X-law detection scheme for monobit transmitted-reference ultra wideband receivers

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ABSTRACT

In ultra-wideband (UWB) communications, monobit receivers offer a low complexity implementation but at the same time exhibit a great performance loss. In this paper, a novel detection scheme, denoted as x-law detection (XLD), is proposed to diminish the performance loss caused by employing monobit analog-to-digital converters in transmitted-reference (TR) UWB receivers. Simulation results show that if the optimal value is employed for \( x \), the XLD-based monobit weighted TR (MWTR) receiver can achieve 14.2–15.5 dB and 8–9.2 dB performance gain over the conventional MWTR receiver in LOS and NLOS scenarios, respectively. Moreover, the XLD-based MWTR receiver performance with the optimal value of \( x \) is only 1.6–3 dB away from the optimum MWTR receiver performance in intra-vehicle UWB channels. Additionally, the XLD-based MWTR receiver is not sensitive to the summation interval. This feature decreases the receiver complexity and guarantees a robust performance over different multipath channels. The significant performance improvement of the XLD scheme comes at a limited complexity increase. Thus, the XLD approach is a good candidate for TR-based and other training-based monobit receivers requiring low complexity, high performance, and low power consumption.

1. INTRODUCTION

Impulse-radio (IR) ultra-wideband (UWB) systems use ultra-short pulses to transmit information. Since such pulses are carrierless and have large bandwidth, IR-UWB systems have the benefits of low complexity [1], high rate [2], low power consumption [1], good material penetration for medical imaging and ground-penetrating radars [3], and fine time-domain resolution suited for tracking or localization applications [4]. These features make the UWB technology a good candidate to be employed in wireless body area networks [2], [5], internet of things, and wireless sensor networks [6]-[8].

The received IR-UWB signals from multipath channels contain a large number of multipath components (MPCs) due to transmitting short-duration pulses [9]. Estimating and tracking the large number of MPCs leads to an unaffordable high complexity [10], [11]. To harvest more energy from the MPCs and obtain full multipath diversity with affordable complexity, noncoherent receivers have been presented such as transmitted-reference (TR)-based receiver [12]-[16]. However, the TR receivers require ultra-wide bandwidth analog delay lines to delay the received signal. Therefore, the implementation of TR-based receivers with integrated circuits becomes challenging. Digitizing the received IR-UWB signals is a simple approach. Nevertheless, because a sampling rate of several Giga samples per second is needed, employing high-
resolution analog-to-digital converters (ADCs) becomes challenging due to unaffordable complexity and power consumption. Thus, using finite-resolution ADCs, especially monobit ones, has been proposed by [17], [18]. Implemented with a few comparators, finite-resolution ADCs have much lower complexity and power consumption than high-resolution ones [18]. However, employing finite-resolution ADCs also leads to severe performance loss for TR-based receivers [17]. In other words, monobit receivers offer a lower complexity implementation but at the same time exhibit a great performance loss. The performance loss of employing the monobit receivers is the problem that this paper aims to solve.

To alleviate the performance loss caused by monobit ADCs, a signal processing scheme, denoted as polarity-invariant square law (PISL), has been proposed in [19] for monobit weighted TR (WTR) receivers as a case study. In [19], it is shown that the PISL-based monobit WTR receiver is insensitive to correlation interval and can achieve a bit-error-rate (BER) performance that outperforms the conventional monobit receiver by more than 10 dB and is less than 1.5 dB away from the performance bound of the optimum monobit TR receivers in intra-vehicle UWB channels. In PISL technology, first, a cleaner reference pulse is generated by coherently averaging multiple reference pulses. Next, the averaged reference pulse is squared while its polarity is preserved to compensate for the monobit digitization which is a severe nonlinear process. Finally, the non-linearly processed reference pulse forms the combining weights to recover information bits.

Inspired by the concept of PISL technology, we propose a digital signal processing scheme, denoted as x-law detection (XLD), for TR-based receivers, to achieve a better performance compared to the PISL scheme. In this paper, the proposed scheme is applied to the WTR receiver as a case study. More specifically, first, by adopting a successive reference enhancement (SRE) technique several reference pulses are coherently added together to form an accumulated reference pulse with a higher signal-to-noise ratio (SNR). Then, the absolute value of the accumulated reference pulse amplitude raised to the power of x while its sign is maintained. Finally, the processed reference pulse is correlated by data pulses to retrieve information bits. In this paper, the BER performance of the XLD scheme is thoroughly investigated by applying it to the WTR receivers in intra-vehicle multipath channels for x ranging from 1 to 4 with a step of 0.5. As shown in [20], intra-vehicle channels have lots of MPCs, and large multipath delay spread. It was shown in [14], [15] that the WTR receiver achieves a higher performance and data-rate in dense multipath channels. Simulation results show that the XLD approach considerably improves the BER performance of the monobit WTR receivers. For simplicity, we have considered a single-user scenario in all the simulations. However, any multiple access scheme like direct-sequence (DS), frequency-hopping (FH), or hybrid DS/FH as in [21] can be applied to the proposed system and the capacity of the proposed system can be derived by employing the approach given in [21].

Our contributions are three-fold: i) a detection process which is a generalization of the PISL technology, denoted as XLD scheme, has been proposed to achieve a better performance compared to the PISL scheme; ii) the effect of the forgetting factor on the optimum value of x is investigated; and iii) the performance of the MWTR receivers has been studied over field measured intra-vehicle channels that have more resolvable MPCs and larger delay spread than typical IEEE 802.15.4a channels [22] for line-of-sight (LOS) and non-LOS (NLOS) scenarios. The paper structure is as follows: Section II models the MWTR receivers and introduces the concept of XLD technology. By using the performance of the optimal MWTR receiver as a benchmark, Section III compares the BER performance of the conventional and the XLD-based MWTR receivers and their sensitivity to the forgetting factor, through simulations. Finally, concluding remarks are given in Section IV.

2. MODEL OF MWTR RECEIVERS

2.1. WTR receiver structure

The transmitted signal in an uncoded WTR system can be written as (1).

\[
(t) = \sum_{j=-\infty}^{\infty} \left( \alpha w_{tr}(t - jT_B) + \beta b_{i,j} w_{tr}(t - jT_B - iT_z - T_d) \right) s_{tr}
\]

Where \(w_{tr}(t)\) denotes a baseband UWB pulse with bandwidth \(B\) and a duration \(T_{wu}\). \(T_z\) is the time interval between two successive data pulses, \(T_g\) indicates the guard time between the reference pulse and the previous/subsequent data pulses (see Figure 1). \(T_d = 2T_g + (N_s - 1)T_z\) indicates the block duration, and \(b_{i,j} \in \{-1, +1\}\) is the \(i^{th}\) binary-phase-shift-keying (BPSK) modulated bit of the \(j^{th}\) block. Finally, \(\alpha\) and \(\beta\) denote the weights of the reference pulse and data pulses in the uncoded WTR system, respectively. To allocate more power to the reference pulse than to the data pulses, we have \(\alpha \geq \beta\). We set \(\alpha^2 + N_s\beta^2 = N_s\) to normalize the transmitted energy per information bit in the uncoded system.
The structure of a monobit WTR (MWTR) receiver equipped with a channel decoder unit is shown in Figure 2. The received signal is expressed by

\[ r(t) = s_{tr}(t) * C(t) + n(t), \]

where * denotes the linear convolution, \( n(t) \) is the additive white Gaussian noise (AWGN) with two-sided power spectral density \( \frac{N_0}{2} \) and \( C(t) = \sum_{k=0}^{K-1} \alpha_k \delta(t - \tau_k) \) is the impulse response of the multipath channel in which \( K \) denotes the number of MPCs; \( \alpha_k \) and \( \tau_k \) indicate the attenuation and delay of the \( k \)th path. The received signal is fed to a low-pass filter (LPF) with an impulse response \( f(t) \) and a bandwidth \( B \) to eliminate out-of-band noise and interferences. Hence, the signal at the output of the LPF can be written as (2).

\[ r(t) = \sum_{j=0}^{\infty} a g(t - jT_B) + \beta \sum_{l=0}^{n_s-1} b_{l,j} g(t - jT_B - lT_s - T_d) \] + \( v(t) = s(t) + v(t) \)  

(2)

Where \( v(t) \) is the filtered AWGN and \( g(t) = w_{tr}(t) * C(t) * f(t) \) is the noise-free received signal resulted from transmitting \( w_{tr}(t) \). Next, \( r(t) \) is sampled at the Nyquist rate and then is quantized by a monobit ADC as (3).

\[ r[n] = \begin{cases} +1, & r(nT) \geq 0 \\ -1, & r(nT) < 0 \end{cases} \]

(3)

Finally, the quantized signal, \( r[n] \), is passed through a digital signal processing (DSP) unit for detection [13]. Let \( T_d = n_d T, T_s = n_s T, \) and \( T_B = n_B T \), where \( n_d, n_s, \) and \( n_B \) are integers.

According to Figure 3, to detect the \( k \)th bit, the reference pulse is delayed by \( kn_s + n_d \) samples and then is multiplied by the \( k \)th data pulse. Next, the result of the multiplication is summed from the \( (kn_s + n_d) \)th sample to the \( (kn_s + n_d + n_t - 1) \)th sample, where \( n_t \) is called the summation interval. Thus, the output of the summation for the \( k \)th information bit of the \( l \)th block can be expressed as (4).

\[ X_{-k}\text{law detection scheme for monobit transmitted-reference ultra wideband receivers (Hassan Khani)} \]
\[ z[k, l] = \sum_{n=0}^{n_l-1} w_{n,k,l} r[n + kn_s + n_d + ln_b], \]  

Where \( w_{n,k,l} \) denotes the combining weight for the \((n + kn_s + n_d + ln_b)^{th}\) data pulse sample that is constructed based on the reference pulses. Finally, the \( k^{th} \) bit of the \( l^{th} \) block is detected as \( \hat{b}_{k,l} \). 

\[ \hat{b}_{k,l} = \text{sign}(z[k, l]) \]  

### 2.2. Successive reference enhancement (SRE) technology

A straightforward approach to estimate \( w_{n,k,l} \), denoted as direct reference (DR) scheme, is obtained by using the reference pulse samples in the same block as the combining weights [17], i.e., 

\[ w_{n,k,l}^{DR} = r[n + ln_b] = \text{sign}(ag(nT) + v((n + ln_b)T)), \quad \text{for} \ n = 0, 1, \ldots, n_l - 1 \]  

This scenario is depicted in Figure 3 when the switch is at position 1. Since monobit digitization imposes severe quantization noise to \( r[n+ln_b] \), the performance of the DR algorithm is far worse than that of the receivers with full-resolution ADCs [15]. An improved algorithm to estimate \( w_{n,k,l} \), denoted as successive reference enhancement (SRE), is obtained by coherently accumulating the synchronized reference pulse samples in different blocks to generate the combining weights [19]. In this case, the accumulated weights can be mathematically expressed as \( (7) \). 

\[ w_{n,k,l}^{SRE} = (1 - q)w_{n,k,l-1}^{SRE} + qr[n + ln_b] = q \sum_{m=-\infty}^{l} (1 - q)^{l-m}r[n + mn_b] \]  

Where a forgetting factor, \( 0 < q < 1 \), is employed to control the effective number of blocks forming the accumulated reference pulse, a forgetting factor \( q \) is used, where \( 0 < q < 1 \). Employing forgetting factor exponentially reduces the portion of the previous blocks in the formation of the accumulated reference pulse as time passes, i.e., the older reference pulse is attenuated more compared with the newer one. Thus, the forgetting-factor based approach can easily adjust the averaging length by the value of \( q \) instead of modifying the receiver hardware compared to the conventional averaging approach used in [18]. This adjustment can be implemented by controlling the value of \( q \) based on the cross-correlation of the accumulated reference pulse and the incoming reference pulse, i.e., \( q \) increases as the cross-correlation decreases and vice versa. Thus, it lets the MWTR receivers adaptively control the number of reference pulses that are to be averaged based on the synchronization accuracy of the receiver and the coherence time of UWB channels. 

Since \( v(nT) \) is a Gaussian noise with a variance of \( \sigma^2 = N_0W \), by using \( (6) \) the probability mass function of \( r[n+ln_b] \) can be given as \( (8) \). 

\[ P(r[n + ln_b]) = \begin{cases} 1 - \varepsilon_n, & r[n + ln_b] = +1 \\ \varepsilon_n, & r[n + ln_b] = -1 \end{cases} \]  

where 

\[ \varepsilon_n = Q(\frac{ag[n]}{\sigma}) \]  

in which \( g[n]=g(nT)/\sigma \) and \( Q(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{+\infty} e^{-t^2/2} dt \). Therefore, the mean and variance of \( w_{n,k,l}^{SRE} \) can be derived as \( (10) \) and \( (11) \). 

\[ \mu_{SRE}[n] = E\{w_{n,k,l}^{SRE}\} = \sum_{m=-\infty}^{l} (1 - q)^{l-m}E\{r[n + mn_b]\} = 1 - 2\varepsilon_n, \]  

\[ \sigma_{SRE}^2[n] = E\{(w_{n,k,l}^{SRE})^2\} - \mu_{SRE}[n]^2 = \frac{q}{2-q}(1 - [1 - 2\varepsilon_n]^2) \]  

Thus, unlike \( w_{n,k,l}^{DR} \) that can only take values of 1 or -1, the polarity of \( w_{n,k,l}^{SRE} \) is the same as that of \( g(nT) \) and the amplitude of \( w_{n,k,l}^{SRE} \) increases with the amplitude of \( g(nT) \). Therefore, if SNR in \( r[n+ln_b] \) increases, the amplitude of \( w_{n,k,l}^{SRE} \) increases. It indicates that the SRE algorithm moderately compensates for the SNR loss of the monobit-quantization in the reference pulse, and subsequently achieves a better performance than the DR algorithm.
3. PROPOSED X-LAW DETECTION (XLD) ALGORITHM

For a WTR receiver with a full-resolution ADC, both reference and data pulses are proportional to 
\(g(nT)\), so the correlation between the reference and data pulses is proportional to \(g^2(nT)\). However, for an MWTR receiver, even when the SRE technology is applied, the correlation between the accumulated reference pulse and data pulses is nearly proportional to \(g(nT)\) (the accumulated reference pulse is based on \(w_{n,k,l}^{SRE}\), which is almost similar to \(g(nT)\) because the data samples are +1 or -1). To compensate for the SNR loss in data pulses and reduce the effect of low-SNR (noise-dominant) samples from the decision statistic, PISL process has been proposed. In this section, we propose x-law detection (XLD) algorithm to obtain a better performance compared to the PISL scheme. In the XLD algorithm \(w_{n,k,l}^{SRE}\) is further processed by the XLD technology as (12).

\[
w_{n,k,l}^{XLD} = sign(w_{n,k,l}^{SRE})|w_{n,k,l}^{SRE}|^x
\]  

(12)

Where \(|z|\) represents the absolute value of \(z\). It is seen from (12) that the XLD technology is equivalent to the PISL technology for \(x=2\). Therefore, XLD technology can be considered as a generalization to the PISL technology. According to (10) and (11), when the value of \(q\) is close to zero, \(w_{n,k,l}^{XLD}\) converges to the value (13).

\[
\lim_{q \to 0} w_{n,k,l}^{XLD} = sign(1 - 2\varepsilon_n)|1 - 2\varepsilon_n|^x,
\]  

(13)

and the correlation for the \(k^{th}\) bit of the \(l^{th}\) block is revised as (14).

\[
z[k, l] = \sum_{n=0}^{n_t-1} w_{n,k,l}^{XLD}[n + kn_s + n_d + ln_b]
\]  

(14)

based on the analysis given in [23, 24], the optimal combining weight for the MWTR receiver is obtained as (15).

\[
w_{n,k,l}^{OP} = \log[(1 - \zeta_n)/\zeta_n]
\]  

(15)

where

\[
\zeta_n = Q(\beta g(nT)/\sigma)
\]  

(16)

Hence, obtaining \(w_{n,k,l}^{OP}\) requires the accurate value of \(g(nT)/\sigma\), which is hardly obtained for a low-power and low-complexity receiver. The optimal combining weights given in (15) are used as the benchmark for suboptimal weights of the XLD technology. Figure 4 shows the optimum combining weights and the expected values of combining weights of the XLD technology for \(x=0, 1, 2, 3, 4\) as a function of \(g(nT)/\sigma\) when \(q\) is close to zero. For a fair comparison, the combining weights are normalized so that when \(g(nT)/\sigma = 1\) (moderate SNR region), the combining weights of different schemes have the same value. From Figure 4, we can conclude that as \(x\) increases, \(w_{n,k,l}^{XLD}\) adds more nonlinearity to its amplitude. More specifically, for noise-dominant samples (i.e., for \(g[n]/\sigma << 1\)), as \(x\) increases \(w_{n,k,l}^{XLD}\) offers a smaller amplitude, and for signal-dominant samples, \(w_{n,k,l}^{XLD}\) offers a larger amplitude. This nonlinear property can effectively lower the contribution of the noise-dominant samples and enlarge the contribution of the signal-dominant samples in the correlation output, and hence further improve the performance of the MWTR receiver. One can suppress noise-dominant samples while retaining the reference pulse shape by the use of hard nonlinearity like dead zone or simply setting to zero the samples below some threshold. This version would be computationally simpler than the proposed piecewise polynomial nonlinearity. However, simulations show that the BER performance of this scheme is slightly better than that of the XLD for \(x=1\) and lies between the performance of the XLD for \(x=1\) and \(x=1.5\).

To compare \(w_{n,k,l}^{XLD}\) with the optimum combining weights, \(w_{n,k,l}^{OP}\), the correlation coefficient between \(w_{n,k,l}^{XLD}\) and \(w_{n,k,l}^{OP}\) is shown in Figure 5 for \(q=0.2, 0.15, 0.1, 0.05, 0.025, 0.001\) when \(x\) varies from 0 to 10. Fig. 5 has been plotted by using (10) and (11). From Fig. 5, it is observed that the optimum value of \(x\) (exponent) increases as \(q\) decreases and for the extreme case reaches to 4 when \(q=0.001\). Therefore, the performance improvement brought by the XLD technology is limited and does not increase unlimitedly as \(x\) increases.
Note that the substantial performance improvement of the XLD technology comes at the expense of a limited increase in the receiver complexity. First, for an appropriate value for \( q \), the multiplication of \((1-q)w[n,k,l]_{SR}\) and \( q w[n,k,l]_{XLD}\) in (7) can be realized by binary shifts and additions. For example, if \( q=0.125 \), \((1-q)w[n,k,l]_{SR} - w[n,k,l]_{XLD}/8 \) and \( qw[n,k,l]_{XLD} = w[n,k,l]_{XLD}/8 \) can be easily obtained through shifting \( w[n,k,l]_{SR} \) to the right by 3 bits. Hence, for the sake of simplicity, the value of \( q \) should be a negative integer exponent of 2, which does not cause a serious limitation. Furthermore, \( w[n,k,l]_{SR} \) would have a limited range between \(-1\) and \(1\) for all values of \(0 < q < 1\). Thus, the calculation for \( w[n,k,l]_{XLD} \) from \( w[n,k,l]_{SR} \) defined in (12) can be done offline and the results are stored into a lookup table for online processing.

\[ 4. \] RESULTS AND DISCUSSION

In this section, we investigate the BER performance of the optimal, DR-based, and XLD-based MWTR receivers in intra-vehicle channels via simulations. A 2001 GMC Savana 3500 van was used to measure the intra-vehicle channels by using PulsON 210 radios made by Time Domain Corporation [24]. Then the channel impulse responses (CIRs) for the intra-vehicle channels were obtained by employing the methodology of [20]. Some key parameters for the obtained CIRs are given in Table 1, where \( \tau_{rms} \) is the root mean square (rms) delay spread (the square root of the second central moment of the power delay profile (PDP), where delays are measured relative to the time of arrival of the first detectable path which is set at \( t_0=0 \)), \( T_{mds} \) is the maximum delay spread (the time interval between the first and last detectable paths), and \( NP \) is the number of MPCs. For all CIRs, only the MPCs within a 15dB range from the strongest MPC are considered. Moreover, the average PDPs over 40 different channels are illustrated in Figure 6 for LOS and NLOS scenarios.

| Table 1. Channel parameters. |
|-----------------------------|
|                               | Min | Average | Max |
|                               | LOS | NLOS    | LOS | NLOS |
| \( \tau_{rms} \) (ns)       | 8.15 | 18.6    | 11.62 | 21.02 |
| \( T_{mds} \) (ns)          | 82.37 | 129.18  | 94.23 | 140.4 |
| \( NP \)                    | 179 | 295    | 209 | 358 |
|                             |     |        |     | 257 |
|                             |     |        |     | 473 |

Figure 4. Comparison of optimum combining weights and the XLD technology combining weights for \( x=0, 1, 2, 3, \) and \( 4 \) when \( q\rightarrow 0 \) and \( \beta/\alpha = 1 \)

Figure 5. Illustration of the effect of \( q \) on the optimum value of \( x \) in the XLD technology
The transmitted UWB pulse, $w_{tr}(t)$, used in simulations is a baseband pulse with a -10dB bandwidth of 500MHz. The sampling rate for the performance evaluations has been set at 1.07GHz. Other parameters for are set at: $\beta/\alpha=0.8$, $N_z=10$, and $n_d=n_s=214$. With $T=1/1.07$ns, $T_d=T_z=214T=200$ns and $T_B=2200$ns. From (7), it can be shown that the value of $w_{n,k,l}^{SRE}$ is mostly constructed by $L=2/q$ recently received blocks. For instance, for $q=0.1$, $L=20$. To obtain $w_{n,k,l}^{SRE}$ with high quality, the duration of $L$ consecutive blocks should be shorter than the channel coherence time. Since the typical channel coherence time for UWB applications is longer than 100$\mu$s [25] and $20T_B=44\mu$s<100$\mu$s, it is practical to choose the value of $q$ at a level greater than or equal to 0.1 for the MWTR receivers with the above block duration.

The BER performance of the XLD-based MWTR receivers have been thoroughly investigated through computer simulations for $x=1, 1.5, 2, 2.5, 3, 3.5, $ and 4 under LOS and NLOS scenarios, and the results are illustrated in Figures 7, 8, and 9 for $q$ values of 0.2, 0.15, and 0.1, respectively. Furthermore, based on the analysis given by [18], [23], the performance bound for the MWTR receivers has been derived and depicted in the same figure as a benchmark. It is worth mentioning that the DR-based MWTR receiver attains the BER of $10^{-3}$ at $E_b/N_0$ values of of 26.2 and 20 dB in LOS and NLOS channels, respectively. From Figures 7 to 9, we can have the following conclusions about the XLD-based MWTR receiver: i) the performance enhances as $q$ decreases. This is because smaller value of $q$ results in ac cumulation of larger number of reference pulses and thereby a higher SNR of the accumulated reference pulse; ii) since the XLD has the power of mitigating the effect of small (low-SNR) samples on the decision variable and such samples are more abundant in LOS channels than in NLOS channels, the performance improvement offered by the XLD scheme is larger in LOS channels, as a result the performance curves of XLD become closer to each other in NLOS scenario; iii) among all the values of $x$ the value of 3 shows almost the best performance in all cases; iv) for $x=3$ and BER=$10^{-3}$, the XLD-based MWTR receiver achieves 14.2~15.5 dB and 8~9.2 dB performance improvement over the DR-based MWTR receiver in LOS and NLOS scenarios, respectively; v) for $x=3$ and BER=$10^{-3}$, the performance curve of the XLD-based MWTR receiver is 1.6~3 dB away from that of the optimal MWTR receiver; vi) for $x=3$, the XLD-based MWTR receiver has a nearly robust performance no matter LOS or NLOS scenario is considered; and vii) unlike the prediction of Figure 5, in all cases except for $q=0.1$ in NLOS channels the performance increases as $x$ increases and the optimum value of $x$ is nearly independent of the value of $q$. 

Figure 6. Average PDPs for LOS and NLOS scenarios
5. CONCLUSIONS

In this paper, the x-law detection (XLD) algorithm was proposed to alleviate the performance loss of monobit quantization in TR-based UWB receivers. Through simulations, it was demonstrated that the XLD-based MWTR receiver for $x=3$ gives almost the best performance in all situations. Moreover, the XLD-based MWTR receiver for $x=3$ exhibits robustness to the channel variations because it gives nearly the same performance in LOS and NLOS scenarios. The XLD-based MWTR receiver for $x=3$ and BER=10$^{-3}$, achieves 14.2~15.5 dB and 8~9.2 dB performance improvement over the DR-based MWTR receiver in LOS and NLOS scenarios, respectively. Besides, the performance curve of the XLD-based MWTR receiver is 1.6~3 dB away from that of the optimal MWTR receiver at BER=10$^{-3}$ under intra-vehicle UWB channels. The XLD-based MWTR offers 6.2 dB more improvement over the DR-based one in LOS channels because the number of low-SNR samples in such channels is larger compared with that in NLOS channels. This implies that the XLD-based MWTR is insensitive to the summation time and is capable to considerably diminish the effect of low-SNR samples on the decision variable. This feature decreases the complexity of receiver design and guarantees a robust performance over different multipath channels. Last but not least, the considerable performance gain brought by the XLD technology comes at the expense of a slight increase in
the receiver complexity. Therefore, the XLD technology is considered as a promising technology for TR-based and other training-based monobit UWB receivers demanding high performance, low power consumption, and low complexity.

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