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Low-complexity code tracking loop with chip-level differential detection for DS/SS receivers

Jia-Chin Lin

A new code tracking loop is proposed for direct-sequence spread-spectrum communications. The main features of such a technique are its lower complexity together with its good tracking performance. Analytical expressions for the error characteristic are derived and timing jitter is evaluated by computer simulation.

Introduction: Direct-sequence spread-spectrum code-division multiple access (DS/CDMA) has recently become the most popular system for commercial applications. Code tracking is one of the most important functions in DS/SS receivers, and substantial effort has gone into solving this problem [1, 2]. A differentially coherent delay-locked loop has also been proposed [3]. However, it cannot deal with data modulation, and can thus only be used in ranging applications. In addition, migration towards digital implementation of modems is currently one of the main trends in communications systems [4]. In this Letter, a fully digital low-complexity code tracking loop suitable for recent commercial applications is proposed.

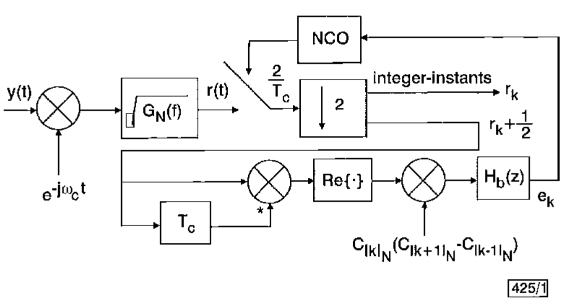

Fig. 1 *Proposed code tracking loop with chip-level differential detection*

Signal model and system description: For bandlimited DS/SS communications [4], the complex envelope of the modulated signal can be expressed as $s(t) = \Sigma_{m=-\infty}^{\infty} a_{\{m\}_M} c_{|m|_N} g_T(t - mT_c)$, where $a_i = \pm 1$ represents BPSK information-bearing symbols, T and T_c are the symbol duration and chip interval, respectively, $M = T/(T_c)$ is the processing gain, $c_i = \pm 1$ is the ith chip value of the PN code, N is the code length, $\{m\}_M$ and $|m|_N$ are the integer quotient and m modulus N, respectively. $g_T(t)$ is the transmitted chip-shape, the Fourier transform of which is $G_T(f) = T_c\sqrt{G_N(f)}$, where $G_N(f)$ is the frequency response of the Nyquist raised-cosine filter. It can be shown that the baseband power spectral density (PSD) of the signal $s(t)$, being proportional to $|G_N(f)|^2$, is constrained to the interval $\pm(1 + \alpha)/2T_c$ where α is the roll-off factor of the raised-cosine pulse. In the receiving end, the baseband equivalent $y(t)$ of the signal at the input of the DS/SS noncoherent down-converter is $y(t) = s(t) + n'(t)$, where $n'(t) = n'_c(t) + jn'_s(t)$ is complex AWGN the quadrature components of which own the PSD $S_n'(f) = N_0/2P$, P being the IF signal power. The complex representation of the baseband signal at the output of the chip-matched filter with its transfer function $\sqrt{G_N(f)}$ is therefore $r(t) = e^{j\theta} \Sigma_{m=-\infty}^{\infty} a_{\{m\}M} c_{|m|N} g(t - mT_c) + n''(t)$, where the PSD of the noise components is now $S_n''(f) = (N_0 G_N(f))/2P$, and the PSD of the overall pulse shape is $G(f) = T_c G_N(f)$. The oversampler samples $r(t)$ at the instants $t_k = (k + \varepsilon_k)T_c$ and $t_{k+1/2} = (k + \varepsilon_k + 1/2)T_c$ (i.e. sampling rate $2/T_c$), where ε_k is the kth normalised chip timing error. The integer-instant samples $r_k = r(t_k)$ are fed into the following information detection processes, while the half-integer-instant samples $r_{k+1/2} = r(t_{k+1/2})$ are exploited in the tracking loop. Note that the actual processing rate of both the information detection processes and the code tracking subsystem is exactly 1 sample/chip. The half-integer-instant samples given by $r_{k+1/2} = r[(k+\varepsilon_k + 1/2)T_c]$ are multiplied by the complex conjugate of its one-chip-delay replicas. We take the real part and then multiply by the difference (i.e. $c_{|k+1|N}c_{|k|N} - c_{|k|N}c_{|k-1|N}$) of the two local differential sequences. The resultant samples are fed into the branch filters h_k which are first-order lowpass filters with bandwidth B_b and the following transfer function: $H_b(z) = 1{-}a/(1{-}az^{-1})$, $a = \exp(-2\pi B_b T_c)$, where B_b may be of the same order of magnitude as $1/T$. The error signal at the input of the NCO can thus be expressed by

$$e_k = \mathrm{Re}\{r_{k+\frac{1}{2}} \cdot r^*_{k-\frac{1}{2}}\} \times [c_{|k|_N}(c_{|k+1|_N} - c_{|k-1|_N})] \otimes h_k$$

It is assumed that code acquisition has been accomplished (i.e. $|\varepsilon_k| \le 1/2$). Therefore, the integration inherently performed by the NCO can be represented by the following loop equation: $\varepsilon_{k+1} = \varepsilon_k - \gamma e_k$, where γ is the NCO sensitivity. The average loop error characteristic can be defined as $\eta(\varepsilon) = \langle E\{e_k|\varepsilon_k = \varepsilon, \forall k\}\rangle$, where the operator $\langle\cdot\rangle$ indicates time-averaging. Since the bandwidth of the branch filters is comparable with $1/T$ (hence, much narrower than $1/T_c$), for $\varepsilon_k = \varepsilon$ after some manipulations, the loop error characteristic can be rewritten as

$$\eta(\varepsilon) = (1 - \frac{1}{M})[g^2(\varepsilon - \frac{1}{2}) - g^2(\varepsilon + \frac{1}{2}) + g(\varepsilon + \frac{1}{2})g(\varepsilon - \frac{3}{2}) - g(\varepsilon + \frac{3}{2})g(\varepsilon - \frac{1}{2})] \quad (1)$$

The error characteristics of the DDLL [4] and the proposed technique derived from the above statistical analyses and using Monte Carlo methods on a computer are shown in Fig. 2. It is obvious that the simulation results are very close to the theoretical ones, and that the proposed technique has a slightly narrower error characteristic and a slightly higher slope at $\varepsilon = 0$.

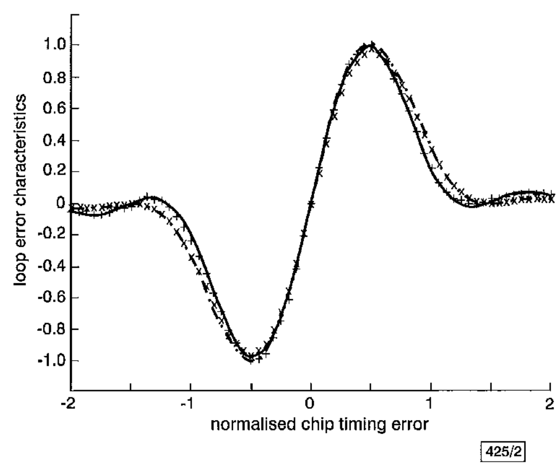

Fig. 2 *Simulated loop error characteristics of DDLL [4] and proposed technique and those obtained by theoretical analysis*

Theoretical:
- · - · DDLL
– – – – proposed

Simulated:
× DDLL
+ proposed

Steady-state timing jitter: Since the overall chip shape is given by $g(\varepsilon T_c) = \sin(\pi\varepsilon)/\pi\varepsilon \cdot \cos(\pi\alpha\varepsilon)/1—(2\alpha\varepsilon)^2$, the slope A of the loop error characteristic at $\varepsilon = 0$ can be found to be

$$\begin{aligned} A &= \frac{d\eta(\varepsilon)}{d\varepsilon}|_{\varepsilon=0} \\ &= 16\cos(\frac{\pi\alpha}{2}) \\ &\quad \cdot\frac{\pi\alpha(1-\alpha^2)\sin(\pi\alpha/2)+2(1-3\alpha^2)\cos(\pi\alpha/2)}{\pi^2(1-\alpha^2)^3} \\ &+8\cos(\frac{3\pi\alpha}{2}) \\ &\quad \cdot\frac{\pi\alpha(1-\alpha^2)\sin(\pi\alpha/2)+2(1-3\alpha^2)\cos(\pi\alpha/2)}{3\pi^2(1-\alpha^2)^2(1-9\alpha^2)} \\ &-8\cos(\frac{\pi\alpha}{2}) \\ &\quad \cdot\frac{3\pi\alpha(1-9\alpha^2)\sin(3\pi\alpha/2)-2(1-27\alpha^2)\cos(3\pi\alpha/2)}{9\pi^2(1-\alpha^2)(1-9\alpha^2)^2} \end{aligned} \quad (2)$$

By exploiting the above analysis and the standard linear loop analysis, the normalised chip timing jitter can be derived. Only the simulation results of the RMS normalised chip timing jitter with the DDLL [4] and the proposed technique are reported in Fig. 3. It can be seen that the performance difference between the DDLL [4] and the proposed technique is very slight.

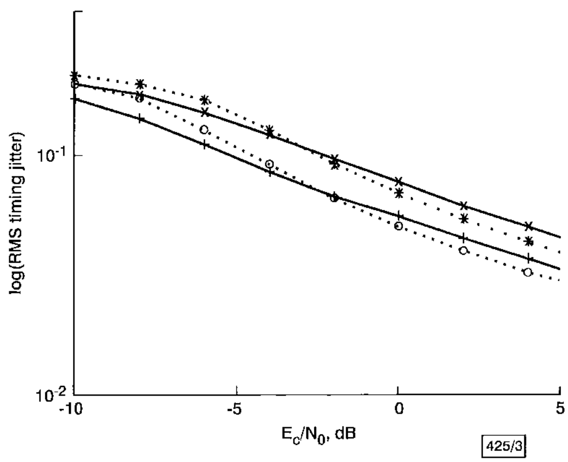

Fig. 3 *RMS normalised chip timing error of DDLL [4] and proposed technique by computer simulation*

····○···· proposed ($B_L T_C$ = 0.01)
····✳···· proposed ($B_L T_C$ = 0.02)
—+— DDLL ($B_L T_C$ = 0.01)
–×– DDLL ($B_L T_C$ = 0.02)

Complexity comparison: Except for the common components employed in both the compared schemes, the proposed technique needs only one real correlator (including the branch filter h_k), two real multipliers, one adder and one delay element, while the DDLL [4] requires four real correlators, four real multipliers, three adders and one delay element. As a result, the proposed technique achieves significant complexity reduction.

Conclusion: A low-complexity code tracking loop has been proposed by taking advantage of chip-level differential detection. Analytical expressions for the error characteristic are derived and the timing jitter has been evaluated by computer simulation.

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Nonlinear space-time decorrelator for multiuser detection in non-Gaussian channels

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A space-time detection scheme combining nonlinear decorrelators and antenna array is investigated for jointly mitigating multiple access interference, multipath fading and impulsive noise. Monte Carlo simulation results of the proposed space-time detector are presented to justify the relative merits of nonlinear signal processing techniques in the spatial-temporal domain.

Introduction: The last few decades have witnessed tremendous progress in multiuser detection [1] as the driving technology for direct-sequence code-division multiple-access (DS-CDMA) communications. A key assumption of these works has been the use of the Gaussian model for the ambient noise. Unfortunately, the wireless environments are often corrupted by interference that exhibits impulsive statistics [2]. Since linear detection schemes often perform poorly in impulsive noise, this motivates the use of nonlinear signal processing techniques. In addition, the urban propagation environments produce multipath fading that degrades signal quality. Antenna array techniques have been found attractive in mitigating multipath fading. In [3], a space-time detector based on the linear decorrelator [4] combined with spatial filters has been investigated. However, a recent study has shown that the least squares-based decorrelator is extremely susceptible to the presence of impulsive noise [5]. This Letter examines a space-time structure by combining the nonlinear decorrelator with adaptive spatial filters.

Signal model: In this Letter we consider a synchronous CDMA system under flat fading channels, i.e. where multipath propagation does not induce temporal delays, but angular spreads. In addition, we assume a slowly varying channel so that the channel parameters are considered as constants. For K users and an M-element receiving array, the received signal at the mth antenna element during the ith symbol of interval T is captured by chip-matched filtering and then sampled at the chip rate $1/T_c$:

$$r_{m,j}(i) = \sum_{k=1}^{K} A_k b_k(i) s_j^k \sum_{l=1}^{L} a_{kl,m} g_{kl} + n_{m,j}(i), \qquad j = 1, \ldots, N \quad (1)$$

or in vector form

$$\mathbf{r}_m(i) = \sum_{k=1}^{K} A_k b_k(i) \mathbf{s}_k \sum_{l=1}^{L} a_{kl,m} g_{kl} + \mathbf{n}_m(i) \in \mathbf{C}^N, \qquad m = 1, \ldots, M \quad (2)$$

where N is the processing gain with $NT_c = T$ and L is the number of multipaths in each user's channel. With respect to the kth user, A_k, $b_k(i) \in \{\pm 1\}$, $s_j^k \in \{\pm 1/\sqrt{N}\}$ and g_{kl} denote, the transmitted amplitude, ith information bit, normalised signature code for the jth chip, complex channel gain of the lth path, respectively. $\mathbf{s}_k = [s_1^k \ldots s_N^k]^T$ and $\mathbf{n}_m(i) = [n_{m,1}(i) \ldots n_{m,N}(i)]^T$ are the vector of independent zero-mean complex spatially and temporally white ambient noise which is being modelled as α-stable random processes