# BLIND UWB TIMING WITH A DIRTY TEMPLATE

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**ABSTRACT**

Ultra-Wideband (UWB) radio is gaining increasing attention thanks to its attractive features that include low-power low-complexity baseband operation and ample multipath diversity. Realization of its potential, however, faces the challenge of low-complexity high-performance timing acquisition. In this paper, we develop a blind timing acquisition algorithm for frame-level synchronization. Relying on simple integrate-and-dump operations over one symbol duration, our algorithm exploits the rich multipath diversity enabled by UWB transmissions. It outperforms existing blind algorithms and has comparable performance to data-aided ones. Equally attractive is its applicability to UWB links with or without time hopping (TH), over frequency-flat or multipath channels. It is also worth stressing that our "dirty" template based scheme is able to achieve timing synchronization at any desirable resolution and is readily applicable to non-UWB systems, so long as inter-symbol interference is absent.

## 1. INTRODUCTION

Since the recent release of the FCC spectral mask, UWB radio has been attracting growing interest especially in the area of short-range indoor wireless communications. This interest stems from its low-power low-complexity baseband operation and ample multipath diversity; see e.g., [1] and references therein. In realizing the unique benefits of UWB radios, clock synchronization constitutes a major challenge, the difficulty of which is accentuated due to the impulse-like low-power UWB transmit-waveforms.

Though straightforward, peak-picking the output of a sliding correlator with the transmit-waveform template is not only suboptimum in the presence of dense multipath, but also results in unacceptably slow acquisition speed and prohibitive complexity when one has to perform exhaustive search over thousands of bins (chips). Attempts to improve sliding-correlator based acquisition speed include a coarse bin reversal search considered in [2] over a noiseless non-modulated pulse sequence and a coded beacon sequence in conjunction with a bank of correlators [3].

Recently, non-data aided and data-aided timing acquisition and tracking schemes have been developed for UWB transmissions through dense multipath channels [4, 5]. However, [4] relies on the cyclostationarity that arises only when there is no TH within each symbol and requires dense multipath that fills up the frame. Moreover, timing acquisition performance suffers from limited multipath energy capture [4]. On the other hand, the data-aided scheme relies on a judiciously designed training pattern and exploits the rich multipath diversity provided by UWB channels [5]. But its high performance comes at the price of reduced bandwidth and/or power efficiency.

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In this paper, we develop a blind timing acquisition algorithm based on integrate-and-dump operations that collect the available multipath diversity[1]. Consequently, energy capture is considerably improved, in comparison to the non-data aided schemes in [4]. As a result, even with medium length observation intervals, timing acquisition and symbol detection performance is considerably better than existing blind algorithms, and is comparable to that of the data-aided scheme in [5].

Similar to [5], but unlike early-late gate and sliding-correlator based schemes, our algorithm does not assume a clean template at the receiver. At the price of using a noisy ("dirty") template, we gain in energy capture whether or not TH is present; and allow for not only frequency-flat, but also frequency-selective channels. Equally important, our scheme is able to achieve timing at any desirable resolution. But due to space limitation, we will focus on frame-level acquisition in this paper. Compared to [5] that utilizes training symbols, our blind scheme saves bandwidth and power, and is attractive for "cold start-up" scenarios.

Section 2 outlines our system model and transceiver operating conditions. Section 3 derives the novel blind acquisition algorithm. Simulation results and summarizing remarks are given in Sections 4 and 5, respectively.

*Notation*: $\lceil\cdot\rceil$ and $\lfloor\cdot\rfloor$ stand for integer ceiling and floor operations, respectively; $(A \bmod B)$ denotes the modulo operation, where $A$ and $B$ are both real.

## 2. MODELING AND PROBLEM STATEMENT

In UWB radio, every information symbol is conveyed by $N_f$ data modulated ultra short pulses $p(t)$, each over one frame of duration $T_f$. The resultant symbol duration is thus $T_s = N_f T_f$ seconds. With $p(t)$ having duration $T_p (\ll T_f)$ at the sub-nanosecond scale, the transmitted signal occupies UWB with bandwidth $B_s \approx 1/T_p$. UWB radio generally adopts modulation methods such as pulse position modulation (PPM), and pulse amplitude modulation (PAM). In this paper, we will deal with binary antipodal PAM, though generalization to PPM is possible.

When multiple users are present, user separation can be accomplished with pseudo-random TH codes, which shift the pulse positions at multiples of the chip duration ($T_c$) [1]. Letting $c_{n_f} \in [0, N_c - 1]$ denote the TH code during the $n_f$th frame with $N_c := \lfloor T_f/T_c \rfloor$ chips, the transmitted symbol waveform $p_T(t)$ containing $N_f$ pulses is given by $p_T(t) := \sum_{n_f=0}^{N_f-1} p(t - n_f T_f - c_{n_f} T_c)$.

The multipath channel is modeled as a tapped-delay line, with $L+1$ taps $\{\alpha_l\}_{l=0}^{L}$, and delays $\{\tau_l\}_{l=0}^{L}$ satisfying $\tau_l < \tau_{l+1}$, $\forall l$. Being quasi-static, the channel coefficients and delays remain invariant over one transmission burst, but are allowed to change across bursts. To isolate the multipath spreading effects from the

[1] This is reminiscent of pilot waveform assisted modulation [6], and transmitted reference (TR) approaches to channel estimation [7], but at the timing acquisition stage (no TR pulses are needed).

propagation delay $\tau_0$, all path delays can be uniquely casted into: $\tau_{l,0} := \tau_l - \tau_0$. Focusing on *a single user link*, and treating multi-user interference (MUI) as noise, the waveform arriving at the receiver is given by:

$$r(t) = \sqrt{\mathcal{E}} \sum_{l=0}^{L} \alpha_l \sum_{k=0}^{+\infty} s(k) p_T(t - kT_s - \tau_{l,0} - t_2) + w(t),$$

where the noise term $w(t)$ includes the MUI, and the first arrival time $t_2$ is nothing but the transmission starting time $t_1$ augmented by the propagation delay $\tau_0$.

To simplify notation, we introduce the received symbol waveform:

$$p_R(t) := \sum_{l=0}^{L} \alpha_l p_T(t - \tau_{l,0}) = \sum_{n_f=0}^{N_f-1} h(t - n_f T_f - c_{n_f} T_c), \qquad (1)$$

where $h(t) := \sum_{l=0}^{L} \alpha_l p(t - \tau_{l,0})$ is the overall channel capturing both pulse shaper and multipath effects. It follows that the received noisy signal is:

$$r(t) = \sqrt{\mathcal{E}} \sum_{k=0}^{+\infty} s(k) p_R(t - kT_s - t_2) + w(t). \qquad (2)$$

Evidently, selecting $T_f \geq \tau_{L,0} + T_p$, and $c_0 = c_{N_f-1} = 0$, the duration of $p_R(t)$ is confined over $[0, T_s)$, and inter-symbol interference is avoided.

Although the received waveform starts at $t_2 = t_1 + \tau_0$, the receiver knows neither the transmission starting time $t_1$, nor the propagation delay $\tau_0$. Upon detecting the energy/amplitude change in the arriving signals, the receiver initiates timing acquisition at $t_3$ ($> t_2$). To acquire timing, the receiver aims at the starting time of a symbol starting after, or at, $t_3$; i.e., at time $t = t_3$, if $(t_3 - t_2)$ is an integer multiple of $T_s$, or, $t = t_2 + \lceil (t_3 - t_2)/T_s \rceil T_s$, otherwise. Notice that $t_1$ plays no role, while the arrival time $t_2$ serves as reference. Without loss of generality, we set $t_2 = 0$. Further, defining $N := \lfloor t_3/T_s \rfloor$ and $\epsilon := t_3 - NT_s$, timing acquisition amounts to correcting $t_3$ by an integer number $n_\epsilon \in [0, N_f - 1]$ of frames, such that $t_3 + n_\epsilon T_f$ is close to either $NT_s$, or, $(N+1)T_s$, within ambiguity of $T_f$ seconds.

In the next section, we develop a low complexity blind synchronization algorithm that relies on integrate-and-dump operations, one per symbol duration, to acquire timing with frame-level resolution.

## 3. BLIND TIMING ACQUISITION

The first step of our blind timing acquisition algorithm is to take from the received waveform a segment of duration $T_s$, starting at time $(t_3 + nT_f + mT_s)$, for integers $n \in [0, N_f)$, and $m \in [0, M-1]$, with $MT_s$ being the observation interval. Denoted by $x_{n,m}(t)$, this waveform can be expressed as:

$$x_{n,m}(t) = r(t + mT_s + nT_f + t_3), \; t \in [0, T_s). \qquad (3)$$

Ignoring noise for brevity, and using $t_3 = NT_s + \epsilon$, where $\epsilon \in [0, T_s)$ by definition, we can express the noise-free $x_{n,m}(t)$ as:

$$x_{n,m}(t) = \sqrt{\mathcal{E}} \sum_{k=0}^{+\infty} s(k) p_R(t - \mathcal{D}_{k,n,m}), \;\; \forall t \in [0, T_s) \qquad (4)$$

where the aggregate delay is given by $\mathcal{D}_{k,n,m} := (k - N - m)T_s - (nT_f + \epsilon)$.

Since $p_R(t)$ has finite support $(0, T_s)$, only a finite number of $k$ values contribute non-zero summands in (4), for any given $m, n$, and $t_3$. To find such $k$ values, let us first express $(nT_f + \epsilon)$ as an integer multiple of $T_s$ plus a residue

$$n_s := \lfloor (\epsilon + nT_f)/T_s \rfloor, \text{ and } \epsilon_s := \epsilon + nT_f - n_s T_s, \qquad (5)$$

where possible $n_s$ values are 0 and 1, and $\epsilon_s \in [0, T_s)$. With this notation, we deduce that the $k$ values that contribute non-zero summands in (4) are as follows:

$$k = N + m + n_s + [0, 1]. \qquad (6)$$

Recalling that $k$ denotes symbol index, (6) indicates nothing but the indices of the symbols that correspond to (possibly) non-zero coefficients in (4). In other words, all other symbols have zero coefficients and thus contribute zero summands in (4). Consequently, (4) becomes:

$$x_{n,m}(t) = \sqrt{\mathcal{E}} \sum_{k=0}^{1} s(k + N + m + n_s) p_R(t - kT_s + \epsilon_s). \qquad (7)$$

Notice that for fixed $n$ and $\epsilon$, $\epsilon_s$ is uniquely determined. Moreover, for any fixed $k$, $p_R(t - kT_s + \epsilon_s)$ depends only on $\epsilon_s$. As a result, as $m$ changes, $p_R(t - kT_s + \epsilon_s)$ corresponding to each $k$ stays invariant for any given $n$ and $\epsilon$.

As outlined in the introduction, we will estimate $n_\epsilon$ relying the cross correlation between successively observed waveforms $x_{n,m}(t)$ of duration $T_s$ each. Integrating-and-dumping, the product of adjacent waveforms $x_{n,m}(t)$ and $x_{n,m+1}(t)$ we find:

$$\begin{aligned} R_{xx}(n; m) &:= \int_0^{T_s} x_{n,m}(t) x_{n,m+1}(t) dt \\ &= \mathcal{E} \int_0^{T_s} \sum_{k=0}^{1} s(k + N + m + n_s) p_R(t - kT_s + \epsilon_s) \\ &\times \sum_{k=0}^{1} s(k + N + m + n_s + 1) p_R(t - kT_s + \epsilon_s) dt \\ &= \mathcal{E} \cdot s(N + m + n_s + 1) \left[ s(N + m + n_s) \int_{\epsilon_s}^{T_s} p_R^2(t) dt \right. \\ &\left. + s(N + m + n_s + 2) \int_0^{\epsilon_s} p_R^2(t) dt \right], \end{aligned} \qquad (8)$$

where in establishing the last equality, we used the fact that $\int_0^{T_s} p_R (t + \epsilon_s) p_R(t + T_s + \epsilon_s) dt$ vanishes due to the finite support of $p_R(t)$, and substitutions $\int_{\epsilon_s}^{T_s} p_R^2(t) dt = \int_0^{T_s} p_R^2(t + \epsilon_s) dt$, and $\int_0^{\epsilon_s} p_R^2(t) dt = \int_0^{T_s} p_R^2(t - T_s + \epsilon_s) dt$. Taking absolute value of $R_{xx}(n; m)$ removes the dependence on the unknown symbol $s(N + m + n_s + 1)$, and yields:

$$\begin{aligned} |R_{xx}(n; m)| = \mathcal{E} \cdot \Big| s(N + m + n_s) \int_{\epsilon_s}^{T_s} p_R^2(t) dt \\ + s(N + m + n_s + 2) \int_0^{\epsilon_s} p_R^2(t) dt \Big|. \end{aligned}$$

Moreover, with information conveying symbols $s(k) \in \{\pm 1\}$ being independent, taking conditional expectation of the absolute value $|R_{xx}(n; m)|$ for any $n \in [0, N_f - 1]$, gives rise to:

$$\begin{aligned} \mathcal{R}_{xx}(n) &:= \mathrm{E}\{|R_{xx}(n; m)| \;|\; h(t), \epsilon\} \\ &= \frac{\mathcal{E}}{2} \left| \int_{\epsilon_s}^{T_s} p_R^2(t) dt - \int_0^{\epsilon_s} p_R^2(t) dt \right| + \frac{\mathcal{E}}{2} \int_0^{T_s} p_R^2(t) dt. \end{aligned} \qquad (9)$$

Notice that the expectation removes the dependence of the absolute value $|R_{xx}(n;m)|$ on $N$, $m$, and $n_s$. But as we will show next, $\mathcal{R}_{xx}(n)$ is still dependent on $n$, for any timing offset $\epsilon$. In fact, it is this dependence that enables us to acquire timing information.

Recalling that the symbol waveform $p_R(t)$ contains $N_f$ pulses, we infer that the first integral corresponds to pulses with indices from $\lfloor \epsilon_s/T_f \rfloor$ to $N_f - 1$, while the second integral corresponds to pulses with indices from 0 to $\lfloor \epsilon_s/T_f \rfloor - 1$. In the absence of a TH code, i.e., $c_{n_f} = 0$, $\forall n_f \in [0, N_f - 1]$, the two integrals in (9) can be respectively expressed (in closed-form) as:

$$\begin{aligned} \int_{\epsilon_s}^{T_s} p_R^2(t)dt &= \left(N_f - \left\lfloor \frac{\epsilon_s}{T_f} \right\rfloor\right) \mathcal{E}_h(T_f) - \mathcal{E}_h(\epsilon_f), \\ \int_{0}^{\epsilon_s} p_R^2(t)dt &= \left\lfloor \frac{\epsilon_s}{T_f} \right\rfloor \mathcal{E}_h(T_f) + \mathcal{E}_h(\epsilon_f), \end{aligned} \tag{10}$$

where $\mathcal{E}_h(\tau) := \int_0^\tau h^2(t)dt$, and $\epsilon_f := (\epsilon_s \bmod T_f)$. From the definition of $\epsilon_s$, it follows that $\epsilon_f = (\epsilon \bmod T_f)$. In the presence of a TH code, the two integrals cannot be expressed in such a neat form. But as shown in [5], they do not vary much (with or without TH), for various $\epsilon_s$ values and channel realizations.

Substituting (10) into (9) results in:

$$\begin{aligned} \mathcal{R}_{xx}(n) &= \frac{\mathcal{E}}{2}\left|\left(N_f - 2\left\lfloor \frac{\epsilon_s}{T_f} \right\rfloor\right) \mathcal{E}_h(T_f) - 2\mathcal{E}_h(\epsilon_f)\right| \\ &+ \frac{\mathcal{E}}{2} N_f \mathcal{E}_h(T_f). \end{aligned} \tag{11}$$

Since $\epsilon_s \in [0, T_s)$ by definition, we deduce that $\lfloor \epsilon_s/T_f \rfloor$ falls in the range $[0, N_f - 1]$. Consequently, it can be readily verified that the maximum of $\mathcal{R}_{xx}(n)$ in (9) is as follows:

$$\begin{aligned} &\max_n \{\mathcal{R}_{xx}(n)\} \\ &= \mathcal{E} N_f \mathcal{E}_h(T_f) - \mathcal{E} \min\{\mathcal{E}_h(\epsilon_f), \mathcal{E}_h(T_f) - \mathcal{E}_h(\epsilon_f)\}. \end{aligned} \tag{12}$$

More specifically, the maximum occurs at $\lfloor \epsilon_s/T_f \rfloor = 0$ if $\mathcal{E}_h(\epsilon_f) < \mathcal{E}_h(T_f)/2$, and at $\lfloor \epsilon_s/T_f \rfloor = N_f - 1$ otherwise. Recalling the definition of $\epsilon_s$ and noticing that $n \in [0, N_f - 1]$, we have the following result: when $\epsilon = 0$, $\arg\max_n\{\mathcal{R}_{xx}(n)\} = -\lfloor \epsilon/T_f \rfloor = 0$; and when $\epsilon > 0$,

$$\begin{aligned} &\arg\max_n \{\mathcal{R}_{xx}(n)\} \\ &= \begin{cases} ((N_f - \lfloor \epsilon/T_f \rfloor) \bmod N_f), & \text{if } \mathcal{E}_h(\epsilon_f) < \mathcal{E}_h(T_f)/2, \\ N_f - \lfloor \epsilon/T_f \rfloor - 1, & \text{otherwise.} \end{cases} \end{aligned} \tag{13}$$

Eq. (13) reveals that for any $\epsilon$, a unique $\arg\max_n\{\mathcal{R}_{xx}(n)\}$ can be found by peak picking $\mathcal{R}_{xx}(n)$ across $n$. In fact, $\arg\max_n \{\mathcal{R}_{xx}(n)\}$ is nothing but $n_\epsilon$, simply because by correcting $t_3$ with $\arg\max_n\{\mathcal{R}_{xx}(n)\}$ frames, we can acquire either $NT_s$, or, $(N+1)T_s$, with ambiguity $< T_f$. To illustrate this point more clearly, we list possible values of $t_3 + \arg\max_n\{\mathcal{R}_{xx}(n)\}T_f$ in Table 1. Evidently, both $\epsilon_f$ and $T_f - \epsilon_f$ are strictly less than $T_f$, by definition. We have thus established that:

**Proposition 1** *Timing acquisition amounts to find $n_\epsilon$, by picking the peak of $\mathcal{R}_{xx}(n)$ in* (9) *across $n \in [0, N_f - 1]$. Correcting the starting time of reception $t_3$ with $n_\epsilon = \arg\max_n\{\mathcal{R}_{xx}(n)\}$ obtained from* (13) *yields the desired timing $NT_s$, or, $(N+1)T_s$, with ambiguity $< T_f$.*

**Table 1**. Possible values of $t_3 + \arg\max_n\{\mathcal{R}_{xx}(n)\}T_f$

| | $\mathcal{E}_h(\epsilon_f) < \frac{\mathcal{E}_h(T_f)}{2}$ | $\mathcal{E}_h(\epsilon_f) > \frac{\mathcal{E}_h(T_f)}{2}$ |
|---|---|---|
| $\epsilon = 0$ | $NT_s$ | impossible |
| $\epsilon \in (0, T_f)$ | $NT_s + \epsilon_f$ | $(N+1)T_s + (\epsilon_f - T_f)$ |
| $\epsilon \in [T_f, T_s)$ | $(N+1)T_s + \epsilon_f$ | $(N+1)T_s + (\epsilon_f - T_f)$ |

An estimate of $\mathcal{R}_{xx}(n)$ can be obtained by computing $\mathcal{R}_{xx}(n;m)$ over pairs of $x_{n,m}(t)$ each of duration $T_s$, and averaging their absolute values across the $M/2$ pairs, as follows:

$$\hat{\mathcal{R}}_{xx}(n) = \frac{2}{M} \sum_{m=0}^{M/2-1} \left| \int_0^{T_s} x_{n,2m}(t) x_{n,2m+1}(t) dt \right|. \tag{14}$$

Summarizing, our *blind timing algorithm* based on "dirty templates" can be carried out as follows:

**Step 0.** Set $n = 0$.
**Step 1.** For a given $n$, take $M$ segments $x_{n,m}(t)$ each of duration $T_s$ from the received signal as in (3). Integrate-and-dump the product of adjacent segments $x_{n,m}(t)$ and $x_{n,m+1}(t)$ as in (8).
**Step 2.** Form an estimate of $\mathcal{R}_{xx}(n)$ by averaging over all pairs the absolute value of the integral obtained in Step 1 [c.f. (14)]. If $n < N_f - 1$, set $n = n + 1$, and go to Step 1. Otherwise, go to Step 3.
**Step 3.** Find an estimate of $n_\epsilon$ by peak-picking $\hat{\mathcal{R}}_{xx}(n)$; i.e., $\hat{n_\epsilon} = \arg\max_n\{\hat{\mathcal{R}}_{xx}(n)\}$.
**Remark:** Correcting $t_3$ with $\hat{n_\epsilon}T_f$ will give rise to an estimate of a symbol starting time. Notice that instead of having a fixed increment 1 (frame duration) per iteration, our algorithm can be applied also with variable non-integer increments using voltage controlled clock (VCC) circuits. The latter then enables not only acquisition, but also tracking, with the possibility of further reduction of synchronization speed.

## 4. SIMULATIONS

In this section, preliminary simulation results will be presented. We select the pulse $p(t)$ as the second derivative of the Gaussian function with unit energy and duration $T_p \approx 1$ ns. Each symbol contains $N_f = 32$ frames, each with duration $T_f = 100$ns, as in [1]. The random channels are generated according to the model in [8, 9] with $(1/\Lambda, 1/\lambda, \Gamma, \gamma) = (2, 0.5, 30, 5)$ns. The diminishing tail of the power profile is truncated to make the maximum delay spread of the multipath channel $T_g = 99$ns.

The timing acquisition and symbol detection performance of our blind algorithm is tested for various $M$ values. Without loss of generality, we set $N = 0$, and generate $\epsilon$ randomly from a uniform distribution over $[0, T_s)$. We employ fast TH spreading codes of period $N_f$, which is generated from a uniform distribution over $[0, N_c - 1]$, with $N_c = 90$, and $T_c = 1.0$ns, independently from frame to frame. We also set $c_0 = c_{N_f - 1} = 0$ to avoid inter-symbol interference.

First, we test the Mean Square Error (MSE) of our timing acquisition algorithm summarized in Proposition 1. The MSE is normalized with respect to $T_s^2$, and plotted versus SNR in Figure 1, for $M = 4, 8, 16, 32, 64, 128$. Also plotted is the normalized variance of the random timing offset without timing acquisition. As

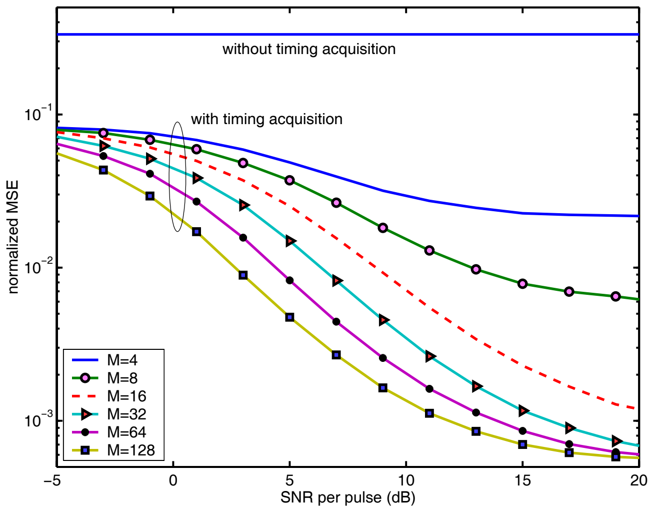


**Fig. 1**. Normalized timing acquisition MSE vs. SNR per pulse with $M = (4, 8, 16, 32, 64, 128)$.

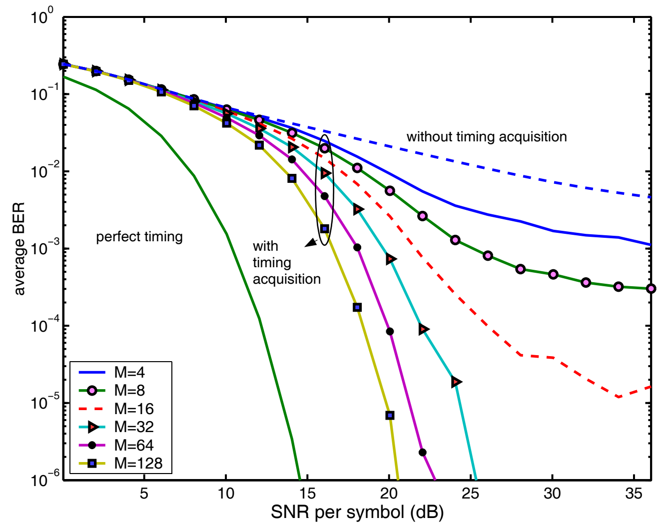


**Fig. 2**. Averaged symbol detection BER vs. SNR per symbol with $M = (4, 8, 16, 32, 64, 128)$.

$M$ increases, the normalized MSE decreases monotonically. The same trend can be observed for increasing SNR. For all $M$ values, the MSE curves flatten at high SNR. This is induced mainly by the random symbol averaging for small $M$ values, and by the frame resolution limit for large $M$ values. The latter can be mitigated through tracking.

We then test the bit-error-rate (BER) performance associated with our timing acquisition scheme, and compare it to the cases without timing acquisition, and with perfect timing. To isolate timing from channel estimation errors, we assume that the channel estimate is error-free. Fig. 2 depicts BER performance with perfect timing, without timing, and with timing acquisition, for various $M$ values. Again, as $M$ increases, the BER performance improves monotonically.

We have also carried out MSE and BER comparisons with the non-data aided and data-aided timing acquisition methods in [4] and [5]. Simulations show significant improvement over the former, thanks to the multipath energy collection through integrate-and-dump operations even with our "dirty template" obtained from the "neighbor." Our blind algorithm with $M \geq 16$ results in MSE and BER performance similar to the data-aided algorithm in [5] using $M/4$ training symbols. Due to lack of space, these figures are not included.

## 5. CONCLUSIONS

We developed a blind timing algorithm for UWB radios based on integrate-and-dump operations between adjacent symbol-long segments of the received waveform. Segments of such pair serve as "dirty templates" for each other. The resultant timing algorithm exploits the ample multipath diversity inherent to UWB transmissions, without knowledge of either information symbols, or, the channel. Simulations and comparisons confirm considerable improvement in both error performance and acquisition speed, over existing blind algorithms. Equally important is that it provides a bandwidth efficient timing method to UWB links with or without TH, over frequency-flat or frequency-selective multipath channels. It is also worth mentioning that our algorithm is readily applicable to non-UWB systems, when inter-symbol interference is absent.

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