COARSE SELF-SYNCHRONIZATION TECHNIQUE FOR GNSS RECEIVERS

Irène Masinjara Mahafeno, Laura Conde-Canencia, Emmanuel Boutillon

LESTER, Centre de recherche, BP 92 116, 56 321 Lorient Cedex, France
E-mail:{masinjara-irene.mahafeno, laura.conde-canencia, emmanuel.boutillon}@univ-ubs.fr

ABSTRACT

In this paper, an original low-complexity technique for fast PN code acquisition is proposed. Its major advantage is that the receiver does not need to know the PN code itself, but only its length. We have applied the proposed technique in the context of GNSS receivers, and more specifically to Galileo, the European contribution to GNSS. The analysis of its performances (mainly in realistic environments) shows that the proposed technique significantly accelerates the signal acquisition process since it greatly reduces the search space. In fact, the correlation process in the synchronization step is no longer carried out for the whole data sequence but within a window of small length, which can be about 100 times smaller than the code length. Moreover, its architecture has a low cost implementation.

***Index Terms*—** synchronization, signal acquisition, GNSS

1. INTRODUCTION

This study deals with code synchronization in Direct -Spread Code Division Multiple Access (DS-CDMA) systems. Code synchronization is in fact one of the most critical features in all spread spectrum systems and, in the case of Global Navigation Satellite Systems (GNSS) receivers, it is directly related to position accuracy and Time To First Fix[1] (TTFF). At reception, the crucial function of despreading the received PN code is accomplished by multiplying the incoming signal by a local replica of the PN code, *i.e.*, by correlation. To achieve a proper despreading the local and received PN codes must be perfectly aligned (synchronized).

Code acquisition, also known as coarse or initial synchronization, is the necessary first step of synchronization. The first alignment precision is typically in the order of half a chip. Acquisition time can attain several tens of minutes, depending on the code lengths, acquisition techniques, etc. However, in GNSS receivers, the time needed to acquire a satellite signal directly impacts on TTFF, hence calling for the development of fast acquisition techniques.

Traditional techniques involve correlating the replica code sequence and the received signal. Correlators can be divided in two main types: time-domain and frequency-domain.

The time-domain correlator performs a direct correlation of the replica code sequence and the received signal in the time domain. This implies that a dedicated processing step is carried out for each possible code phase and this results in prohibitely high TTFF.

Software-based receivers are the most promising and flexible ones, as they are highly configurable and can accommodate all possible modes of operation. For these, a second type of correlator relying on frequency-domain acquisition techniques is more suited. The Fast Fourier Transform enables parallel processing for all possible code delays and significantly reduces computational complexity. The basic idea is the calculation of the correlation function via the frequency-domain. Several algorithms have been proposed to optimise this frequency approach in terms of speed [2], reduced complexity [3] and Doppler effect compensation [4]. They constitute valuable issues as software receivers constitute the major trend in GNSS receiver technologies.

The technique we propose involves a serial search but not in a blind manner. In fact, the correlation process in the synchronization step is carried out within a window of small length, which can be about 100 times smaller than the code length.

This paper is organized as follows: An original architecture for fast code acquisition is proposed in Section 2. Simulation results considering noise and inter-satellite interference are given in Section 3. In Section 4, we evaluate the gain introduced by the proposed technique in comparison with the traditional serial acquisition. Finally, Section 5 draws conclusions and future work.

2. A NEW CODE ACQUISITION TECHNIQUE

In Galileo Signal-In-Space (SIS), a data channel is the result of modulating a PN code, a sub-carrier (if present), and a secondary code (if present) with a navigation data stream. In this study, we consider the E1-B data channel Binary Offset Carrier (BOC) (1,1) described in [1], where the chip rate is $R_{c,\text{E1-B}} = 1.023 \times 10^6$ chips/s. Detecting the beginning of the data sequence provides a first coarse alignment between the received sequence and the local replica of the PN code. This involves detecting the data transition in the data stream.

[1] The time before the receiver first computes a position solution when powered on.

2.1. The Coarse Self-Synchronization (CSS) technique

In this section, we describe the proposed algorithm, which easily determines the beginnings of the code-subcarrier sequences. At reception, the signal from one satellite can be expressed as follows:

$$r(k) = \alpha\, \nu(k) + w(k) \tag{1}$$

where

$$\nu(k) = \sum_{i=-\infty}^{+\infty} d_{[i]_N} c_{|i|_N} \times rect_N\,(k - i \times N) \times e^{j(\phi + i\theta)} \tag{2}$$

is the emitted signal with amplitude α and phase ϕ. N is the number of chips in a sequence (N_{chips}) multiplied by the number of samples per chip ($n_{samples}$), d_k is the k-th symbol of the navigation message, c_k is the k-th chip of the PN code, θ is the Doppler shift and $w(k)$ is an additive white Gaussian noise (AWGN) with zero mean and variance σ_w^2. Following the notation in [1], $[i]_N$ denotes the integer part of $\frac{i}{N}$ and $|i|_N$ denotes i modulo N. The function $rect_N(k)$ is equal to 1 for $k \in [0, N-1]$ and equal to 0 elsewhere. The proposed coarse alignment architecture scheme is presented in Fig. 1.

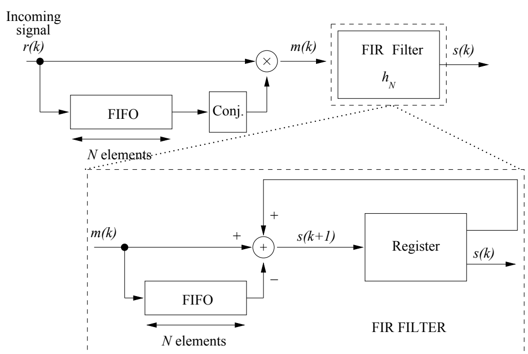

Fig. 1. Coarse alignment architecture scheme and implementation of the FIR filter

In the proposed algorithm, we start by computing the signal $m(k)$, given by the following equation:

$$m(k) = r(k) \times r^*(k - N) \tag{3}$$

where r^* denotes the complex conjugate of r. Considering (1), (2) and assuming $\alpha = 1, \sigma_w^2 = 0$ (no noise) and $c^2_{|i|_N} = 1$ (BPSK), equation (3) leads to:

$$\begin{aligned} m(k) &= d_{[k]_N} d^*_{[k-N]_N} e^{-jN\theta} \\ &= d_{[k]_N} d^*_{[k]_N - 1} e^{-jN\theta} \end{aligned} \tag{4}$$

The signal $m(k)$ is then sent through a FIR filter $h_N = \sum_{i=0}^{N-1} z^{-i}$ to generate $s(k)$:

$$s(k) = \sum_{i=0}^{N-1} d_{[k-i]_N} \times d^*_{[k-i]_N - 1} \times e^{-jN\theta} \tag{5}$$

Take $k = aN + b$, with $a, b \in \mathcal{N}$. Clearly:

$$\begin{cases} [k-i]_N = a & \text{if} \quad i \le b \\ [k-i]_N = a - 1 & \text{if} \quad i > b \end{cases}$$

The coarse alignment signal $s(k)$ can then be written as:

$$\begin{aligned} s(k) &= \sum_{i=0}^{b} d_{[k-i]_N} d^*_{[k-i]_N - 1} e^{-jN\theta} \\ &+ \sum_{i=b+1}^{N-1} d_{[k-i]_N} d^*_{[k-i]_N - 1} e^{-jN\theta} \\ &= (b\, d_a\, d_{a-1} + (N-b) d_{a-1} d_{a-2})\, e^{-jN\theta} \end{aligned} \tag{6}$$

Fig. 2 presents the output of the CSS architecture for E1-B data channel when $\theta = \phi = 0$ and $\sigma_w^2 = 0$ (no noise). We have considered a 10-bit ($N_{bits} = 10$) random information message and 4 samples per chip ($n_{samples} = 4$). The code length is $N_{chips} = 4092$ chips.

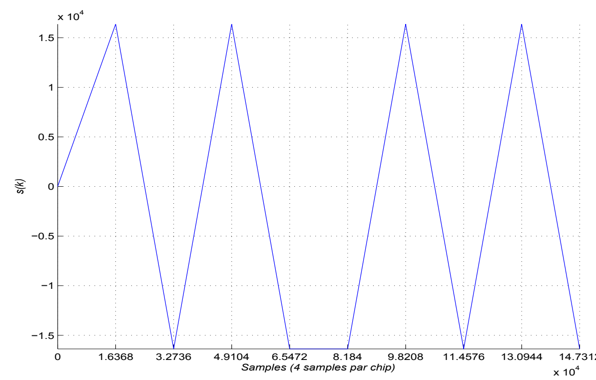

Fig. 2. Example of coarse aligment signal $s(k)$ (10-bit random information message, 4092-chip code, 4 samples per chip), when $\theta = \phi = 0$ and $\sigma_w^2 = 0$

This figure shows that a change in the slope of the coarse alignment signal corresponds to a data transition and thus to a beginning of the PN code sequence (*i.e.* the synchronization instants). Each transition instant corresponds indeed to a multiple of $N = N_{chips} \times n_{samples}$.
The FIR filter is a simple sliding window accumulator and can be easily implemented using:

$$s(k) = s(k-1) - m(k-N) + m(k) \tag{7}$$

where $s(k-1) = \sum_{i=0}^{N-1} m(k-1-i)$ (see Fig. 1).

The proposed architecture can be seen as performing an auto-correlation through a sliding window of size N, without the need of a local replica of the PN code. For this reason, the technique is called *Coarse Self-Synchronization*.

2.2. The accumulation method

Let us consider a single-satellite system, whose signal is only affected by a delay D. We set $\theta = \phi = 0$ in (2) and $\sigma_w^2 = 0$. Fig. 3 represents the coarse alignment signal $s(k)$ (full line) and its absolute value $|s(k)|$ (dashed line), for $D = 1000$ chips. The changes in the slope of $s(k)$ are located at $(i \times N_{chips} \times n_{samples}) + D$, where $i = 0, 1, \ldots, (N_{bits} - 1)$. Each of these instants corresponds to a beginning of a code sequence.

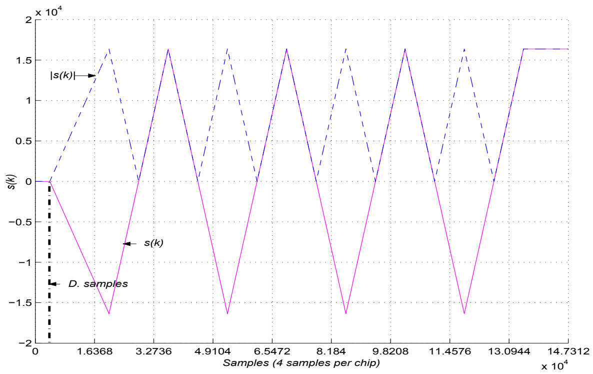

Fig. 3. Coarse aligment signal $s(k)$ and its absolute value $|s(k)|$ (10-bit information message, 4092-chip code, 4 samples per chip), when $\theta = \phi = 0$, the signal delay $D = 1000$ chips and $\sigma_w^2 = 0$

In order to identify the signal delay corresponding to a particular satellite, a simple method that detects changes in the slope of $s(k)$ is proposed. This method named *accumulation method* involves summing the function $|s(k)|$ over N_{bits} consecutive bit periods (*i.e.* every N samples) to obtain:

$$s_{add}(k) = \sum_{q=1}^{N_{bits}} |s(k - qN)| \tag{8}$$

In fact, according to equation (6), the function $|s(k)|$ in the interval $[aN - D, (a-1)N - D]$ is equal to: $|s(k)| = |d_{a-2}d_{a-1}(N - |k|_N) + d_{a-1}d_a(|k|_N)|$ if $[k]_N < a$ and $|s(k)| = |d_{a-1}d_a(N - |k|_N) + d_a d_{a+1}(|k|_N)|$ if $[k]_N \geq a$. Thus, on an interval of length N, $|s(k)|$ can only have 4 different shapes, as shown in Fig. 4.

The function $s_{add}(k)$ is then the accumulation of N_{bits} curves of type **a)**, **b)**, **c)** or **d)**. Since the bit d_a is random, each curve has the same probability $p = 0.25$ to appear. Thus, in average, the function will have a maximum equal to $N_{bits}N$ for $k' = D$, and a minimum equal to $N_{bits}N/2$ for $k' = D + N/2$. In a non-noise environment, the index Δ of the maximum of $s_{add}(k)$ is then equal to the delay D. In practice, in a noisy environment, the estimated delay Δ will be given by the index of the maximum value of $s_{add}(k)$, *i.e.*:

$$\Delta = \arg\max\{s_{add}(k); k = 1..N\} \tag{9}$$

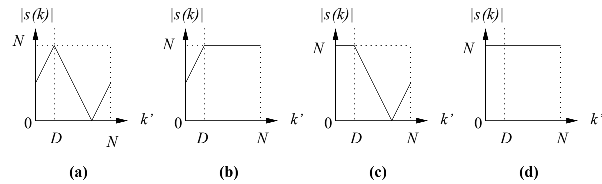

Fig. 4. Shape of the curve $|s(k)|$ in the interval $[aN - D, (a-1)N - D]$, with $k' = k - aN + D$: **a)**$-d_{a-2}d_{a-1} = d_{a-1}d_a = -d_a d_{a+1}$; **b)**$-d_{a-2}d_{a-1} = d_a d_{a+1} = d_{a-1}d_a$; **c)**$d_{a-2}d_{a-1} = d_a d_{a+1} = -d_{a-1}d_a$ **d)**$d_{a-2}d_{a-1} = d_a d_{a+1} = d_{a-1}d_a$

2.3. Complexity of the proposed technique

The CSS technique combined with the accumulation technique has a very low hardware complexity. The operations and the memories required to perform this algorithm are: 2 additions (ADD), 4 multiplications (MULT) and $2 \times N$ memories (MEM) for equation (3); 2 ADD and $2 \times N$ MEM for equation (7); 2 ADD and N MEM for equation (8); $\frac{1}{N_{bits}}$ ADD for equation (9). The number of ADD and MULT are considered for real variables (*i.e.* a complex ADD requires 2 real ADD and a complex MULT requires 2 real ADD and 4 real MULT). The very low complexity of this algorithm (7 ADD and 4 MULT per sample, plus 5 FIFO of size N) makes it very useful for software techniques or very low complexity hardware receiver.

3. SIMULATION RESULTS CONSIDERING NOISE AND INTER-SATELLITE INTERFERENCE

In this section, we first present simulation results when the system is affected by AWGN. Then, we present a multiple-satellite scheme and a method to mitigate the inter-satellite interference.

3.1. Gaussian noise impact

Let us now consider that the single-satellite system is affected by an AWGN noise, a Doppler shift and a phase shift. The complex received signal can be written as follows:

$$r(k) = \alpha\, \nu(k - D) + w(k) \tag{10}$$

The noise is defined in terms of carrier-to-noise ratio C/N_0, expressed in dB-Hz.

Fig. 5 presents the signal $s_{add}(k)$ for a typical value of $C/N_0 =$ 45 dB-Hz and a fixed delay $D = 1000$ chips. θ and ϕ are generated randomly. A receiver filter with a 4.092 MHz bilateral bandwidth is used. In this example, only the real part of $s(k)$ is considered when computing $s_{add}(k)$. The estimated delay is $\Delta = 1027$ chips, so the estimation error is $\delta = |D - \Delta| = 27$ chips.

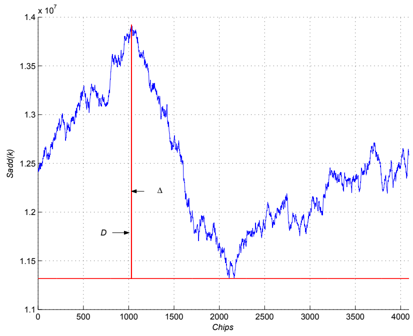

Fig. 5. Signal after the accumulation technique for 1 satellite (100-bit information message, 4092-chip code and 4 samples per chip, $C/N_0 = 45$ dB-Hz, $D = 1000$ chips, $\phi \neq 0, \theta \neq 0$)

To perfectly determine D, the idea is to apply the classical correlation search between the received signal and the code replica, starting from the value of given Δ by the CSS algorithm in a *zigzag* mode (*i.e.*, synchronization for a delay of Δ, then $\Delta + 1$, then $\Delta - 1$, then $\Delta + 2$ and so on). The search process is then stopped when the correct synchronization is found. In this example, we have $\delta = |D - \Delta| = 27$. Thus, using the *zigzag* scan, around 54 chip values will be tested before obtaining the exact synchronization.
So, with this technique, the time for code synchronization is a function of the time of the coarse alignment, plus the time of the correlation process using a *zigzag* scan starting from Δ.

3.2. Application to a multiple satellite incoming signal

We now consider a system with a multiple-satellite incoming signal affected by an AWGN noise. The objective is to compute position, time and speed when signals from at least four satellites are available. Fig. 6 presents the schematics of the problem.

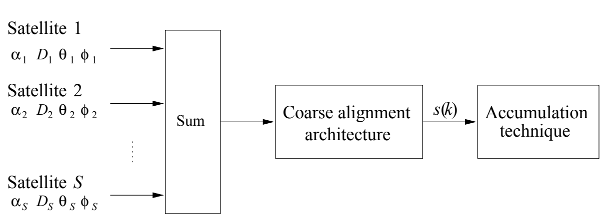

Fig. 6. Context of the study and architectural overview of the detector

The global received signal corresponds to the superposition of delayed copies of the multiple (four or more) transmitted signals. Each signal from each satellite is affected by a delay D_s, a Doppler shift θ_s and a phase shift ϕ_s. A simplified model of the global received signal at the input of the architecture can be written as:

$$r(k) = \sum_{s=1}^{S} \alpha_s \times \nu_s(k - D_s) + w(k) \qquad (11)$$

where S is the number of satellites in sight, $\nu_s(k)$ is the signal emitted from satellite s with an amplitude α_s. A Doppler shift of $\theta_s \neq 0$ leads to a rotation of the signal $s(k)$ by an angle $-N\theta_s$, as shown in equation (6). This rotation can be easily compensated by multiplying $s(k)$ by $e^{j\psi}$, with $\psi = N\theta_s$:

$$s_\psi(k) = Re\left[s(k) \times e^{j\psi}\right] \qquad (12)$$

Since $N\theta_s$ is unknown, several values of ψ are systematically tested; typically $\psi \in \{0, \frac{\pi}{4}, \frac{\pi}{2}, \frac{3\pi}{4}\}$. The best projection is then selected.

Simulation results are obtained for $S = 4$ satellites with same C/N_0, which is the worst case. In fact, if one satellite has a significant higher C/N_0 than the others, its contribution will dominate the CSS output. This domination leads to a much more precise delay estimation. Fig. 7 shows the signal $s^0_{add}(k)$ computed using $\psi = 0$, *i.e.* $s_0(k) = Re\left[s(k)\right]$. Different delays, phase shifts and Doppler shifts are considered.

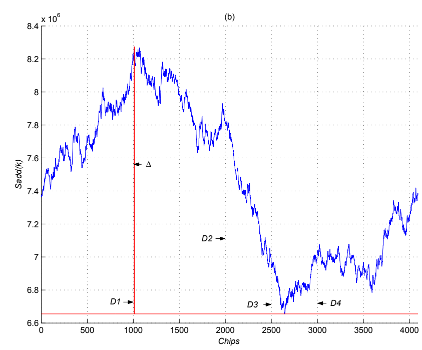

Fig. 7. Signal $S_{add}(k)$ for $S = 4$ satellites (150-bit information message, 4092-chip code and 4 samples per chip, $C/N_0 = 45$ dB-Hz, $D_1 = 1000$, $D_2 = 2500$, $D_3 = 2000$, $D_4 = 3000$, $\theta \neq 0, \phi \neq 0, \psi = 0$)

In this figure, the different delays are also plotted. We set: $D_1 = 1000$, $D_2 = 2500$, $D_3 = 2000$ and $D_4 = 3000$. As for the phase shifts and the Doppler shifts, they are generated randomly. With this example, the estimated delay is $\Delta = 1010$ chips. It corresponds to a good estimation of the delay of the satellite number 1.

3.3. Projecting the coarse alignment signal to minimize inter-satellite interference

Since each Doppler implies a different rotation of the CSS signal, we propose to perform systematically the projection of $s(k)$ on the four axes defined by the angles $\psi = \frac{a\pi}{4}$, $a = 0..3$.

Four signals $S_{add}^{a}(k)$ are then generated using the accumulation method. On each axis, the index of the maximum of $Re\left[S_{add}^{a}(k)\right]$ gives a potential candidate Δ_a.
Using again the example of the previous section, Fig. 8 corresponds to the projection of $s(k)$ on the axis $\psi_1 = \frac{\pi}{4}$. The corresponding estimated delay is $\Delta_1 = 2539$ chips, which is close to the delay of satellite 3 ($D_3 = 2500$).

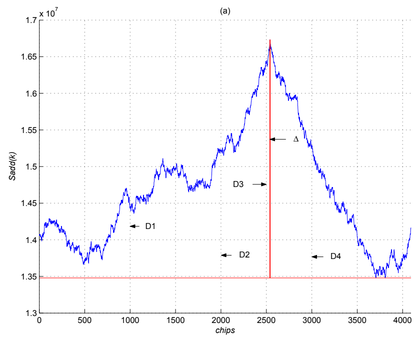

Fig. 8. Signal $S_{add}(k)$ for $S = 4$ satellites (150-bit information message, 4092-chip code and 4 samples per chip, $C/N_0 = 45$ dB-Hz, $D_1 = 1000$, $D_2 = 2500$, $D_3 = 2000$, $D_4 = 3000$, $\theta \neq 0$, $\phi \neq 0$, $\psi = \frac{\pi}{4}$

So, applying a projection over differents axes (different values of ψ) helps to determinate the different delays D_s. However, a single estimation Δ of a delay D_s is required to initiate a *zigzag* coarse synchronization search. The problem is then how to choose the best Δ_a ($a = 1..Q$), where Q is the number of projections. In our simulations, we apply a method that involves selecting one Δ_a which gives the smallest value of $|\Delta_a - \Delta_b|$ when $a \neq b$. By the following, this method is called *selection of the vote of majority*. Fig. 9 gives the schematic of the detector used in the sequels.

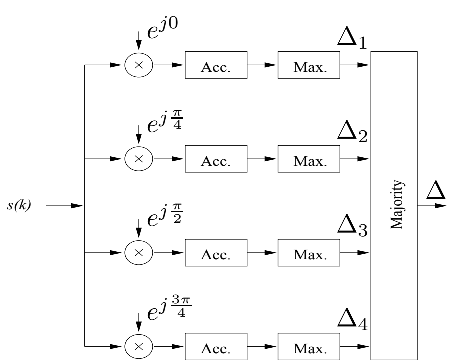

Fig. 9. Schematic representation of the detector

The next step of our study should permit to associate each candidate delay with each satellite signal. This task could be included in the fine synchronization step. It can be carried out using an iterative process, *i.e.*, by iteratively separating the different signals.

4. EVALUATION OF THE GAIN INTRODUCED BY THE CSS TECHNIQUE

In order to evaluate the gain introduced by the proposed CSS technique, compared to the traditional serial acquisition, we calculate the probability density function (*pdf*) of the minimal distance D_{min} between the estimated delay Δ and the exact delay $\{D_s; s = 1..S\}$. This distance is defined as:

$$D_{min} = \min_{s=1..S}\{|\Delta - D_s|_{N_{chips}}, N_{chips} - |\Delta - D_s|\} \quad (13)$$

We can schematize the problem as in Fig. 10 for a four-satellite system, where the correlation process is represented by a torus of size N_{chips}. The 4 fixed points denoted $\{D_s; s = 1..4\}$ represent the delay of each satellite signal.

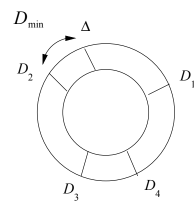

Fig. 10. Representation of the correlation process by a torus

One the one hand, for the traditional serial acquisition, the starting point Δ is a random value between 1 and N_{chips}, since the correlation process is done in a blind manner. On the other hand, the correlation process is started at a known instant Δ (Δ is the estimated signal delay) when using the proposed CSS technique, as already mentioned in sub-section 3.1.
Based on this approach, we compare the *pdf* of D_{min} of the CSS technique with that of the traditional serial acquisition. A Monte-Carlo estimation of the *pdf* of D_{min}, obtained with the CSS algorithm is presented in Fig. 11 for the following conditions: $S = 4$; $C/N_0 = 45$ dB-Hz; $N_{chips} = 4092$, $n_{samples} = 4$; delays, phase shifts and Doppler shifts generated randomly; $N_{bits} = 150$; a number of draws equal to 50,000 blocks of N_{bits} bits. We apply 4 projections of $s(k)$ using the following values of: $\psi_a = \{0, \frac{\pi}{4}, \frac{\pi}{2}, \frac{3\pi}{4}\}$. The selection by *the vote of majority* has been used to estimate Δ among the 4 projections. As a reference, we also plot in Fig. 11: the *pdf* of D_{min} if a *genius* helps us to always select the best estimation among Δ_1, Δ_2, Δ_3 and Δ_4 (optimal minimal distance) and the *pdf* of D_{min} if we simply draw a random point (traditional serial method).
This figure shows that, for the proposed technique, the mean values of D_{min} are significantly smaller than those of the traditional serial method. One should note that the majority

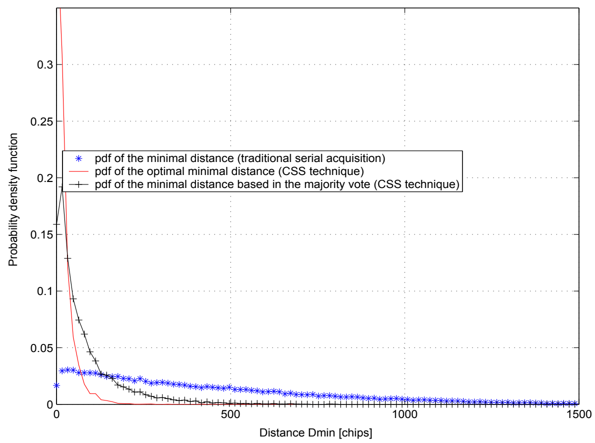

Fig. 11. The probability density function of the minimal distance between the fixed points D_s and the correlation process starting point Δ (150-bit information message for the CSS technique)

vote technique is not optimal. In fact, the genius aided selection method presents significantly better performance. Further study needs to be done in order to improve the performance.

Other simulations show that better values of D_{min} are obtained for the CSS technique when the number of information bits is increased or/and for higher values of C/N_0. As an example, the first statistical orders (*i.e.* the means $E[D_{min}]$) of the *pdf* for the CSS technique are plotted in Fig. 12 when $N_{bits} = \{10, 25, 50, 100, 150, 200\}$ and $C/N_0 = \{40, 45, 50, 55, 60\}$ dB-Hz.

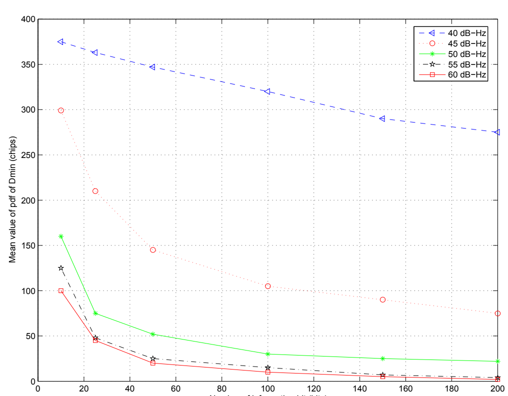

Fig. 12. The mean value of *pdf* of D_{min} for different values of N_{bits} and C/N_0 for the CSS technique

The mean value of the *pdf* of D_{min} decreases as a function of N_{bits} or/and C/N_0. Then, the proposed technique achieves better performance for large values of N_{bits} or/and C/N_0. Note that with the traditional serial method D_{min} is always about 500 chips for any value of N_{bits} and C/N_0 (with the same conditions as in Fig. 12).

5. CONCLUSIONS

In this paper, an original technique for fast low-complexity code acquisition has been proposed and preliminary results have been presented. This study assesses the principles and the possible benefits offered by the proposed CSS architecture. One of its major advantages is that only the length of the code needs to be known and not the code itself. Indeed, it does not rely on any special property of the spreading codes and can thus be applied to any set of PN sequences. Its applicability to the Galileo system has been discussed. Simulation results have shown that the proposed technique accelerates the acquisition process since it reduces the search space. Moreover, it can be easily implemented. Further work on the subject should be the evaluation of the benefits of the proposed technique in a more realistic environment, using namely a noisy multipath channel and the Galileo spreading codes. Finally, this study has been carried out in the context of GNSS receivers but the proposed aquisition technique can be applied to other contexts.

6. ACKNOWLEDGEMENTS

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7. REFERENCES

[1] European Space Agency / Galileo Joint Undertaking, “Galileo Open Service Signal In Space Interface Control Document (OS SIS ICD)”, 2006.

[2] D.J.R. van Nee, A.J.R.M. Coenen, “New fast GPS code-acquisition technique using FFT”, Electronics Letters, Vol. 27, No.2, 17th January 1991.

[3] P.K. Sagijaru, S. Again, D. Akopian “Reduced complexity acquisition of GPS signals for software embedded applications”, IEE Proc. Radar Sonar Navig., Vol. 153, No. 1, Feb. 2006

[4] R. A. Stirling-Gallacher, A. P. Hulbert, G. J. R. Povey, “A Fast Acquisition Technique for a Direct Sequence Spread Spectrum Signal in the Presence of a Large Doppler Shift”, Proc. of IEEE 4th Int. Symposium on Spread Spectrum Techniques and Applications, Sept. 1996.