 20 \\
L_{max}/2, & 15 \leq PMR_1 < 20 \\
L_{max}, & PMR_1 < 15 
\end{cases} \tag{22}
\]

where \( L_{max} (L_{max} \geq 4) \) is the maximum number of non-coherent integration allowed by the acquisition scheme, \( L_2 \) is
The real and imaginary parts of the GNSS signal data are quantized in 4 bits, and the sampling frequency $f_s$ is 30.69 MHz. After down sampling to 4.092 MHz, the zero-IF data is obtained, and its real and imaginary parts are all quantized in 8 bits.

This acquisition scheme runs at the frequency of 130.944 MHz (i.e. $f_{\text{FPGA}} = 32 \times 4.092$ MHz) during experiments. In addition, the ‘zero-IF data RAM’ is designed to contain 81,840 (i.e. $4092 \times 20$) complex memory cells so that it is capable of storing 20 ms zero-IF data. Then the coherent integration time is 1 ms, and the allowed maximum number of non-coherent integration is 10, that is, $L_{\text{max}} = 10$. If the Doppler frequency is 5 kHz, the PRN code chip rate will be shifted by about 3.25 chips/s (i.e. $5000 \times 1.023 / 1575.42$), which indicates that the PRN code replica and the received PRN code will only shift by about 1/30 chip per 10 ms. As a result, this study ignores the shift on the received PRN code chip rate caused by Doppler frequency.

Furthermore, the FPGA-based circuit of the proposed acquisition scheme is designed in very high speed hardware description language, and it is implemented by the Xilinx's Integrated Development Environment (IDE) ‘ISE Design Suite 14.7’ in the FPGA platform of xc7z030-3fgg484. The implementation results are shown in Table 4.

First, the GNSS signals studied here are acquired individually without the proposed new aided acquisition method, and the acquisition results are shown in Figure 10.

It is noted that the Galileo E1B and BDS-3 B1C are acquired in the proposed three-step method. Regrettably, the

**Figure 7** Block diagram of the proposed innovative FPGA-based and low-complexity MCC GNSS acquisition scheme

**Figure 8** Comparison figure of the ACF performance

9 | EXPERIMENTAL RESULTS

In order to verify the effectiveness and performance of the proposed acquisition scheme, unambiguous acquisition of sBOC(1,1) signals based on the maximum likelihood criterion, three-steps method, and aided acquisition method, real GNSS signals data at carrier center frequencies of 1575.42 and 1561.098 MHz were recorded in Beijing China (116° 20'E, 39°56'N) at 06:54:00 Greenwich Mean Time on November 11, 2019 by using SPIRENT GSS6450 front end.

**Figure 9** Typical coherent relation between two different signals from the same satellite
real GNSS signals data does not include the data of the GLONASS L1 due to the limits of the laboratory. Nevertheless, referring to the acquisition process of the GPS L1 C/A and BDS B1I, it is apparent that the proposed acquisition scheme also works for the GLONASS L1. Although the GPS L1C is not broadcasted in the sky at present, it is similar to the BDS B1C in terms of signal structure and signal modulation mode. Hence, referring to the BDS-3 B1C, it is obvious that the proposed acquisition scheme also works for GPS L1C signals.

**TABLE 4** Hardware resources consumption of the proposed acquisition scheme after FPGA implementation, which is based on the Xilinx’s xc7z030-3fbg484 platform

| Items             | The number | Percentage |
|-------------------|------------|------------|
| Slice registers   | 48,644     | 30         |
| Slice LUTs        | 25,828     | 32         |
| Occupied slices   | 13,530     | 68         |
| LUT flip flop pairs | 51,525   |            |
| DSP48E1s          | 18         | 4          |

**FIGURE 10** Acquisition results of GNSS signals studied in this paper

Second, based on the proposed new aided acquisition method, that is, Equations (20)–(22), the BDS-3 B1C is acquired with three types of auxiliary information from the BDS B1I. As Figure 10 shows, the B1I of PRN 23, 24, 25, 30, and 32 are from BDS satellites, and then the comparison between individual and aided acquisitions is shown in Table 5. Table 5 shows that the new aided acquisition method not only significantly reduces the 2D search space, but also decreases the redundant number of non-coherent integration for the BDS-3 B1C with high power, which brings about a significant decline of acquisition time according to Equation (19). Compared with conventional aided acquisition methods merely based on two types of auxiliary information about code phase and Doppler frequency, respectively, the new aided acquisition method is better, since it spends less time on acquiring signals with high power.

10 | CONCLUSIONS

First, since we deduce and establish appropriate linear expression about several random variables of the ACF of sBOC(1,1) signals, it is proven by Monte Carlo simulations that the maximum

**TABLE 5** The comparison between individual and aided acquisitions about the BDS-3 B1C

| BDS-3 B1C: PRN | 23   | 24   | 25   | 30   | 32   |
|----------------|------|------|------|------|------|
| Individual acquisition | $\tau_2$ = 1–40,920; $N_c=4092$  |
|                              | $f_{d2}$ = 5 to 5 KHz  |
|                              | $L_2 = 10$  |
|                              | $T_{\text{eq}} = 100.31$ ms  |
| Aided acquisition | $\tau_2 = \tau_1 + 4092 \cdot k - 40 \sim \tau_1 + 4092 \cdot k + 40; N_c = 80$  |
|                              | $f_{d2} = f_{d,1} - 500$ Hz $\sim f_{d,1} + 500$ Hz  |
|                              | $L_2 = 1$  |
|                              | $T_{\text{eq}} = 0.51$ ms  | $2.27$ ms | $0.51$ ms | $2.27$ ms | $0.51$ ms |
likelihood criterion is capable of achieving unambiguous acquisition of sBOC(1,1) signals with higher probability than 94%. Therefore, applying the maximum likelihood criterion for obtaining decision statistics is able to not only ensure the compatibility between BPSK and sBOC(1,1) signals but also reduce system complexity.

Second, the proposed ISPMF structure has lowest complexity as compared with conventional PMF and SPMF structures, and it makes all operations perform in a pipeline so as to simplify the implementation in FPGA.

Third, in order to reduce the complexity of the acquisition scheme, the proposed three-step method is specially used to acquire GNSS signals with a long PRN code successfully, and it also ensures a wide dynamic searching range of frequency. Hence, the three-step method makes the compatibility among PRN codes with different lengths possible and simple.

Fourth, the proposed new aided acquisition method significantly reduces the acquisition time of signals such as BDS-3 B1C. Since the proposed third type of auxiliary information—signal strength—is taken into account, the proposed new aided acquisition method is better than conventional aided acquisition methods due to the reduction in the redundant number of non-coherent integration.

Fifth, in order to reduce system complexity and hardware resources consumption, only signals whose PRN code rates are less than or equal to 2.046 Msps and sub-carrier frequency is 1.023 MHz can be acquired in this study. As an IF of the 4.092 MHz is adopted in this study, it ensures a low compatible IF data rate. Therefore, only a small-size IF data RAM and ISPMF are required.

Consequently, the proposed innovative FPGA-based MCC GNSS acquisition scheme is based on the above-mentioned five techniques, and therefore it has low complexity. According to experimental results, this acquisition scheme is capable of acquiring multi-constellations civil GNSS signals including GPS L1 C/A, GPS L1C, BDS B1I, BDS-3 B1C, Galileo E1, and GLONASS L1.

Moreover, as long as the signal meets the above-mentioned fifth requirement, it can be acquired by the proposed acquisition scheme with the corresponding PRN codes and parameter configurations.

What is more, the first four techniques are all without loss of generality, and therefore these techniques can be extended to acquire other GNSS signals like GPS L5, as long as the IF, IF data RAM and etc. meet the demand.

As a consequence, the proposed scheme has high practical value.

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**REFERENCES**

1. Mongredien, C., et al.: Opportunities and Challenges for Multi-Constellation, Multi-Frequency Automotive GNSS Receivers, pp. 157–170. Microelectronic Systems (2011)
2. Chebib, S., et al.: GNSS signals acquisition and tracking in unfavourable environment. Radioengineering. 27(2), 557–571 (2018)
3. Guo, C., Xu, B., Tian, Z.: Research on multi-constellation GNSS compatible acquisition strategy based on GPU high-performance operation. EURASIP J. Wirel. Commun. Netw. 1 (2018), 112 (2018)
4. Zahid, M., Ali, R., Afzal, M.: Acquisition and analysis of multi-global navigation satellite system signals based on fast Fourier transform. J. Commun. Technol. Electron. 64(11), 1288–1297 (2019)
5. Tran, V., Shvaramaiah, N., Dempster, A.: Feasibility analysis of baseband architectures for multi-GNSS receivers. GPS Solutions. 21(1), 1–11 (2017)
6. Kong, S: High sensitivity and fast acquisition signal processing techniques for GNSS receivers. IEEE Signal Proc. Mag. 34(5), 59–71 (2017)
7. Leckere, J., Bottoner, C., Farine, P.: Comparison framework of FPGA-based GNSS signals acquisition architectures. IEEE Trans. Aerosp. Electron. Syst. 49(3), 1497–1518 (2013)
8. Yang, Y., Ba, X., Chen, J.: A novel VLSI architecture for multi-constellation and multi-frequency GNSS acquisition engine, Q Control Trans. 7, 655–665 (2019)
9. Povey, G.J., Grant, P.M.: Simplified matched filter receiver designs for spread spectrum communications applications. Electron. Commun. Eng. J. 5(2), 59–64 (1993)
10. Spillard, C., Spangenberg, S.M., Povey, G.J.: A serial-parallel FFT correlator for PN code acquisition from LEO satellites. Int. Symp. Spread Spectr. Tech. Appl. 2(1), 446–448 (1998)
11. Ba, X.R., et al.: Design of a universal acquisition engine for GNSS receiver. In: Proc. CSNC, pp. 133–142 Academy Exchange Center of China Satellite Navigation Office, Nanjing, Jiangsu, China (2014)
12. Grant, P.M., et al.: Doppler estimation for fast acquisition in spread spectrum communication systems. In: Proc. Spread Technology to Africa, pp. 106–110.IEEE, Sun City, South Africa (1998)
13. Guo, W., et al.: A new FFT acquisition scheme based on partial matched filter in GNSS receivers for harsh environments. Aerosp. Sci. Technol. 61, 66–72 (2017)
14. Betz, J.W.: The offset carrier modulation for GPS modernization. In: Proc. ION NTM, pp. 639–648. Institute of Navigation, San Diego, CA (1999)
15. Lohan, E.S., et al.: Unambiguous techniques modernized GNSS signals: surveying the solutions. IEEEN Signal Proc. Mag. 34(5), 38–52 (2017)
16. Hao, F., et al.: Unambiguous acquisition/tracking technique based on sub-correlation functions for GNSS sine-BOC signals. Sensors. 20, 485–507 (2020)
17. Han, Q., et al.: BOC signal acquisition algorithm based on similar enfoldment. Int. J. Aerosp. Eng. 2020(4), 1–19 (2020)
18. Fishman, P.M., Betz, J.W.: Predicting performance of direct acquisition for the M-code signal. In: Proc. ION NTM, pp. 574–582. Institute of Navigation, Anaheim, CA (2000)
19. Ward, P.W., Liljo, W.E.: Ambiguity removal method for any GNSS binary offset carrier (BOC) modulation. In: Proc. ION ITM, pp. 406–419. Institute of Navigation, Anaheim, CA (2009)
20. Julien, O., et al.: ASPcCT Unambiguous sine-BOC(n, n) acquisition/tracking technique for navigation applications. IEEE Trans. Aerosp. Electron. Syst. 43(1), 150–162 (2006)
21. Yao, Z., Lu, M., Feng, Z.: Unambiguous sine-phased binary offset carrier modulated signal acquisition technique. IEEE Trans. Wireless Commun. 9(2), 577–580 (2010)
22. Yao, Z, et al.: Pseudo-correlation-function-based unambiguous tracking technique for sine-BOC signals. IEEE Trans. Aerosp. Electron. Syst. 46(4), 1782–1796 (2010)
23. Gao, Y., Yao, Z., Lu, M.: Design and implementation of a real-time software receiver for BDS-3 signals. J. Inst. Navig. 66(1), 83–97 (2019)
24. Borio, D.: A statistical theory for GNSS signal acquisition, Ph.D. Thesis. Politecnico Di Torino (2008)
25. Parkinson, B.W., Spilker, J.J.: Global positioning system: theory and applications. Prog. Astronaut. Rocketry. 163, 3–55 (1996)
26. Blunt, P.: Advanced global navigation satellite system receiver design, Ph.D. Thesis. University of Surrey (2007)
27. Macchigernot, F., Petovello, M.G., Lachapelle, G.: Combined acquisition and tracking methods for GPS L1 C/A and L1C signals. Int. J. Navig. Obser. 1–19 (2010)
28. Liu, Q., et al.: Mean acquisition time analysis for GNSS parallel and hybrid search strategies. GPS Sol. 23(4), 94–106 (2019)

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