NON-DATA AIDED TIMING ACQUISITION OF ULTRA-WIDEBAND TRANSMISSIONS USING CYCLOSTATIONARITY*

Liuqing Yang[1], *Zhi Tian*[2], and *Georgios B. Giannakis*[1]

[1]Dept. of ECE, Univ. of Minnesota, Minneapolis, MN 55455

[2]Dept. of ECE, Michigan Tech. Univ., Houghton, MI 49931

ABSTRACT

Low-complexity rapid timing acquisition constitutes a major challenge in realizing the high potential that ultra-wideband (UWB) wireless technology promises for indoor communications. We derive and test two such timing acquisition algorithms which capitalize on the cyclostationarity that is naturally present in UWB transmissions. Our novel schemes are blind, they do not require multiple antennas or oversampling, and rely on frame-rate sampling which reduces complexity and acquisition delay, considerably.

1. INTRODUCTION

With the Ultra Wideband (UWB) spectral mask (in the range of 3.1-10.6GHz) released by FCC in February 2002, the interest for commercial applications of UWB technology is growing fast, especially in the areas of indoor wireless for short-range communications. This interest stems from several attractive features UWB comes with: low-power, impulse-like, baseband transmissions with improved penetration capability; rich multipath diversity that can be collected with low-complexity RAKE reception; a large number of users allowed access with Time Hopping (TH) codes; and potential to overlay existing narrowband systems, such as IEEE 801.11 and Bluetooth, with reduced interference (noise-like) characteristics; see e.g., [2, 5, 7] and references therein.

One of the most critical challenges (at least at the physical layer) in enabling the unique benefits of UWB transmissions, is the clock synchronization step (a.k.a. timing offset estimation), the difficulty of which is accentuated in UWB due to the fact that the information bearing waveforms are impulse-like, and have low amplitude [6]. Peak-picking the output of a sliding correlator between the received signal and the transmit-waveform template is not only sub-optimum in the presence of dense multipath, but also results in unacceptably slow acquisition times, and has prohibitive complexity when one has to perform exhaustive search over thousands of bins (chips). Recent attempts to improve acquisition speed include a coarse bin reversal search considered in [4] for the noiseless case, and the design of a coded beacon sequence in conjunction with a bank of correlators in the context of data-aided localization [1].

This paper develops low-complexity rapid acquisition schemes for *non-data aided* (a.k.a. blind) timing acquisition of UWB transmissions. The novel approach exploits the cyclostationarity (CS) that emerges as UWB pulses are periodically repeated (one per frame) across the multiple frames comprising each symbol. Unique to UWB, this is on top of the well known CS that arises (even in narrowband systems) due to the repeated use of the symbol waveform [9]. For narrow-band systems, CS-based timing acquisition was dealt with in [3]. Different from [3], the timing algorithms here neither require oversampling nor they rely on multiple antennas to induce CS, simply because they capitalize on the CS that is naturally present in UWB signaling. Because they rely on frame-rate (as opposed to chip-rate) samples, they can reduce complexity and acquisition speed by one or two orders of magnitude. Equally attractive is their blind estimation capability that makes them particularly suitable for "cold start-up" scenarios.

The ensuing Section 2 outlines our system model and operating transceiver conditions. Section 3 derives two novel acquisition algorithms: one based on the periodically time-varying correlation, and the other based on the cyclic correlation. Section 4 presents preliminary simulations. For tracking issues, we refer the reader to [9]. Due to lack of space, detailed derivations of our acquisition and tracking algorithms are included in the journal version [8].

2. MODELING AND PROBLEM STATEMENT

In impulse radio multiple access, every information symbol is transmitted by repeating over N_f frames (each of duration T_f) an ultra short pulse $p(t)$ that has duration $T_p \ll T_f$. The pulse (a.k.a. monocycle) can have rectangular, triangular, or, typically Gaussian shape [7]. With T_p at the sub-nanosecond scale, $p(t)$ is UWB with bandwidth $B_s \approx 1/T_p$. User separation is accomplished with pseudo-random TH codes, a different one per user, which time-shift the pulse positions at multiples of the chip duration (T_c) [7]. Here we focus on *a single user link*, and treat multi-user interference (MUI) as noise. Although generalizations to pulse position modulation (PPM) are possible, for convenience we will deal with UWB binary pulse amplitude modulation (PAM), as in [5]. The information-bearing PAM symbols $s(k)$ have zero mean and variance 1, and the transmitted waveform is given by:

$$u(t) = \sqrt{\mathcal{P}} \sum_{k=-\infty}^{+\infty} s(k) p_s(t - kN_fT_f - c_kT_c), \qquad (1)$$

where $c_k \in [0, N_c - 1]$ is the TH code during the kth symbol duration, and $\mathcal{P}$ stands for the power. Notice that the transmit filter here refers to the *symbol waveform* which contains N_f monocyles:

$$p_s(t) := \sum_{n=0}^{N_f-1} p(t - nT_f). \qquad (2)$$

After multipath propagation, the received waveform is given by: $x(t) = \sqrt{\mathcal{P}} \sum_{k=-\infty}^{+\infty} s(k) \sum_{l=0}^{L} \alpha_l p_s(t - kN_fT_f - c_kT_c - \tau_l) + w(t)$, where $L+1$ is the number of paths, each with amplitude α_l, and delay τ_l satisfying $\tau_l < \tau_{l+1}$, $\forall l$. We assume that the channel is quasi-static, which implies that $\{\alpha_l\}_{l=0}^{L}$ and $\{\tau_l\}_{l=0}^{L}$ remain invariant over one symbol period, but they are allowed to change independently from symbol to symbol. The multipath is sufficiently

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rich to ensure that $\tau_{l+1} - \tau_l < 2T_p, \forall l$. This assumption is well justified for indoor environments. It simplifies our development, but even when it is violated, the basic approach herein carries over with simple preprocessing [8]. The additive noise $w(t)$ is independent of $s(k)$, $\{\alpha_l\}_{l=0}^{L}$ and $\{\tau_l\}_{l=0}^{L}$, zero-mean, wide-sense stationary, but not necessarily white and/or Gaussian, as it consists of both ambient noise and MUI.

In a *non-data aided* mode, the receiver cannot distinguish between two time delays that are separated by an integer multiple of the symbol duration, e.g., τ_0 and $\tau_0 + kN_fT_f$. Thus, without loss of generality, we can confine our timing offset acquisition problem to within a symbol duration (N_f frames), and express the first arrival time as: $\tau_0 = N_\epsilon T_f + \epsilon$, where $N_\epsilon \in [0, N_f - 1]$, and $\epsilon \in [0, T_f)$. Accordingly, other path delays will be described uniquely by: $\tau_{l,0} := \tau_l - \tau_0$. With these definitions, the received waveform can be expressed as:

$$x(t) = \sqrt{\mathcal{P}} \sum_{k=-\infty}^{+\infty} s(k) \sum_{l=0}^{L} \alpha_l \cdot p_s(t - kN_fT_f - c_kT_c - \tau_{l,0} - N_\epsilon T_f - \epsilon) + w(t). \quad (3)$$

Since the frame duration T_f is a design parameter, we choose it as $T_f = \tau_{L,0} + T_p$, so as to avoid inter-frame interference.

Given frame-rate samples of $x(t)$, our timing *acquisition* goal in this paper amounts to estimating N_ϵ. Due to lack of space, we defer development of *tracking* algorithms (estimation of ϵ) to [8].

3. TIMING ACQUISITION

With $p_s(t)$ as a template, the correlator receiver yields the continuous-time output $y(t) = \int_{t-N_fT_f}^{t} p_s(\lambda)x(\lambda)d\lambda$. Sampling at frame-rate, we obtain:

$$y(n) := y(t)|_{t=(n+N_f)T_f} = \sqrt{\mathcal{P}} \sum_{k=-\infty}^{+\infty} s(k) r_{pp}(n - kN_f) + \eta(n), \quad (4)$$

where $\eta(n)$ denotes sampled noise, and $r_{pp}(n)$ is the discrete-time impulse response of the overall channel, capturing the transmit filter, the multipath propagation channel, as well as the correlator at the receiver. Substituting $p_s(t)$ from (2) and $x(t)$ from (3), we can express $r_{pp}(n - kN_f)$ as [c.f. (4)]:

$$r_{pp}(n - kN_f) = \sum_{m_1=0}^{N_f-1} \sum_{m_2=0}^{N_f-1} \sum_{l=0}^{L} \alpha_l \int_0^{T_p} p(\lambda)p(\lambda - \Delta - \tau_{l,0})d\lambda, \quad (5)$$

where the delay $\Delta := (m_2 - m_1 - kN_f + N_\epsilon - n)T_f - c_kT_c - \epsilon$. Due to the finite nonzero support of $p(t)$, and depending on n and k, only a few (m_1, m_2) pairs contribute nonzero summands to $r_{pp}(n - kN_f)$. Because $\tau_{L,0} < T_f$, those (m_1, m_2) pairs that might yield nonzero contributions must satisfy the condition: $\Delta \in (-T_f, T_p)$. Consequently, the integers m_1 and m_2 that contribute nonzero terms in (5) must satisfy:

$$m_2 - m_1 = n - kN_f - N_\epsilon - q, \; q = 0, 1, 2. \quad (6)$$

Complying to this constraint, (5) becomes:

$$r_{pp}(n - kN_f) = \sum_{q=0}^{2} A_q(n - kN_f)\beta_q(k), \quad (7)$$

where $A_q(n) := N_f - |n - N_\epsilon - q|, \forall n \in [-N_f + N_\epsilon + q, N_f + N_\epsilon + q]$, and

$$\beta_q(k) := \sum_{l=0}^{L} \alpha_l \int_0^{T_p} p(\lambda)p(\lambda + qT_f - c_kT_c - \epsilon - \tau_{l,0})d\lambda.$$

The dependence of $\beta_q(k)$ on the symbol index k comes from the TH code c_k, and from the channel parameters, when they change from symbol to symbol. Moreover, $\beta_q(k)$ also depends on ϵ implicitly, because the value of the latter affects the monocycle correlation. Notice that since the paths picked up by $\beta_q(k)$'s are far apart for different q values, the gains of those paths are independent; i.e., $\beta_q(k)$'s are independent across different q values. Substituting (7) back into (4), we have:

$$y(n) = \sqrt{\mathcal{P}} \sum_{k=-\infty}^{+\infty} s(k) \sum_{q=0}^{2} A_q(n - kN_f)\beta_q(k) + \eta(n). \quad (8)$$

Recalling that each symbol is transmitted over N_f frames, we let $n := k_fN_f + n_f$ with $k_f := \lfloor n/N_f \rfloor$, and $n_f := n - k_fN_f$. Consequently, the $A_q(n - kN_f)$ term in (8) becomes $A_q((k_f - k)N_f + n_f)$. And the constraint associated with the definition of $A_q(n)$ implies that

$$(k - k_f)N_f \in [-N_f + n_f - N_\epsilon - q, N_f + n_f - N_\epsilon - q].$$

In other words, for any given k_f, we only need to consider $k \in [k_f - 2, k_f + 1]$. As a result, (8) becomes:

$$\begin{aligned} y(n) &= \sqrt{\mathcal{P}} \sum_{m=-1}^{2} s(k_f - m) r_{pp}(n - (k_f - m)N_f) + \eta(n) \\ &= \sqrt{\mathcal{P}} \sum_{q=0}^{2} \sum_{m=-1}^{2} s(k_f - m) A_q(n_f + mN_f)\beta_q(k_f - m) + \eta(n). \end{aligned} \quad (9)$$

Intuitively speaking, $y(n)$ is regarded by the receiver as a frame-rate sample resulting from the k_fth symbol. Nevertheless, due to the timing offset, $y(n)$ could relate to any of $s(k_f - m)$, $m \in [-1, 2]$, or combinations of them. While the multiple m values include the effect of N_ϵ, it is ϵ and the TH delay that determines which paths are picked up by the correlator, and thus determines the parameters associated with q.

3.1. Periodically Time-Varying Correlation Approach

The inherent pulse repetition pattern of a UWB signal gives rise to cyclostationarity at the frame level, which can be exploited for timing acquisition. Let the time-varying correlation of a general non-stationary process $y(n)$ be $r_{yy}(n; \nu) := \mathrm{E}[y(n)y(n + \nu)]$, where ν is integer. Skipping for brevity the noise in (9), and using the independence between the random channel and the symbol sequence, we find

$$\begin{aligned} r_{yy}(n; \nu) = \mathcal{P} \sum_{m_1=-1}^{2} \sum_{m_2=-1}^{2} &\mathrm{E}[s(k_f - m_1)s(k_f^{(\nu)} - m_2)] \\ &\cdot \mathrm{E}[r_{pp}(n - (k_f - m_1)N_f) r_{pp}(n + \nu - (k_f^{(\nu)} - m_2)N_f)], \end{aligned}$$

where $k_f^{(\nu)} := \lfloor (n + \nu)/N_f \rfloor$. Notice that a nonzero contribution to $r_{yy}(n; \nu)$ results only when $k_f^{(\nu)} - m_2 = k_f - m_1$; hence,

$$\begin{aligned} r_{yy}(n; \nu) &= \mathcal{P} \sum_{m=-1}^{2} \mathrm{E}[r_{pp}(n - (k_f - m)N_f) r_{pp}(n + \nu - (k_f - m)N_f)] \\ &= \mathcal{P} \sum_{q=0}^{2} \sum_{m=-1}^{2} \mathcal{A}_{q,m}(n_f; \nu)\mathcal{B}_q, \end{aligned} \quad (10)$$

where $\mathcal{A}_{q,m}(n_f; \nu) := A_q(n_f + mN_f)A_q(n_f + \nu + mN_f)$, and $\mathcal{B}_q := \mathrm{E}[\beta_q^2(k_f - m)]$. For each $q \in [0, 2]$ and $m \in [-1, 2]$, $\mathcal{A}_{q,m}(n_f; \nu) \neq 0$ necessitates $n_f + \nu \in [-(m + 1)N_f + N_\epsilon +$

$q, -(m-1)N_f + N_\epsilon + q]$. It is worth clarifying that the expectation here is taken over not only all possible channel realizations, but also over all possible TH codes. Therefore, although $\beta_q^2(k_f - m)$ is $k_f - m$ dependent, its expectation $\mathcal{B}_q$ is not. Adding kN_f (with k integer) to n in (10) does not change the correlation, simply because $n + kN_f$ shares the same n_f with n. Therefore, $r_{yy}(n;\nu)$ is indeed periodic in n with period N_f.

Next, we will show that $r_{yy}(n;\nu)$ in (10) has a single peak $\forall q$, whose position is related to N_ϵ. Recalling the constraint associated with the definition of $A_q(n)$, we deduce that the range of n_f and ν values over which $\sum_m \mathcal{A}_{q,m}(n_f;\nu)$ might be nonzero is given by $n_f + \nu \in [-3N_f + N_\epsilon + q, 2N_f + N_\epsilon + q], \forall q$. Let us partition this range of $n_f + \nu$ values into five segments $\{seg_i\}_{i=0}^4$, each of which defined as $seg_i := [-(3-i)N_f + N_\epsilon + q, -(2-i)N_f + N_\epsilon + q]$. We have studied the change of $r_{yy}(n;\nu)$ with respect to ν (for any n_f, N_ϵ, and q), and summarize our results in the next table:

	seg_0	seg_1	seg_2	seg_3	seg_4
$N_\epsilon + q - n_f < -\frac{N_f}{2}$	$\nearrow$	$\nearrow$	$\nearrow$	$\searrow$	$\searrow$
$N_\epsilon + q - n_f \in [-\frac{N_f}{2}, 0)$	$\nearrow$	$\nearrow$	$\nearrow$	$\nearrow$	$\searrow$
$N_\epsilon + q - n_f \in [0, \frac{N_f}{2})$	$\nearrow$	$\nearrow$	$\nearrow$	$\searrow$	$\searrow$
$N_\epsilon + q - n_f > \frac{N_f}{2}$	$\nearrow$	$\nearrow$	$\searrow$	$\searrow$	$\searrow$

The table entries reveal that $r_{yy}(n;\nu)$ is segment-wise monotonic with respect to ν, $\forall n$, and peaks at either $\nu = N_\epsilon + q - n$, or, $\nu = N_\epsilon + q - n \pm N_f$. Due to lack of space, please refer to [8] for the detailed proof. Based on this observation, we can recover N_ϵ as

$$N_\epsilon := \text{round}\{[\arg\max_\nu r_{yy}(n;\nu) + n]_{N_f}\}, \tag{11}$$

where $[\cdot]_N$ performs the modulo operation with respect to integer N, and round$\{\cdot\}$ performs the rounding operation.

In a nutshell, we have established that:

Proposition 1 (Acquisition by periodic correlation) *Timing offset can be acquired by picking the peak of the periodically time-varying correlation of the frame-rate sampled correlator outputs. An estimate can be obtained based on* (11) *for each* $n_f \in [0, N_f - 1]$*, followed by averaging across* n_f*.*

Due to the aggregate effect of $q = 0, 1, 2$, for any n, the maximum of the periodic correlation $r_{yy}(n;\nu)$ will be determined by the q associated with the strongest $\mathcal{B}_q$. In other words, by observing the ν corresponding to the maximum value of $r_{yy}(n;\nu)$, the estimate of N_ϵ comes with an ambiguity since we do not have sufficient information to distinguish among $q = 0, 1, 2$. As a matter of fact, this ambiguity is induced by ϵ as well as by the TH delay $c_k T_c$. Nevertheless, the mismatching induced by this ambiguity is at most 2 frames, the effect of which on decoding performance is negligible, with N_f in the order of hundreds.

Estimation of $r_{yy}(n;\nu)$ necessitates taking into account the noise correlation. Let us temporarily assume that $w(t)$ is white with variance σ_w^2. Then the correlation of $\eta(n)$ in (9) is:

$$r_{\eta\eta}(\nu) = \begin{cases} (N_f - |\nu|)\sigma_w^2 & \text{, if } \nu \in [-N_f, N_f] \\ 0 & \text{, otherwise.} \end{cases} \tag{12}$$

Collecting noisy frame-rate samples of the correlator output over N symbols, we can estimate $r_{yy}(n;\nu)$ as follows:

$$\hat{r}_{yy}(n;\nu) = \frac{1}{N}\sum_{k=0}^{N-1} y(kN_f + n)y(kN_f + n + \nu) - r_{\eta\eta}(\nu),$$

$\forall n \in [0, N_f]$, and $\nu \in [-3N_f, 3N_f]$. The performance of $\hat{N}_\epsilon$ based on $\hat{r}_{yy}$ heavily depends on the signal-to-noise ratio (SNR). Our objective in the ensuing section will be to suppress stationary noise effects by going to the cyclic correlation domain. This is particularly important because the noise is generally colored.

3.2. Cyclic Correlation Approach

Being periodic in n, $r_{yy}(n;\nu)$ accepts a Fourier Series expansion, with so-termed cyclic correlation coefficients given by:

$$\begin{aligned} R_{yy}(l;\nu) &:= \frac{1}{N_f}\sum_{n=0}^{N_f-1} r_{yy}(n;\nu)e^{-j\frac{2\pi}{N_f}ln} \\ &= \frac{\mathcal{P}}{N_f}\sum_{q=0}^{2}\mathcal{B}_q R_q(l;\nu) + \frac{1}{N_f}\sum_{n=0}^{N_f-1} r_{\eta\eta}(\nu)e^{-j\frac{2\pi}{N_f}ln} \\ &= \frac{\mathcal{P}}{N_f}\sum_{q=0}^{2}\mathcal{B}_q R_q(l;\nu) + r_{\eta\eta}(\nu)\delta(l), \end{aligned} \tag{13}$$

where $R_q(l;\nu)$ is introduced for notational brevity, and is given by:

$$\begin{aligned} R_q(l;\nu) &:= \sum_{m=-1}^{2}\sum_{n=0}^{N_f-1}\mathcal{A}_{q,m}(n;\nu)e^{-j\frac{2\pi}{N_f}ln} \\ &= \sum_{m=-1}^{2}\sum_{n=0}^{N_f-1} A_q(n + mN_f)A_q(n + \nu + mN_f)e^{-j\frac{2\pi}{N_f}ln}. \end{aligned}$$

Changing variables and noticing that integer multiples of N_f have no effect on the exponential term, we find:

$$\begin{aligned} R_q(l;\nu) &= \sum_{m=-1}^{2}\sum_{n=-mN_f}^{(1-m)N_f-1} A_q(n)A_q(n+\nu)e^{-j\frac{2\pi}{N_f}l(n-mN_f)} \\ &= \sum_{n=-2N_f}^{2N_f-1} A_q(n)A_q(n+\nu)e^{-j\frac{2\pi}{N_f}ln}. \end{aligned}$$

Plugging in the definition of $A_q(n)$, and recalling the associated constraint, we have:

$$\begin{aligned} R_q\ (l;\nu) &= \sum_{n=-N_f+N_\epsilon+q}^{N_f+N_\epsilon+q}(N_f - |n - N_\epsilon - q|)(N_f - |n + \nu - N_\epsilon - q|)e^{-j\frac{2\pi}{N_f}ln} \\ &= e^{-j\frac{2\pi}{N_f}l(N_\epsilon+q)}\sum_{n=-N_f}^{N_f}(N_f - |n|)(N_f - |n+\nu|)e^{-j\frac{2\pi}{N_f}ln}, \end{aligned} \tag{14}$$

where change of variables was used in establishing (14). Fourier Transforming the finite sequence $N_f - |n|$ yields

$$G(f) := \sum_n (N_f - |n|)\exp(-j2\pi fn) = \frac{\sin^2(\pi f N_f)}{\sin^2(\pi f)}. \tag{15}$$

Notice that $G(f)$ is real and symmetric in f; i.e., $G(f) = G(-f)$. Using Parseval's relation, we can rewrite the sum in (14) as

$$\begin{aligned} \mathcal{S} &:= \sum_{n=-N_f}^{N_f}(N_f - |n|)(N_f - |n+\nu|)e^{-j\frac{2\pi}{N_f}ln} \\ &= \int_{-1/2}^{1/2} G\left(\lambda - \frac{l}{N_f}\right)G(\lambda)e^{j2\pi\lambda\nu}d\lambda \\ &= e^{j\frac{\pi}{N_f}l\nu}\int_{-1/2-l/(2N_f)}^{1/2-l/(2N_f)} G\left(\lambda - \frac{l}{2N_f}\right)G\left(\lambda + \frac{l}{2N_f}\right)e^{j2\pi\lambda\nu}d\lambda. \end{aligned}$$

Since $G(f)$ is symmetric, we deduce that the product $G\left(\lambda - \frac{l}{2N_f}\right)$ $G\left(\lambda + \frac{l}{2N_f}\right)$ is also symmetric. Further noticing that changing the integral limits to ± 0.5 does not alter the result, we find

$$\begin{aligned}\mathcal{S} &= e^{j\frac{\pi}{N_f}l\nu} \int_{-1/2}^{1/2} G\left(\lambda - \frac{l}{2N_f}\right) G\left(\lambda + \frac{l}{2N_f}\right) e^{j2\pi\lambda\nu} d\lambda \\ &= e^{j\frac{\pi}{N_f}l\nu} \int_{0}^{1/2} 2G\left(\lambda - \frac{l}{2N_f}\right) G\left(\lambda + \frac{l}{2N_f}\right) \cos(2\pi\lambda\nu) d\lambda \\ &= \mathcal{S}_r(\nu) e^{j\frac{\pi}{N_f}l\nu}, \end{aligned} \quad (16)$$

in which $\mathcal{S}_r(\nu)$ contains no imaginary part, and thus is guaranteed to be real. Substituting (16) and (14) into (13), we obtain:

$$R_{yy}(l;\nu) = \frac{\mathcal{P}\mathcal{S}_r(\nu)}{N_f} \sum_{q=0}^{2} \mathcal{B}_q e^{-j\frac{\pi}{N_f}l(2N_\epsilon + 2q - \nu)} + r_{\eta\eta}(\nu)\delta(l). \quad (17)$$

When $l \neq 0$, the noise term vanishes, and we can recover N_ϵ based on the phase of our cyclic correlation as

$$N_\epsilon := \text{round}\left\{ \left[\frac{1}{2}(\nu - \theta(l;\nu)\frac{N_f}{l\pi} \right]_{N_f} \right\}, \quad (18)$$

where $\theta(l;\nu) := \angle R_{yy}(l;\nu)$. As in the preceding section, N_ϵ is actually the estimate of $N_\epsilon + q$, where q corresponds to the strongest $\mathcal{B}_q$. Summarizing, we have established:

Proposition 2 (Acquisition by cyclic correlation) *Timing can be acquired from the phase of the cyclic correlation of the frame-rate sampled correlator outputs. An estimate can be obtained using* (18) *for each* $\nu \in (-N_f, N_f)$*, followed by averaging over* ν*. To avoid phase wrapping,* l *should be set to* ± 1.

In practice, $R_{yy}(l;\nu)$ can be estimated via

$$\hat{R}_{yy}(l;\nu) = \frac{1}{N} \sum_{n=0}^{N-\nu-1} y(n) y(n+\nu) e^{-j\frac{2\pi}{N_f}ln}. \quad (19)$$

Random channel effects on the estimator (18) can be further alleviated by averaging (18) over $l \in \{\pm 1\}$.

4. SIMULATIONS

We choose $p(t)$ as the second derivative of the Gaussian pulse, normalized to have unit energy, and pulse width $T_p = 0.7ns$. The system parameters are as follows: binary PAM, $N_f = 101$, $T_f = 100ns$ [7], which is also the maximum delay spread.

The channel we simulated has 400 paths equally spaced in time within the maximum delay spread; $\{\alpha_l\}_{l=0}^{L}$ are generated as Gaussian variables, linearly weighted with weights decreasing to zero at the maximum delay spread. We used a random TH code uniformly distributed over $[0, N_c - 1]$, with $N_c = 90$ and $T_c = 1ns$.

To verify our timing acquisition methods, we test the performance of the two timing acquisition methods detailed in Propositions 1 and 2. In our simulations, ϵ is randomly taken in the range $[0, T_f)$, and $N_\epsilon = 89$ and an SNR of $20dB$ is used. The normalized mean square error (MSE) versus the number of symbols N used for estimation is plotted in Fig. 4 for both estimators.

The performance of both estimators improves when averaging over more symbols, while the estimator based on Proposition 1 yields a consistently lower MSE.

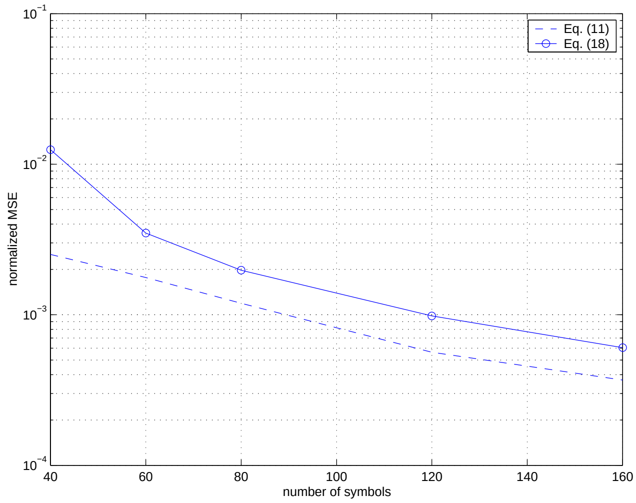

Fig. 1. Normalized MSE of $\hat{N}_\epsilon$ vs. number of symbols.

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