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ON SYNCHRONIZATION OF DOPPLER-STRETCHED GPS SIGNALS

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ABSTRACT

The problem of synchronization of Doppler-stretched GPS signals is addressed. A new synchronization technique for the course acquisition code is proposed, which does not assume that the so called narrow-band condition is satisfied. Thus, the Doppler effect is modeled as a frequency shift on the carrier and a stretching of the complex envelope. As a consequence, observation intervals significantly longer than those adopted in conventional techniques can be adopted leading to a higher immunity against disturbance signals. The proposed technique exploits the cyclostationarity of the received signal and the fact that transmitted and received signals are jointly spectrally correlated. Simulation results show the better performance of the proposed method with respect to the conventional technique that models the Doppler effect just as a frequency shift of the carrier.

***Index Terms*—** Spectrally correlated processes, cyclostationarity, Doppler effect, synchronization, GPS signal

1. INTRODUCTION

The first step in the acquisition of the global position system (GPS) signal requires a search of the periodical course acquisition (C/A) code that identifies the satellite [6]. At this step, Doppler effect and phase mismatch (delay) of the received C/A code and the replica stored in the receiver must be compensated. For each candidate C/A code, the receiver correlates the local replica with a delayed and frequency shifted version of the incoming signal. That is, it computes the narrow-band cross-ambiguity function (NB-CAF) [15, Sect. 10.1] of the received signal and the local replica for delay and frequency shift ranging in predetermined intervals and with predetermined steps [6, Sect. 5.8].

In the presence of constant relative radial speed between transmitter and receiver within the observation interval, the Doppler effect introduces a frequency shift in the carrier of the received signal with respect to that of the transmitted one and also introduces a time-scale factor (a time stretch) in the argument of the complex envelope of the received signal [8, Sect. 7.3.3], [15, pp. 339-340].

In the described acquisition procedure of the GPS signal, the Doppler effect is modeled just as a frequency shift of the carrier and the time-scale factor in the argument of the complex envelope is assumed to be unity. This widely adopted model for the Doppler effect is valid when the so called "narrow-band condition" is satisfied, that is, when the product of transmitted-signal bandwidth and data-record length is much smaller than the ratio of the medium propagation speed and the relative radial speed between transmitter and receiver [8, Sect. 7.5.1], [15, pp. 339-340]. Under the narrow-band condition and under the assumption of constant relative radial speed between transmitter and receiver, locating the peak of the magnitude of the NB-CAF provides maximum likelihood (ML) estimates of delay and frequency shift in additive white Gaussian noise (AWGN) at the high signal-to-noise ratio (SNR) regime [15, Sects. 9.2, 10.1]. The drawback of this technique is that it is very time consuming.

If the transmitted-signal bandwidth and a maximum value for the relative radial speed are fixed, the narrow-band condition puts an upper bound to the maximum data-record length that can be adopted in order to effectively model the Doppler effect as a simple carrier frequency shift. For the GPS-L1 signal, in order to satisfy the narrow-band condition, the first step of the C/A code synchronization is made on a single period of the C/A code, that is, over an observation interval equal to 1 ms [1, Chap. 5, pp. 69-72], [10], [14].

The limit on the data-record length puts a limit to the minimum SNR for which satisfactory performance of the synchronization algorithm can be achieved. In order to properly operate at very low SNR, as in indoor scenarios, the data-record length needs to be significantly augmented with respect to 1 ms [13]. In such a case, however, the narrow-band condition is not satisfied, and the non-unit time-scale factor in the argument of the complex envelope of the received signal must be accounted for [8, Sect. 7.3.3], [11], [13]. That is, the received signal is a stretched version of the transmitted one. Consequently, the estimation procedure based on the NB-CAF performs poorly since of the non accurate model for the Doppler effect.

In this paper, a new synchronization technique for L1-GPS signal is proposed, which does not assume the narrow-band condition be satisfied. Thus, it is suitable to be exploited with very large data-record lengths in severe disturbance environments. The received signal is modeled as an attenuated, time-scaled, delayed, and frequency-shifted replica of the transmitted one. The proposed method exploits the cyclostationarity property of the transmitted L1-GPS signal [9] to estimate the time-scale factor and the frequency shift of the received signal. It is shown that the received signal is still cyclostationary but with cyclic features which are different from those of the transmitted signal. Moreover, transmitted and received signals are shown to be not jointly cyclostationary but, rather, jointly spectrally correlated. (Jointly) spectrally correlated processes have a Loève bifrequency (cross-)spectrum whose spectral masses are concentrated on a countable set of support curves in the bifrequency plane [8, Chap. 4]. They include the (jointly) almost-cyclostationary and cyclostationary processes as special cases when the support curves are lines with unit slope. A linear time-variant (LTV) equalization of the received signal is performed to compensate the time-scale factor and the frequency shift. Then, the delay and phase shift introduced by the Doppler propagation channel are estimated starting from the equalized signal and the local replica of the C/A code stored in the receiver.

Simulation experiments are carried out to show the effectiveness of the proposed method to provide satisfactory performance in severe noise and interference environment where the 1 ms data-record length is not sufficient to obtain reliable estimates by the NB-CAF and, hence, larger data record-lengths need to be adopted. In the large data-record regime, the proposed method is shown to outperform the technique based on the NB-CAF and able to properly counteract the presence of noise and interference due to the signal selectivity properties of cyclostationarity.

The paper is organized as follows. In Section 2, the GPS-L1 signal model is briefly reviewed and its second-order cyclostationarity properties presented. The model for the received signal is described in Section 3 and the proposed estimation method presented in Section 4. Numerical results are shown in Section 5 and conclusions are drawn in Section 6.

2. GPS-L1 SIGNAL

The continuous-time GPS-L1 signal [5, Par. 3.2-3.3, pp. 3-17], [6, Chap. 4, pp. 113-116] is a quadrature phase-shift keying (QPSK) signal

$$\begin{aligned} x(t) &= \sqrt{2}\, A\, d(t)\, c(t)\, \cos(2\pi f_{\mathrm{L1}} t + \phi_0) \\ &\quad + A\, d(t)\, p(t)\, \sin(2\pi f_{\mathrm{L1}} t + \phi_0) \end{aligned} \tag{1}$$

where $f_{\mathrm{L1}} = 1575.42$ MHz. In (1):

1) $d(t)$ is the *navigation message*. It is obtained by interleaving two periodic components and a binary pulse-amplitude-modulated (PAM) signal with bit period T_{b}; the PAM signal is multiplied by a periodic signal [5, Par. 20.3, pp. 68-122].
2) $c(t)$ is the *course acquisition (C/A) code* signal. It is the periodic replication with period $T_{\mathrm{CA}} = N_{\mathrm{c}} T_{\mathrm{c}}$ of a fixed Gold sequence with chip period T_c and $N_{\mathrm{c}} = 1023$ chips that identifies the satellite [5, Par. 3.3, pp. 26-30]. It results $T_c = 0.9775\mu$s, $T_{\mathrm{CA}} = 1$ ms, and $T_{\mathrm{b}} = 20 N_{\mathrm{c}} T_{\mathrm{c}}$.
3) $p(t)$ is the *precision P(Y) code* signal. It can be modeled as a binary PAM signal with i.i.d. symbols and bit period $T_{\mathrm{p}} = T_{\mathrm{c}}/10$ within realistic observation intervals [5, Par. 3.3, pp. 18-25].

Due to the presence in $x(t)$ of periodic replication operations and PAM signals, in [9], it is shown that the complex signal associated to $x(t)$

$$z(t) = \sqrt{2}\, A\, d(t)\, c(t) + j\, A\, d(t)\, p(t) \tag{2}$$

is second-order wide-sense cyclostationary. That is, its expected value and autocorrelation function are periodic functions of time. The analytical expression of (conjugate) cyclic autocorrelation functions and (conjugate) cyclic spectra of $z(t)$ are derived in [9]. These expression are very complicate due to the complex structure of the GPS-L1 signal (1). In [9], it is shown that different signal models should be considered depending on the length T of the observation interval. Consequently, different cyclic features are evidenced depending on the value of T.

1) Let $T = T_{\mathrm{CA}} = 1$ ms. Assuming that $d(t) = 1$ and no data-bit transition is present within the observation interval, the signal model for $z(t)$ is

$$z(t) = \sqrt{2} A\, c(t) - jA\, p(t) \tag{3}$$

where the C/A code signal $c(t)$ must be modeled as a binary PAM signal with bit period T_c. Thus, $c(t)$ exhibits cyclostationarity at cycle frequencies $\alpha = k/T_c$, k integer, which are pure second-order cycle frequencies [3]. The signal $p(t)$ is a PAM signal with period $T_p = T_c/10$. It has cycle frequencies $\alpha = k/T_p$, k integer, which are a subset of the cycle frequencies $\alpha = k/T_c$.

2) Let be $T = 20\, T_{\mathrm{CA}} = T_b = 20$ ms. If the observation interval is such that no data-bit transition is present, then the model for $z(t)$ is that in (3), where, unlike the case $T = T_{\mathrm{CA}}$, the signal $c(t)$ must be modeled as the periodic replication with period T_{CA} of a deterministic signal. The signal exhibit cyclostationarity with cycle frequencies $\alpha = k/T_c = kN_c/T_{\mathrm{CA}}$ which, in such a case, are impure second-order cycle frequencies [3].

Both autocorrelation function and conjugate autocorrelation function are necessary for a complete second-order characterization in the wide sense of complex-valued signals [12]. In the following, notation $(*)$ will be adopted for an optional complex conjugation in order to consider, in the same formula, both cyclic statistics and conjugate cyclic statistics.

3. RECEIVED SIGNAL

In the case of free-space propagation, wide-band transmitting and receiving antennas, and if the relative radial speed v_r between satellite and GPS receiver can be assumed constant within the observation interval, the complex envelope of the (noise-free) received signal is given by [8, Sect. 7.3.3], [11], [13]

$$y(t) = a\, z(st - d_0)\, e^{j2\pi\nu t} \tag{4}$$

where, in the case of stationary GPS receiver, $s = c/(c + v_r)$ is the time-scale factor, $\nu = (s-1) f_{\mathrm{L1}}$ is the frequency shift, d_0 is the time delay, and $a = |a| e^{j\phi_a}$ is a complex gain that accounts for propagation losses, phase ϕ_0, and phase introduced by the propagation channel. $c \simeq 3 \cdot 10^8$ m s^{-1} is the medium propagation speed. Note that the parameters ν and s are linked only if perturbation ionospheric effects and receiver oscillator instabilities are neglected [11].

For typical values of the elevation and Earth central angles, it can be shown that $|1-s| = |v_r/(v_r+c)| \simeq 2.5 \cdot 10^{-6}$.

Assuming for the bandpass GPS-L1 signal an approximate bandwidth $B \simeq 1/T_p = 10.23$ MHz, where T_p is the width of the narrowest rectangular pulse in the signal model, we have that $T = 1$ ms $\Rightarrow$ $BT \simeq 10.23 \cdot 10^3$ and $T = 10$ ms $\Rightarrow$ $BT \simeq 1.02 \cdot 10^5$. Thus, the so called "narrow-band condition" [8, Sect. 7.5]

$$BT \ll 1/|1-s| \simeq c/|v_r| \tag{5}$$

is practically satisfied for $T = 1$ ms and is not satisfied for $T = 10$ ms. When (5) is satisfied, then $s = 1$ can be assumed in the argument of the complex envelope $z(\cdot)$ in (4) leading to the classical model for the Doppler effect.

From (4) it follows that if $z(t)$ exhibits cyclostationarity with cycle frequency α, then $y(t)$ exhibits cyclostationarity with cycle frequency $s\alpha$ and the cyclic spectra of $y(t)$ and $z(t)$ are linked by

$$S^{s\alpha}_{yy^*}(f) = |a|^2 e^{-j2\pi\alpha d_0}\, \frac{1}{|s|} S^{\alpha}_{zz^*}\left(\frac{f-\nu}{s}\right). \tag{6}$$

Moreover, if $z(t)$ exhibits conjugate cyclostationarity with conjugate cycle frequency β, then $y(t)$ exhibits conjugate cyclostationarity with conjugate cycle frequency $s\beta + 2\nu$ and the conjugate cyclic spectra of $y(t)$ and $z(t)$ are linked by

$$S^{s\beta+2\nu}_{yy}(f) = a^2\, e^{-j2\pi\beta d_0}\, \frac{1}{|s|} S^{\beta}_{zz}\left(\frac{f-\nu}{s}\right). \tag{7}$$

Due to the presence of the non unit time-scale factor s in (4), it follows that $y(t)$ and $z(t)$ are not jointly almost-cyclostationary but, rather, jointly spectrally correlated [8, Chap. 4]. The Loève bifrequency cross-spectrum of $y(t)$ and $z(t)$ is given by

$$\mathrm{E}\left\{Y(f_1)\, Z^{(*)}(f_2)\right\} = \frac{a}{|s|}\, e^{-j2\pi(f_1-\nu)d_0/s} \sum_{\alpha_n \in A_{zz^{(*)}}} S^{\alpha_n}_{zz^{(*)}}\left(\frac{f_1-\nu}{s}\right) \delta\left(f_2 - {\scriptstyle(-)}\left(\alpha_n - \frac{f_1-\nu}{s}\right)\right) \tag{8}$$

where ${\scriptstyle(-)}$ is an optional minus sign linked to $(*)$. That is, it is constituted by spectral masses concentrated on a countable set of lines with non unit slope. In (8), $A_{zz^{(*)}}$ is the set of (conjugate) cycle frequencies of z, and $Y(f)$ and $Z(f)$ are the Fourier transforms of $y(t)$ and $z(t)$, respectively, defined in a distributional sense [4, Chap. 6].

4. PARAMETER ESTIMATION

Let

$$r(t) = y(t) + n(t) \tag{9}$$

be the noisy received signal, where $n(t)$ represents a zero-mean possibly non stationary disturbance signal. Assuming $y(t)$ and $n(t)$ statistically independent, the (conjugate) cyclic spectrum of $r(t)$ is given by

$$S^{\gamma}_{rr^{(*)}}(f) = S^{\gamma}_{yy^{(*)}}(f) + S^{\gamma}_{nn^{(*)}}(f)\,. \tag{10}$$

According to (8), the signals $z(t)$ and $y(t)$ are jointly spectrally correlated. Their Loève bifrequency cross-spectrum has spectral masses concentrated on a countable set of lines with slope $-{\scriptstyle(-)}1/s$. Since the parameter s is unknown, the location of the spectral lines is unknown and the corresponding spectral correlation densities can be estimated only with some uncertainty [7], [8, Sect. 4.5].

In this section, an interference-tolerant technique for estimating the parameters s, ν, d_0, and a is proposed, which is based on the cross-correlation between a replica of $z(t)$ stored in the receiver and a time-stretched version of $r(t)$.

Let α_0 and β_0 be a cycle frequency and a conjugate cycle frequency, respectively, of $z(t)$. The parameters s and ν can be estimated starting from estimates of the (conjugate) cycle frequencies of the received signal. Let

$$\lambda_{rr^{(*)}}(\alpha) \triangleq \int_{\mathbb{R}} \left|\widehat{S}^{\alpha}_{rr^{(*)}}(f)\right|^2 \mathrm{d}f \tag{11}$$

where $\widehat{S}^{\alpha}_{rr^{(*)}}(f)$ denotes the (conjugate) frequency-smoothed cyclic periodogram obtained observing signals in $[0, T]$.

Let s and ν be such that for some $\Delta\alpha$ and $\Delta\beta$ there is only one cycle frequency of $z(t)$ in the set $J(\alpha_0, \Delta\alpha) \triangleq [\alpha_0 - \Delta\alpha/2, \alpha_0 + \Delta\alpha/2]$ and only one conjugate cycle frequency of $z(t)$ in the set $J(\beta_0, \Delta\beta)$, and, moreover $S^{\alpha}_{nn^*}(f) = 0$ for $\alpha \in J(\alpha_0, \Delta\alpha)$ and $S^{\beta}_{nn}(f) = 0$ for $\beta \in J(\beta_0, \Delta\beta)$. In the ideal case of perfect measurements (that is, for infinite data-record length and infinitely small spectral resolution) we have $\widehat{S}^{\alpha}_{rr^*}(f) = S^{\alpha}_{rr^*}(f)$ and $\widehat{S}^{\beta}_{rr}(f) = S^{\beta}_{rr}(f)$

with probability 1 (w.p.1)). Moreover, when $\alpha \in J(\alpha_0, \Delta\alpha)$ and $\beta \in J(\beta_0, \Delta\beta)$, accounting for (6) and (7), the statistics $\lambda_{rr^*}(\alpha)$ and $\lambda_{rr}(\beta)$ are different from zero w.p.1 only for $\alpha = s\alpha_0$ and $\beta = s\beta_0 + 2\nu$, respectively. Thus, in the case of finite data-record length and spectral resolution, the following estimates for the cycle frequency α and the conjugate cycle frequency β can be considered:

$$\widehat{\alpha} = \arg \max_{\gamma \in J(\alpha_0, \Delta\alpha)} \lambda_{rr^*}(\gamma) \tag{12}$$

$$\widehat{\beta} = \arg \max_{\gamma \in J(\beta_0, \Delta\beta)} \lambda_{rr}(\gamma) \,. \tag{13}$$

Then, accounting for (6) and (7), estimates of s and ν can be obtained by

$$\widehat{s} = \frac{\widehat{\alpha}}{\alpha_0} \qquad \widehat{\nu} = \frac{1}{2}(\widehat{\beta} - \widehat{s}\beta_0) \,. \tag{14}$$

Once the estimates $\widehat{s}$ and $\widehat{\nu}$ are available, a LTV equalization of the signal $r(t)$ can be made in order to compensate time-scale factor and frequency shift (but not delay and phase) in $y(t)$. This is obtained by a LTV equalizer with impulse-response function

$$h_{\mathrm{E}}(t, u) = \delta(u - t/\widehat{s})\, e^{-j2\pi\widehat{\nu}u} \,. \tag{15}$$

The equalized version of $r(t)$ is

$$\begin{aligned} r_{\mathrm{E}}(t) &= \int_{\mathbb{R}} h_{\mathrm{E}}(t, u)\, r(u)\, \mathrm{d}u \\ &= a\, z\Big(\frac{s}{\widehat{s}}t - d_0\Big)\, e^{j2\pi(\nu-\widehat{\nu})t/\widehat{s}} + n(t/\widehat{s}) \end{aligned} \tag{16}$$

The first term in the rhs of (16) is a “good estimate” of $a\, z(t - d_0)$ provided that the following conditions are satisfied

$$BT \ll \Big|1 - \frac{s}{\widehat{s}}\Big|^{-1} \qquad \Big|\frac{\nu - \widehat{\nu}}{\widehat{s}}\Big| T \ll 1 \tag{17}$$

where B is the bandwidth of $z(t)$. The first condition in (17) assures that the time-varying delay introduced by the (approximatively unit) time-scale factor $s/\widehat{s}$ within the observation interval $[0, T]$ can be neglected in the argument of $z(\cdot)$. The second condition in (17) assures that the maximum phase-change in $e^{j2\pi(\nu-\widehat{\nu})t/\widehat{s}}$ is negligible when t ranges in $[0, T]$. The equalized signal $r_{\mathrm{E}}(t)$ is approximatively jointly cyclostationary with $z(t)$, provided that conditions (17) hold.

The time-delay d_0 can be estimated by cross-correlating $r_{\mathrm{E}}(t)$ and a replica of $z(t)$ stored in the receiver. An interference-tolerant cyclostationarity-based alternative to the cross-correlation technique is the spectral coherence alignment (SPECCOA) method for complex signals [2]. The estimates of delay d_0 and phase ϕ_a are given by

$$\widehat{d_0} = \arg\max_{\tau} |I(\tau)| \qquad \widehat{\phi}_a = \measuredangle[I(\widehat{d_0})] \tag{18}$$

where $\measuredangle[\cdot]$ is the argument of a complex number and

$$I(\tau) \triangleq \int_{\mathbb{R}} \widehat{S}^{\alpha_0}_{r_{\mathrm{E}}z^*}(f)\, \widehat{S}^{\alpha_0}_{zz^*}(f)^*\, e^{j2\pi f\tau} \mathrm{d}f \,. \tag{19}$$

Since the signals of more satellites characterized by different C/A codes are generally present at the receiver input, the estimation procedure described by (12), (13), and (14) should be modified as follows. Assuming that a maximum of n satellites are visible, the n strongest peaks of the functions $\lambda_{rr^*}(\gamma)$ and $\lambda_{rr}(\gamma)$ must be located when γ ranges in $J(\alpha_0, \Delta\alpha)$ and $J(\beta_0, \Delta\beta)$, respectively. One obtains n estimates for $\widehat{\alpha}$ and $\widehat{\beta}$ and consequently n^2 pair estimates $(\widehat{s}, \widehat{\nu})$. The association among satellites (i.e., C/A codes) and corresponding $(\widehat{s}, \widehat{\nu})$ pair can be made choosing, for each local replica of C/A code, the pair $(\widehat{s}, \widehat{\nu})$ that maximizes the value of the peak of the function $|I(\tau)|$ with $r_E(t)$ given by (16).

5. NUMERICAL RESULTS

In this section, performance analysis of the parameter estimation procedure described in Section 4 is carried out in terms of root mean-square error (rmse) of the time-scale, frequency-shift, delay, and phase estimates for the receiver signal. Results are compared with those obtained by the NB-CAF.

The received L1-GPS signal is contaminated by circular AWGN with SNR = 0 dB and by an interfering BPSK signal with the same carrier frequency of the GPS signal and baud rate equal to $1/41T_s$, where $T_s = T_p/4$ is the sampling period. The signal-to-interference ratio (SIR) is equal to 5 dB. $\alpha_0 = \beta_0 = 1/T_c$ are used in the proposed method.

In Fig. 1, the normalized sample rmse of estimated parameters as function of the number N_b of processed chips of the C/A code is reported. In the experiment, 100 Monte Carlo runs are carried out. For values of N_b such that the narrow-band condition is satisfied ($N_b = 1023$), that is, for a data-record length equal to 1 ms, the proposed method and the NB-CAF technique have approximatively the same performance except for the delay, where the NB-CAF technique presents a smaller rmse. In contrast, when N_b, and hence the integration time, is increased in order to counteract the effects of the disturbance signals, the performance of the proposed method significantly improves while that of the NB-CAF technique worsens since it is based on the wrong received signal model ($s = 1$ in (4)).

6. CONCLUSION

A new synchronization technique for L1-GPS signal is proposed, which does not assume that the so called narrow-band condition is satisfied. Thus, data-record lengths significantly larger than those adopted with the classical narrow-band cross-ambiguity function technique can be adopted. The proposed method exploits the cyclostationarity of the transmitted and received signals and suitably models the transmitted and received signals as jointly spectrally correlated. In the large data-record regime, the proposed method is shown to significantly outperform the technique based on the narrow-band cross-ambiguity function in terms of rmse of the estimates. Moreover, it is shown to properly counteract the presence of noise and interference.

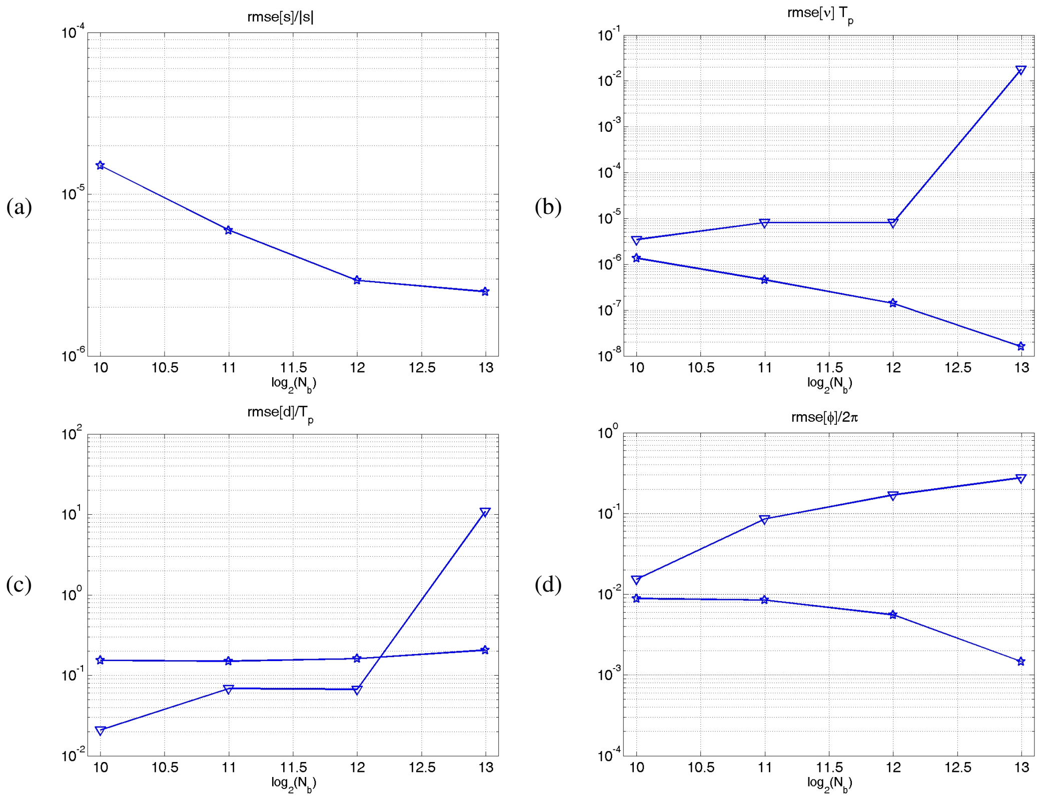

Fig. 1. Normalized sample rmse of estimated parameters as function of the number N_b of processed chips of the C/A code. $\star$ proposed method; $\triangledown$ NB-CAF method. (a) rmse of $\widehat{s}/|s|$; (b) rmse of $\widehat{\nu}T_p$; (c) rmse of $\widehat{d}_0/T_p$; (d) rmse of $\widehat{\phi}_a/2\pi$.

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