# PACRIM 2009

**Conference Program**

**Sunday August 23, 2009**

18:00 - 20:00

Welcoming Reception (University Club)

**Monday August 24, 2009**

8:45 - 9:00

Opening Ceremony (ECS-125)

9:00 - 9:30

Keynote Address (ECS-125)

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Acquisition of Spread Spectrum Codes with Consecutive Correlator Outputs in the Presence of Fractional Doppler Frequency Offset

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Abstract

A novel code acquisition scheme is proposed for the acquisition in time of spread spectrum codes in the presence of fractional Doppler frequency offset (FDFO). In the proposed acquisition scheme, the decision variable for detection is formed by combining two consecutive correlator outputs so that the influence of the reduction in the correlation peak due to the FDFO can be alleviated. Numerical results show that the proposed acquisition scheme can offer better mean-time-to-synchronization performance than the conventional scheme based on the cell-by-cell detection.

1. Introduction

The code synchronization is one of the important technical issues in spread-spectrum (SS) systems [1]. Normally, achieving the synchronization is called acquisition and maintaining the synchronization is called tracking, of which the former is dealt with in this paper.

Many SS-based wireless systems, e.g., the global positioning system (GPS) and CDMA2000, often cause a large Doppler frequency offset (DFO) between the transmitter and receiver [2]. In the presence of the DFO, the most common approach is to search, sequentially or in parallel, all possible DFOs (cells) over the DFO uncertainty region discretized with the cell spacing [3]-[5]. However, since the actual DFO is usually not an integer multiple of the cell spacing, a fractional DFO (FDFO) would still remain after the searching process is completed. In [4], such an effect of the FDFO is mentioned to a certain degree; however, no method for mitigating the influence of the FDFO is suggested.

In this paper, we formulate the influence of the FDFO on the acquisition of SS codes and then propose a novel detection method to alleviate the influence of the FDFO without a significant increase in the hardware complexity and acquisition time. In the proposed acquisition scheme, the decision variable for the correlation-based detection is formed by combining the correlation values corresponding to two consecutive cells, so that the decrease in the correlation value due to the FDFO can be compensated for without decreasing the cell spacing. Numerical results demonstrate that the proposed acquisition scheme can provide a significant improvement in performance over the conventional detector.

2. System Model

Considering a standard direct-sequence spread-spectrum (DS/SS) system, we can express the complex baseband
equivalent \( r(t) \) of the received signal as

\[
r(t) = \sqrt{P}e^{i(2\pi f_d t + \phi)}c(t - \tau T_c)d(t - \tau T_c) + w(t). \tag{1}
\]

In (1), \( P \) is the received signal power; \( f_d \) is the DFO; \( \phi \) is the carrier phase distributed uniformly over \([0, 2\pi)\); \( c(t) = \sum_{i=\infty}^{\infty} c_i p_{r_c}(t - i T_c) \) is the pseudo-noise (PN) code of period \( L \) chips with \( c_i \in \{-1, +1\} \) the \( i \)-th chip and \( p_{r_c}(t) \) the PN code waveform defined as a unit rectangular pulse over \([0, T_c)\); \( \tau \) is the code phase normalized to the PN code duration \( T_c \); \( d(t) \) is the data waveform; and \( w(t) \) is a complex additive white Gaussian noise process with mean zero and one-sided power spectral density \( N_0 \). For simplicity, we assume that the receiver is chipsynchronized to the received signal and uses a step size of one chip for the code synchronization. It is also assumed that there is a preamble for acquisition, so that no data modulation is present during acquisition (i.e., \( d(t) = 1 \)).

A typical PN code acquisition scheme that incorporates the DFO search process is as follows. The received signal \( r(t) \) is first compensated by a candidate DFO \( f_D^{(n)} \) (the \( n \)-th cell), where \( n = 1, 2, \cdots, U \) with \( U \) denoting the total number of cells in the DFO uncertainty region. Subsequently, the compensated signal is correlated with a locally generated PN code over a correlation time \( T \) in the non-coherent correlator. Finally, the correlator output is compared with a threshold. If the correlator output exceeds the threshold, the synchronization process is transferred to the tracking process: otherwise, the acquisition process resumes with the next candidates of the code phase and DFO.

3. Code Acquisition

3.1. The Influence of the DFO

When the phase of the locally generated code is \( \hat{\tau} \) (normalized to \( T_c \)), the correlator output \( x_n \) corresponding to \( f_D^{(n)} \) is expressed as

\[
x_n = \left| \frac{1}{T} \int_{(n-1)T}^{nT} r(t)e^{-i2\pi f_D^{(n)}r}c(t - \hat{\tau} T_c)dt \right|^2
\]

\[
= \left| \sqrt{P}R_c((\tau - \hat{\tau})T_c)\text{sinc}((f_d - f_D^{(n)})T) \right|^2
\]

\[
\cdot \left[ e^{\frac{i\pi}{2}(f_d - f_D^{(n)})T(2n-1) + \phi} + w_n \right]^2 \tag{2}
\]

for \( n = 1, 2, \cdots, U \), where \( R_c(\cdot) \) is a partial aperiodic autocorrelation function of the PN code \( c(t) \), sinc(\( y \)) is defined as \( \sin(\pi y)/(\pi y) \), and \( \{w_n\}_{n=1}^{U} \) are independent and identically distributed (i.i.d.) complex Gaussian random variables with mean zero and variance \( \sigma_w^2 = N_0/T \). When the noise is absent and the code synchronization is achieved (i.e., \( \tau = \hat{\tau} \)), \( x_n \) can be written as

\[
x_n = P\sin^2((f_d - f_D^{(n)})T). \tag{3}
\]

We can rewrite \( f_d - f_D^{(n)} = (p + \delta)\Delta_f \), where \( \Delta_f \) represents the cell spacing, \( p \) is an integer, and \( \delta \in [0, 1) \) is the DFO normalized to the cell spacing \( \Delta_f \). The cell spacing is typically of the form \( \Delta_f = (2^k T)^{-1} \), where \( k \) is a non-negative integer [3]. The correlator outputs \( x_n \) in the absence and presence of the DFO are shown in Fig. 1, where \( \Delta_f = T^{-1} \) and the arrows along the horizontal axis represent sampling instants. From this figure, we can clearly see that, when the DFO \( \delta \) is non-zero, the correlation peak would be smaller (than when \( \delta = 0 \)) even if the code synchronization is achieved, and consequently, the detection probability would be lower. Note that the detection in the code acquisition process is based on the comparison between the correlation value and a given threshold.

3.2. Proposed Scheme

In Fig. 1, it is observed that the signal power in the single correlation peak for \( \delta = 0 \) splits into smaller correlation peaks when \( \delta \neq 0 \), with most of the signal power contained in the two peaks

\[
x_{n-1} = P\sin^2(1 - \delta) \tag{4}
\]
and
\[ x_n = P \text{sinc}^2(\delta). \]  

The proposed acquisition scheme is motivated by the implication of (4) and (5): if the sum of two consecutive correlator outputs is used as the decision variable during the detection process, (most of) the signal power split by the FDFO can be combined and then used efficiently for detection.

As the proposed acquisition scheme can maintain relatively constant signal power under the variation of \( \delta \), the proposed acquisition scheme is expected to offer better and more robust detection performance in the presence of the FDFO, compared to the conventional cell-by-cell scheme in which the correlator outputs are individually used for detection. It should be mentioned that we can obtain similar results for any \( \Delta_f < T^{-1} \); 1 − \( \delta \) in (4) and (5) are simply replaced by \( (1 - \delta)T\Delta_f \) and \( \delta T\Delta_f \), respectively, when \( \Delta_f < T^{-1} \).

It is noteworthy that several techniques similar to the proposed acquisition scheme have been introduced and analyzed to deal with the fractional chip offset problem in the literature. Specifically, exploiting a differential combining (DC) of two matched filter outputs, a differentially coherent detection technique is proposed for code acquisition in chip-synchronous/asynchronous environments in [6]. The DC scheme was shown to be effective in dealing with the fractional chip offset. Unfortunately, it turns out that the DC is not suitable for addressing the fractional frequency offset problem of this paper as explained in the following. Assume that a DC is taken over two matched filter outputs. From (2) and Fig. 1, then, it can be easily seen that the signal amplitudes of \( x_{n-1} \) and \( x_n \) would be \( P \text{sinc}(2 - \delta) \text{sinc}(1 - \delta) \) and \( P \text{sinc}(1 - \delta) \text{sinc}(\delta) \), respectively, when the code synchronization is achieved. Here, we can observe that both of the signal amplitudes \( P \text{sinc}(2 - \delta) \text{sinc}(1 - \delta) \) and \( P \text{sinc}(1 - \delta) \text{sinc}(\delta) \) become 0 when the fractional frequency offset \( \delta = 0 \) or 1. Furthermore, it is straightforward to show that \( P \{\text{sinc}^2(1 - \delta) + \text{sinc}^2(\delta)\} \) is larger than any other logical combination \( P \{\text{sinc}(2 - \delta) \text{sinc}(1 - \delta) + \beta \text{sinc}(1 - \delta) \text{sinc}(\delta)\} \) of the DC outputs, where \( |\alpha| \leq 1 \) and \( |\beta| \leq 1 \).

These observations imply that a detection technique using the signal amplitudes obtained through the DC, either jointly or individually, would result in a code acquisition scheme with performance inferior to that of the proposed acquisition scheme in the fractional frequency offset problem. In essence, a DC-based scheme for the fractional frequency offset problem cannot maintain a constant signal power under the variation of \( \delta \) when compared with the proposed acquisition scheme. In addition, to deal with the fractional chip offset problem again, the Scheme 2 in [7] exploits the sum of two samples obtained via fractional sampling, a processing similar to that in the proposed acquisition scheme. Yet, the proposed acquisition scheme is vividly distinct from the Scheme 2 in that the proposed acquisition scheme does not need a cell spacing reduction (which is the counterpart of the fractional sampling) in its operation, while the Scheme 2 requires a fractional sampling prior to summing two samples in solving the fractional chip offset problem. It should also be added that the proposed acquisition scheme offers good performance in mitigating the influence of the fractional frequency offset, regardless of whether or not a cell spacing reduction is employed, as we shall see in Figs. 2 and 3 later in Section 5.

4. Performance Analysis

In this section, we derive the detection and false alarm probabilities of both the proposed acquisition scheme and conventional cell-by-cell scheme, which will eventually be used in the evaluation of the mean-time performance in Section 5. In deriving the probability expressions, the correlation time \( T \) is assumed to be long enough compared to the chip duration \( T_c \) so that the correlation between the received and locally generated PN codes can be accurately approximated as zero when the phases of the received and locally generated codes are not synchronized.

4.1. Detection Probability

From (2), it is easy to see that the correlator output \( x_n \) is a noncentral chi-square variable with two degrees of freedom when the phases of the received and locally generated codes are synchronized. As a consequence, the decision variable

\[ y_n = x_n + x_{n-1} \]  

in the proposed acquisition scheme obeys the noncentral chi-square distribution with four degrees of freedom when the code synchronization is achieved. Hence, it should be noted that \( x_0 = x_U \). Thus, the probability density functions (pdfs) \( f^1_{x_n}(x) \) and \( f^1_{y_n}(y) \) of \( x_n \) and \( y_n \) are given by

\[ f^1_{x_n}(x) = e^{-(x+s_n)}I_0(2\sqrt{s_n}x) \]  

and

\[ f^1_{y_n}(y) = \sqrt{y \over s_n + s_{n-1}}e^{-(y+s_n+s_{n-1})}I_1(2\sqrt{(s_n+s_{n-1})y}), \]  

respectively, for \( n = 1, 2, \ldots, U \), when normalized by \( s_w^2 \).

In (7) and (8), \( s_n = PTN_0^{-1}\text{sinc}^2((f_D-f'_{D})(T)), s_0 = s_U \), and

\[ I_l(u) = \sum_{m=0}^{\infty} \frac{(u/2)^{l+m}}{m!\Gamma(l+m+1)} \]  

(9)
is the \(l\)-th order modified Bessel function of the first kind [8] with

\[
\Gamma(\alpha) = \int_0^\infty x^{\alpha-1} e^{-x} \, dx
\]  

(10)

the gamma function.

The detection in a code acquisition process can be defined as an event that the decision variable of a detector exceeds a given threshold when the phases of the received and locally generated codes are synchronized. Considering that there are \(U\) decision variables (corresponding to \(U\) cells) under the situation that the phases of the received and locally generated codes are synchronized, the detection probability is the probability that any one of the \(U\) decision variables exceeds the threshold. Thus, the detection probabilities \(P_D^p\) and \(P_D^c\) of the proposed acquisition scheme and conventional cell-by-cell scheme are given by

\[
P_D^p = 1 - \Pr \{ y_1 < \eta_p, y_2 < \eta_p, \cdots, y_U < \eta_p | \gamma = \hat{\gamma} \}
\]  

(11)

and

\[
P_D^c = 1 - \Pr \{ x_1 < \eta_c, x_2 < \eta_c, \cdots, x_U < \eta_c | \gamma = \hat{\gamma} \}
\]

\[
= 1 - \prod_{n=1}^U \left[ 1 - Q\left( \sqrt{2\eta_n}, \sqrt{2\eta_n} \right) \right]
\]  

(12)

respectively. Here, \(\eta_p\) and \(\eta_c\) denote the thresholds of the proposed acquisition scheme and conventional scheme, respectively, and

\[
Q(a, b) = \int_b^\infty e^{-(u^2+a^2)/2} I_0(au) \, du
\]  

(13)

is the Marcum \(s\) \(Q\) function [9]. Unfortunately, it is highly complicated, if not impossible, to express \(P_D^p\) in a closed form due to the dependence between adjacent decision variables. In this paper, hence, Monte Carlo integration [10] is employed to evaluate \(P_D^p\).

4.2. False Alarm Probability

A false alarm occurs if any one of the \(U\) decision variables exceeds the threshold when the phases of the received and locally generated codes are not synchronized. In the case of the code phase being out of synchronization, the decision variables \(x_n\) and \(y_n\) are central chi-square distributed variables with two and four degrees of freedom, respectively. The pdfs \(f_{x_n}^0(x)\) and \(f_{y_n}^0(y)\) of \(x_n\) and \(y_n\), normalized by \(\sigma_n^2\), are thus given by

\[
f_{x_n}^0(x) = e^{-x}
\]  

(14)

for \(n = 1, 2, \cdots, U\) from (7) by noting \(I_0(0) = 1\), and

\[
f_{y_n}^0(y) = ye^{-y}
\]  

(15)

from (8) by noting that

\[
\sqrt{y} I_1\left(2\sqrt{(s_n + s_{n-1})y}\right) = \frac{\sqrt{y}}{\sqrt{s_n + s_{n-1}}}
\]

\[
\cdot \left\{ \sqrt{(s_n + s_{n-1})y} + \frac{(s_n + s_{n-1})y}{2} + \cdots \right\}
\]

\[
\rightarrow y
\]  

(16)

when \(s_n + s_{n-1} \rightarrow 0\). Thus, the false alarm probabilities \(P_{FA}^p\) and \(P_{FA}^c\) of the proposed acquisition scheme and conventional cell-by-cell scheme can be obtained as

\[
P_{FA}^p = 1 - \Pr \{ y_1 < \eta_p, y_2 < \eta_p, \cdots,
\]

\[
y_U < \eta_p | \gamma \neq \hat{\gamma} \}
\]  

(17)

and

\[
P_{FA}^c = 1 - \Pr \{ x_1 < \eta_c, x_2 < \eta_c, \cdots,
\]

\[
x_U < \eta_c | \gamma \neq \hat{\gamma} \}
\]

\[
= 1 - [1 - e^{-\eta_c}]^U
\]  

(18)

respectively. As in (11), Monte Carlo integration is used to evaluate \(P_{FA}^c\). The thresholds \(\eta_p\) and \(\eta_c\) are determined by specifying the false alarm probabilities (17) and (18), respectively.

5. Numerical Results and Discussion

In evaluating the performance, we assume the following parameters: a PN code of \(L = 32767\) chips (generated from an \(m\)-sequence with the primitive polynomial \(1 + z + z^{15}\)), chip rate = 1 MHz, DFO range = \(\pm 10\) kHz, \(T = 1000T_c\), \(\Delta_f = T^{-1}\) and \((2T)^{-1}\) (corresponding to \(\{\Delta_f = 1\) kHz, \(U = 21\}\) and \(\{\Delta_f = 500\) Hz, \(U = 41\}\), respectively), and \(P_{FA}^p = P_{FA}^c = 10^{-2}\). Monte Carlo integrations with \(10^7\) samples are performed to evaluate \(P_D^p\) and \(P_{FA}^p\). We consider a two-dwell system consisting of a search mode, followed by a verification mode in the form of a coincidence detector (CD) [11]. In the verification mode, the CD declares an acquisition if at least \(B\) out of \(A\) tests exceed a threshold. As suggested in [11], we have chosen \(A = 4\) and \(B = 2\) for the verification mode.

The mean-time-to-synchronization (MTTS), the time that elapses prior to acquisition on the average, is used as the performance metric. The MTTS can be calculated as

\[
T_M = \frac{2 - P_D P_{D1}}{2P_D P_{D1}} \left[ 1 + A P_{FA} + J P_{FA} P_{FA1} \right] \cdot \frac{LT}{T}
\]  

(19)

using the flow-graph method in [12], where the penalty time \(J \) in chips due to a false alarm is set to \(10^4\), and

\[
P_{D1} = \sum_{n=B}^{A} \left( \begin{array}{c} A \nonumber \cr n \end{array} \right) P_D (1 - P_D)^{A-n}
\]  

(20)
correlator outputs, as mentioned in Section 3. The proposed acquisition scheme is more robust to the variation of the signal power split, utilizing more accurate information on the signal from more reliable correlator outputs during the detection process. Consequently, the proposed acquisition scheme provides better performance than the conventional scheme.

Let us add a brief discussion on the performance of the proposed acquisition scheme with respect to combined tim-

![Figure 2. Mean-time-to-synchronization of the proposed acquisition scheme and conventional scheme for $\delta = 0$ and 0.5 when $\Delta f = T^{-1}$.](image)

![Figure 3. Mean-time-to-synchronization of the proposed acquisition scheme and conventional scheme for $\delta = 0$ and 0.5 when $\Delta f = (2T)^{-1}$.](image)

The proposed acquisition scheme is higher than that of the conventional scheme, resulting in better MTTS performance. Another important observation in Fig. 2 is that the proposed acquisition scheme is more robust to the variation of $\delta$ than the conventional scheme. This is because the proposed acquisition scheme inherently maintains relatively constant signal power under the variation of $\delta$, through the combining of two consecutive correlator outputs, as mentioned in Section 3.

and

$$P_{FA1} = \sum_{n=B}^{A} \binom{A}{n} P_{FA}^{n} (1 - P_{FA})^{A-n} \quad (21)$$

are the detection and false alarm probabilities, respectively, in the verification mode, with $\binom{A}{n}$ the binomial coefficient. Fig. 2 shows the MTTS of the proposed acquisition scheme and conventional scheme for $\delta = 0$ (best case) and 0.5 (worst case) when $\Delta f = T^{-1}$. Here, the signal to noise ratio per chip (SNR/chip) is defined as $P_{Tc}/N_0$. When $\delta = 0$, only a single correlation peak exists at which all of the signal power is concentrated. Thus, the sum of two consecutive correlator outputs in the proposed acquisition scheme naturally leads to an increase of the noise variance in the correlation peak. Consequently, the performance of the proposed acquisition scheme is slightly inferior to that of the conventional scheme.

On the other hand, when $\delta = 0.5$, most of the signal power spreads over two correlation peaks of normalized amplitude $\text{sinc}^2(0.5) \approx 0.405$. The proposed acquisition scheme exploits two consecutive correlator outputs jointly for detection, combining more efficiently the signal power spread by the FDFO than the conventional scheme which uses the correlator outputs individually for detection. Thus, the average detection capability of the proposed acquisition scheme is higher than that of the conventional scheme, resulting in better MTTS performance. Another important observation in Fig. 2 is that the proposed acquisition scheme is more robust to the variation of $\delta$ than the conventional scheme.

When $\delta = 0.5$, the signal power spreads over mainly two correlation peaks of normalized amplitude $\text{sinc}^2(0.25) \approx 0.811$: unlike in the case of $\Delta f = T^{-1}$, on the other hand, the two neighboring peaks have a larger normalized amplitude (0.811 versus 0.405). Thus, the performance for $\delta = 0.5$ is closer to that for $\delta = 0$ in both the proposed acquisition scheme and conventional scheme.

In short, while the correlator outputs that contain the signal power split by the FDFO are used individually in the conventional scheme, the correlator outputs are used jointly in the proposed acquisition scheme. Thus, the proposed acquisition scheme more efficiently combines the signal power split, utilizing more accurate information on the signal from more reliable correlator outputs during the detection process. Consequently, the proposed acquisition scheme provides better performance than the conventional scheme.
ing and frequency synchronization. If we took the frequency acquisition into account in the analysis via two-stage approaches for each of the time and frequency acquisition processes (for example, delay or timing detection, followed by delay verification, followed by frequency detection and perhaps frequency verification), the absolute performance of the proposed acquisition scheme and conventional scheme would be different from that shown in this paper. Nonetheless, we believe that the analysis and results provided in this paper should still be applicable in that the relative performance of the proposed acquisition scheme and conventional scheme would not change when an identical frequency acquisition scheme is employed in both schemes. On the other hand, if some other combined synchronization techniques (for example, joint delay-and-frequency detection followed by a single verification stage) are considered, the superiority of the proposed acquisition scheme over the conventional scheme in terms of the performance of the combined time and frequency acquisition is not quite apparent. We have reserved the issue of performance analysis of various combined synchronization techniques for further study.

6. Conclusion

In this paper, we have proposed a novel detection scheme for the acquisition of SS codes in the presence of fractional Doppler frequency offset.

We have derived the detection and false alarm probabilities of the proposed acquisition scheme, which have in the sequel been used in examining the mean-time performance. The mean-time-to-synchronization performance of the proposed acquisition scheme has been analyzed and discussed in comparison with that of the conventional cell-by-cell detection scheme. From the numerical results we have obtained, it is observed that the proposed acquisition scheme has better mean-time-to-synchronization performance regardless of whether or not the cell spacing reduction is used and that the proposed acquisition scheme is more robust to the variation of the fractional Doppler frequency offset than the conventional cell-by-cell detection scheme.

Acknowledgments

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References


