Low Complexity Suboptimal Monobit Receiver for Transmitted-Reference Impulse Radio UWB Systems

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***Abstract*—In this paper, a low complexity suboptimal monobit receiver, denoted as Q-function estimation (QFE) based monobit receiver, is proposed for transmitted-reference (TR) based impulse radio (IR) ultra wideband (UWB) systems to mitigate the performance degradation caused by severe quantization noise due to employing monobit ADCs. Through accumulating reference samples over multiple blocks, the QFE-based monobit receiver can considerably reduce the noise-on-noise effect and optimize the contribution of each sample to the decision statistic. Computer simulations demonstrate that in both line-of-sight (LOS) and non-LOS (NLOS) channels the QFE-based monobit weighted TR receiver can achieve a close-to-optimal bit-error rate (BER) performance without requiring full channel state information. Furthermore, because the optimal combining weights can be obtained offline and saved into a lookup table, the complexity of the proposed suboptimal monobit receiver is quite low. Therefore, it is considered as a promising technology for IR-UWB applications requiring low system complexity and low power consumption.**

***Index Terms*—Forgetting factor, monobit ADC, quantization noise, Q-function estimation, transmitted reference, ultra-wideband.**

I. Introduction

In impulse radio (IR) ultra wide band (UWB) systems, very short-duration pulses are used to carry information. Since those pulses can be transmitted without a carrier, IR-UWB systems have the advantages of low complexity, low power, and good time-domain resolution (for location and tracking applications). Nevertheless, transmitting short-duration pulses also introduces a critical problem in the receiver design for IR-UWB systems. In multipath environments, the received IR-UWB signals consist of a large amount of resolvable multipath components (MPCs) [1]. Tracking and estimating those MPCs lead to an unacceptably high complexity for the well-known Rake receiver [1, 2]. In order to capture more energy from MPCs and gain multipath diversity with low complexity, a group of transmitted-reference (TR) based receivers have been proposed to detect IR-UWB signals in multipath environments [3-5]. Unfortunately, all those TR receivers require delay elements with ultra wide bandwidth to provide delayed versions of the received IR-UWB signals. Because such delay elements are difficult (if not impossible) to be implemented with analog integrated circuits, the feasibility of TR-based receivers becomes a major technology bottleneck. Converting IR-UWB signals into digital signals is a straightforward approach to remove this bottleneck. However, as a sampling rate of several giga samples per second is needed to digitize IR-UWB signals, implementing high-resolution analog-to-digital converters (ADCs) becomes formidable due to the impractically high power consumption and complexity. Thus, applying finite-resolution ADCs, especially single-bit (monobit) ADCs, has attracted a lot of interest [6, 7]. Made by one or several comparators, finite-resolution ADCs work with much lower power consumption and complexity than high-resolution ones [6]. Nevertheless, finite-resolution ADCs, especially monobit ones, also lead to serious degradation on the performance of IR-UWB systems [7].

In order to mitigate the performance degradation caused by severe quantization noise due to employing monobit ADCs, in this paper, we have proposed a novel suboptimal detection approach, denoted as Q-function estimation (QFE) based detection approach, which calculates the optimal weights to combine the received IR-UWB data samples from the estimation results for the Marcum Q-function. Furthermore, we have applied this newly proposed approach to a monobit weighted TR (WTR) receiver [8] to improve the bit-error rate (BER) performance of the receiver. In a conventional monobit WTR receiver, reference samples are directly used as the weights to combine data samples. However, in the QFE-based detection approach, the reference samples in consecutive blocks are accumulated together to estimate the value of the Q-function, then the data samples are combined by using the suboptimal weights obtained from estimation results for the Q-function.

Through computer simulations, we have demonstrated that in both line-of-sight (LOS) and non-LOS (NLOS) intra-vehicle channels the QFE-based monobit WTR receiver can achieve a BER performance that is much better than the conventional monobit WTR receiver and only 2~3dB worse than the optimal monobit receiver requiring full channel state information. Furthermore, because the calculations to obtain the optimal weights to combine the data samples from the estimation results for the Q-function can be performed offline and the obtained results can be saved into a lookup table for online operations, the complexity to implement the proposed suboptimal detection approach is quite low.

The organization of the paper is as follows: Section II presents the system model of the conventional monobit WTR receiver, the optimal monobit receiver, the matched filter monobit receiver, and the QFE-based monobit WTR receiver. In Section III the BER performance of the above described receivers is investigated through computer simulations, and related discussions are given. Finally, conclusions are presented in Section IV.

II. System Model

A. Conventional Monobit WTR Receiver

In a WTR system, data are transmitted in blocks and each block consists of one reference pulse followed by N_s data pulses closely spaced in time. As shown in Fig. 1, to provide a guard time for the reference pulse, the first data pulse is transmitted T_d seconds after the reference pulse and the new block is transmitted T_d seconds after the last pulse of the previous block. Such a configuration significantly reduces inter-pulse interference (IPI) and inter-block interference (IBI) on the reference pulse and results in a nearly clean reference pulse. The destructive effect of noise on the reference pulse is mitigated by allocating more power to the reference pulse than to the data pulses. Hence, in the WTR system the transmitted signal can be expressed as follows:

$$s_{tr}(t) = \sum_{j=-\infty}^{\infty} \left\{ \begin{array}{l} \alpha w_{tr}(t - jT_B) \\ +\beta \sum_{i=0}^{N_s-1} b_{j,i} w_{tr}(t - jT_B - iT_s - T_d) \end{array} \right\}, \tag{1}$$

where $w_{tr}(t)$ is a baseband unit-energy UWB pulse with a bandwidth of W and a duration of T_w, T_s is the time spacing between two consecutive data pulses, T_d is the guard time between the reference pulse and the previous/subsequent data pulses, N_s denotes the number of data pulses per block, and $T_B=2T_d+(N_s-1)T_s$ is the duration of the block. In the same equation, $b_{j,i} \in \{-1,+1\}$ represents the i^{th} information bit of the j^{th} block. Finally, α and β represent the weights of the reference pulse and data pulses, respectively. In order to normalize energy per bit, α and β satisfy the following equation:

$$\alpha^2 + N_s \beta^2 = N_s \,. \tag{2}$$

The structure of a WTR receiver employing a monobit ADC is depicted in Fig. 2. The received signal is first passed through a low-pass filter (LPF) with an impulse response of $f(t)$ and a bandwidth of W to eliminate out-of-band noise and interferences. Thus, the received signal at the output of the LPF is given by:

$$\begin{aligned} r(t) &= \sum_{j=-\infty}^{\infty} \left\{ \begin{array}{l} \alpha g(t - jT_B - \theta) \\ +\beta \sum_{i=0}^{N_s-1} b_i g(t - jT_B - iT_s - T_d - \theta) \end{array} \right\} + v(t) \\ &= s(t) + v(t)\,, \end{aligned} \tag{3}$$

where θ is the asynchronous delay between the receiver and the transmitter and $v(t)$ is the filtered additive white Gaussian noise (AWGN) with a two-sided power spectral density of $N_0 |F(f)|^2 / 2$. Here $F(f)$ is the Fourier transform of $f(t)$. In the same equation, $g(t) = w_{tr}(t) \otimes C(t) \otimes f(t)$ ($\otimes$ denotes linear convolution) is the received noise-free waveform corresponding to a transmitted pulse, where $C(t)$ is the impulse response of the multipath channel. Without loss of generality, throughout this paper, we assume that $f(t)$ is an ideal LPF and θ=0.

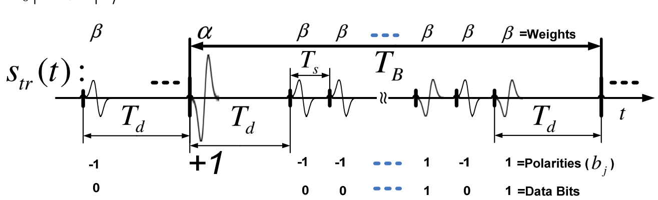

Fig. 1. A typical transmitted signal of the WTR system.

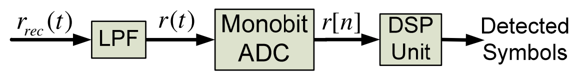

Fig. 2. The receiver structure of the WTR system using a monobit ADC.

Second, the received signal is sampled at the Nyquist rate, resulting in a sampling interval of $T=1/(2W)$:

$$r(nT) = s(nT) + v(nT), \tag{4}$$

and then digitized by a monobit ADC as follows:

$$r[n] = \begin{cases} +1, & r(nT) \geq 0 \\ -1, & r(nT) < 0 \end{cases}. \tag{5}$$

In this paper, it is assumed that $T_d=n_dT$, $T_s=n_sT$, and $T_B=n_bT$, where n_d, n_s, and n_b are integers.

Finally, the digitized signal is fed to a digital signal processing (DSP) unit to detect information bits. As shown in Fig. 3, in a conventional monobit WTR receiver (i.e. the switch is rest at position 1), in order to demodulate the k^{th} information bit of the block, the received signal is first multiplied by its kn_s+n_d delayed replica (so that the reference samples are aligned with the k^{th} data samples). The result of the multiplications are then summed from the $(kn_s+n_d)^{th}$ sample to the $(kn_s+n_d+n_i-1)^{th}$ one, i.e. over a duration of n_i samples and typically $n_i \leq n_d$. Here n_iT is equivalent to the integration interval, T_i, in an analog WTR receiver. Thus, the summation output corresponding to the k^{th} information bit in the conventional monobit WTR receiver is given by:

$$z[k] = \sum_{n=0}^{n_i-1} r[n] r[n + kn_s + n_d]\,, \tag{6}$$

and the k^{th} information bit is detected by a sign detector as follows:

$$\hat{b}_k = sign(z[k])\,. \tag{7}$$

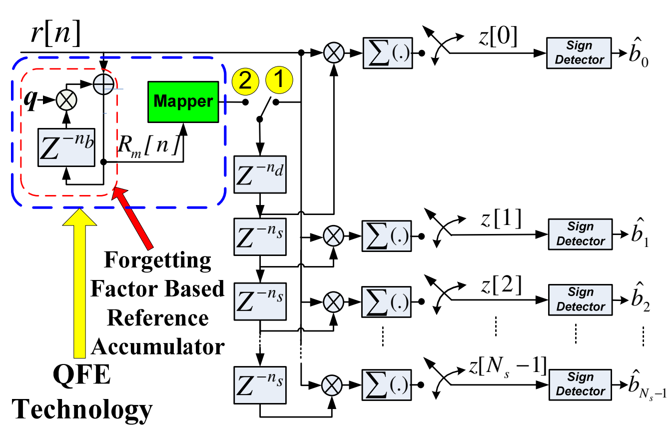

Fig. 3. The diagram of the DSP unit for the conventional (switch is at position 1) and the QFE-based (switch is at position 2) monobit WTR receivers.

It is worth mentioning that in order to reduce the system complexity even more the approach given in [9] can be used for digital receiver shown in Fig. 3 which halves the required delay elements.

B. Optimal and Matched Filter Monobit Receivers

As shown in [6], the optimal detection approach for monobit WTR receivers should be a weighted linear combination of the data samples followed by a sign detector:

$$\hat{b}_k = sign\left(\sum_{n=0}^{n_i-1} w_n^{op} r[n+kn_s+n_d]\right), \tag{8}$$

where for $n \in \{0,1,\ldots,n_i\text{-}1\}$ the optimal combining weights are given by [6]:

$$w_n^{OP} = \log\left(\frac{1-Q(\beta\, g[n]/\sigma)}{Q(\beta g[n]/\sigma)}\right) = \log\left(\frac{1-\varepsilon_n^{\beta}}{\varepsilon_n^{\beta}}\right). \tag{9}$$

In the above equation, $g[n]$ are the full-resolution samples of $g(t)$, the received noise-free waveform corresponding to a transmitted pulse, $Q(x)$ is the Marcum Q-function, and $\sigma = \sqrt{N_0 W}$.

If the combining weights in (8) are set as follows:

$$w_n^{MF} = Cg[n], \tag{10}$$

where C is a constant, the detection approach becomes the well-known matched filter (MF) receiver. It should be noted that although the MF receiver is not an optimal monobit receiver, for most application scenarios it can achieve the similar performance as the optimal monobit receiver. It is because that when the value of $\beta g[n]/\sigma$ is not very large, i.e. the signal-to-noise ratio (SNR) of the monobit data samples is not very high, w_n^{MF} and w_n^{OP} always have the similar values.

C. Q-function Estimation Based Monobit Receiver

Although the optimal and the MF monobit receivers can achieve a much better performance than the conventional monobit WTR receiver, they require full channel state information to derive $g[n]$, which is very difficult (if not impossible) to obtain in real world. In this paper, through estimating the value of ε_n^{β} from the monobit reference samples in consecutive blocks, referred to as Q-function estimation (QFE), we have proposed a suboptimal detection approach for the monobit WTR receiver, which can achieve a close-to-optimal performance without requiring prior knowledge about $g[n]$.

As shown in Fig. 3, when the switch is at position 2, by delaying the monobit reference samples by n_b clock cycles, they align with the monobit reference samples of the subsequent block. Therefore, the monobit reference samples in different blocks can be accumulated coherently. In order to control the number of the blocks to be accumulated, a forgetting factor, $0<q<1$, is introduced to exponentially reduce the contribution of the reference samples in early received UWB blocks. The optimal value for q is determined by the coherence time of the UWB channel and the synchronization accuracy of the WTR receiver.

Assuming that the IPI and IBI for reference pulses are negligible, we can mathematically express the accumulated reference samples for the m^{th} block as follows:

$$R_m[n] = \sum_{l=-\infty}^{m} q^{m-l} r_{n,l}\,, \tag{11}$$

where

$$r_{n,l} = sign(\alpha g[n] + v[n+ln_b]),\ \ n \in \{0,1,\ldots,n_i-1\}. \tag{12}$$

Because $v(n)$ is a filtered AWGN with a variance of $\sigma^2 = N_0 W$, the probability distribution of $r_{n,l}$ can be expressed as:

$$P(r_{n,l}) = \begin{cases} 1-\varepsilon_n^{\alpha}, r_{n,l} = +1 \\ \varepsilon_n^{\alpha}, \quad\ r_{n,l} = -1 \end{cases}, \tag{13}$$

where

$$\varepsilon_n^{\alpha} = Q\left(\frac{\alpha g[n]}{\sigma}\right). \tag{14}$$

Thus, from (11)-(14), the mean value of $R_m[n]$ can be derived as:

$$E\{R_m[n]\} = \sum_{l=-\infty}^{m} q^{m-l} E\{r_{n,l}\} = \frac{1}{1-q}(1-2\varepsilon_n^{\alpha}). \tag{15}$$

Meanwhile, the second moment of $R_m[n]$ is given by:

$$\begin{aligned} E\left\{R_m^2[n]\right\} &= E\left\{\left(\sum_{l=-\infty}^{m} q^{m-l} r_{n,l}\right)^2\right\} \\ &= E\left\{\sum_{l=-\infty}^{m}\sum_{l'=-\infty}^{m} q^{m-l} q^{m-l'} r_{n,l} r_{n,l'}\right\} \\ &= \sum_{l=-\infty}^{m}\sum_{\substack{l'=-\infty \\ l'\neq l}}^{m} q^{m-l} q^{m-l'} E\{r_{n,l} r_{n,l'}\} + \sum_{l=-\infty}^{m} q^{2(m-l)}. \end{aligned} \tag{16}$$

Because $r_{n,l}$ and $r_{n,l'}$ are independent to each other when $l \neq l'$, the above equation can be rewritten as:

$$\begin{aligned} E\left\{R_m^2[n]\right\} &= \sum_{l=-\infty}^{m}\sum_{\substack{l'=-\infty \\ l'\neq l}}^{m} q^{m-l} q^{m-l'} E\{r_{n,l}\} E\{r_{n,l'}\} + \sum_{l=-\infty}^{m} q^{2(m-l)} \\ &= \left[1-2\varepsilon_n^{\alpha}\right]^2 \left(\frac{1}{(1-q)^2} - \frac{1}{1-q^2}\right) + \frac{1}{1-q^2}. \end{aligned} \tag{17}$$

Therefore, from (15) and (17), the variance of $R_m[n]$ can be derived as:

$$\sigma_{R_m[n]}^2 = E\{R_m^2[n]\} - E^2\{R_m[n]\} = \frac{1}{1-q^2}\left\{1-\left[1-2\varepsilon_n^{\alpha}\right]^2\right\}. \tag{18}$$

Furthermore, from (15) and (18), the ratio between the mean value and the standard deviation of $R_m[n]$ can be expressed as:

$$\frac{E\{R_m[n]\}}{\sigma_{R_m[n]}} = \sqrt{\frac{1+q}{1-q}}\,\frac{1-2\varepsilon_n^{\alpha}}{2\sqrt{\varepsilon_n^{\alpha}(1-\varepsilon_n^{\alpha})}}. \tag{19}$$

Clearly, when q takes a value close to one, the mean value of $R_m[n]$ is much larger than its standard deviation, and hence based on (15) we can estimate the value of ε_n^{α} from that of $R_m[n]$ with the following equation:

$$\hat{\varepsilon}_n^{\alpha} = \{1-(1-q)R_m[n]\}/2 \,. \tag{20}$$

From the above equation, the value of ε_n^{β} can be estimated as:

$$\hat{\varepsilon}_n^{\beta} = Q\left[\frac{\beta}{\alpha}Q^{-1}\left(\hat{\varepsilon}_n^{\alpha}\right)\right] = Q\left[\frac{\beta}{\alpha}Q^{-1}\left(\frac{1}{2}\{1-(1-q)R_m[n]\}\right)\right], \tag{21}$$

where $Q^{-1}(x)$ is the inverse Marcum Q-function. Substituting ε_n^{β} given by (21) into (9) yields the following suboptimal weights for the QFE-based monobit receiver:

$$w_n^{QFE} = \log\left(\frac{1-\hat{\varepsilon}_n^{\beta}}{\hat{\varepsilon}_n^{\beta}}\right). \tag{22}$$

According to (11), the maximum and the minimum values of $R_m[n]$ are $1/(1\text{-}q)$ and $-1/(1\text{-}q)$, respectively. When $R_m[n]$ reaches to its maximum or minimum value, from (20) $\hat{\varepsilon}_n^{\alpha}$ is equal to zero or one, and correspondingly from (21) $\hat{\varepsilon}_n^{\beta}$ is equal to zero or one as well. Consequently, the value of w_n^{QFE} given by (22) becomes positive or negative infinite; therefore a small estimation error in $\hat{\varepsilon}_n^{\beta}$ for high SNR samples leads to a large detection error and consequently severe performance loss. In order to overcome this weakness, we have revised (22) as follows:

$$w_n^{QFE} = \begin{cases} log\left(\dfrac{1-\varepsilon}{\varepsilon}\right), & \hat{\varepsilon}_n^{\beta} \le \varepsilon, \\ log\left(\dfrac{1-\hat{\varepsilon}_n^{\beta}}{\hat{\varepsilon}_n^{\beta}}\right) & \varepsilon < \hat{\varepsilon}_n^{\beta} < 1-\varepsilon, \\ log\left(\dfrac{\varepsilon}{1-\varepsilon}\right) & \hat{\varepsilon}_n^{\beta} \ge 1-\varepsilon, \end{cases} \tag{23}$$

where ε is a small positive value.

It should be noted that because $R_m[n]$ has a limited range, $-1/(1\text{-}q) \le R_m[n] \le 1/(1\text{-}q)$, once the values of q, α, β, and ε are determined, the calculations defined by (21) and (23) can be performed offline and the calculation results are saved into a lookup table. As shown in Fig. 3, in online mode for each given value of $R_m[n]$ the Mapper obtains the value of w_n^{QFE} from the lookup table. Thus, the complexity to implement the proposed suboptimal detection approach is quite low.

III. SIMULATION RESULTS AND DISCUSSION

In this section, through computer simulations the BER performances of the following four receivers are thoroughly evaluated in LOS and NLOS intra-vehicle channels given by [10]: *i*) a conventional monobit WTR receiver, *ii*) an optimal monobit receiver, *iii*) an MF monobit receiver, and *iv*) a QFE-based monobit WTR receiver.

In the computer simulations, a baseband UWB pulse with a -10dB bandwidth equal to 500MHz has been used as $w_{tr}(t)$, and the ADC sampling rate is set at 1.07GHz. Other simulation parameters are summarized in Table I, where T=1/1.07ns is the ADC sampling duration. Furthermore, it is assumed that the received UWB signals are perfectly synchronized in all four receivers. Finally, after extensive simulations the value of ε is set at 0.001.

TABLE I. SIMULATION PARAMETERS

System	Conventional	QFE	Optimal/MF
N_s	10	10	10
T_d	$107T\approx$100ns	$107T\approx$100ns	NA
T_s	$107T\approx$100ns	$107T\approx$100ns	$118T\approx$110ns
T_B	$1177T\approx$1100ns	$1177T\approx$1100ns	$1180T\approx$1103ns
T_i	$107T\approx$100ns	$107T\approx$100ns	$118T\approx$110ns
R_b	9.09Mbps	9.09Mbps	9.07Mbps
β/α	0.5	1	NA

From (11), it is straightforward to derive that approximately 99% of the energy of $R_m[n]$ is provided by the most recently received 2/(1-q) UWB blocks. For example, when q=0.9, $R_m[n]$ is mainly determined by the most recently received 20 UWB blocks. In order to obtain $R_m[n]$ with a high quality, the duration of the $2/(1\text{-}q)$ UWB blocks must be shorter than the coherent time of the channel. Since typically the channel coherent time for UWB systems is larger than 100μs [11] and $20T_B$=22μs<100μs, it is reasonable to set the value of q at 0.8 or 0.9 for the systems with the parameters given by Table I.

The sensitivity of the conventional and the QFE-based monobit WTR receivers to the value of β/α is investigated over both LOS and NLOS channels and the obtained results are illustrated in Fig. 4. It can be seen that the conventional monobit receiver becomes more sensitive at higher β/α values, and when E_b/N_0=14dB the optimum β/α value is 0.5. On the contrary, the QFE-based receiver is more sensitive at lower β/α values, and when E_b/N_0=12dB, the optimum β/α value is 0.7 for q=0.8 and 0.8 for q=0.9. Because maintaining the same weight for the reference pulse and the data pulses reduces the peak-to-average power ratio of the system and correspondingly reduces the complexity of the system, so we have fixed the value of β/α at one for the QFE-based receiver

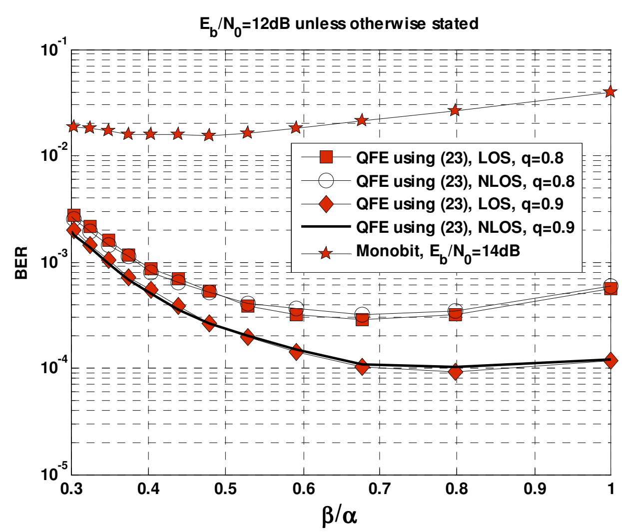

Fig. 4. The effect of β/α value on the performance of the conventional monobit and the QFE-based WTR monobit receivers.

in the rest simulations. As shown in Fig. 4, for both LOS and NLOS scenarios, the performance of QFE-based receiver increases with q. This is because a larger value of q leads to averaging over more reference pulses and accordingly a reference pulse with a higher quality is obtained. As a result, according to (20), a more accurate estimation of $\hat{\varepsilon}_n^{\alpha}$ is guaranteed. Furthermore, Fig. 4 represents that the QFE-based monobit WTR receiver is not sensitive to the delay spread of multipath channels and has almost the same performance in both LOS and NLOS channels. This implies the following two conclusions. First, in both LOS and NLOS scenarios, almost all the energy of the received pulses is included in the integration interval. Second, low SNR (noise dominant) samples which have larger number in LOS scenario than NLOS one due to more dispersion in NLOS, can barely degrade the performance of the QFE-based WTR receiver.

Then the BER performances of the four monobit receivers as a function of Eb/N0 are investigated through computer simulations. The simulation results for the LOS scenarios are depicted in Figs. 5 and 6 for q=0.8 and 0.9, respectively, and the simulation results for the NLOS scenarios are illustrated in Fig. 7 when q=0.8 and in Fig. 8 when q=0.9. From the simulation results, we can draw the following conclusions for the four monobit receivers. The conventional monobit WTR receiver has the worst performance among the four receivers, even with the optimal value for β/α. For example, when BER is 10^{-3}, as shown in Fig. 5, the conventional monobit WTR receiver is 10.3dB worse than the optimal monobit receiver in the LOS scenarios; as shown in Fig. 7, it is 11.5dB worse than the optimal monobit receiver in the NLOS scenarios.

The optimal monobit and the MF monobit receivers have almost the same BER performance. As shown in Fig. 5, only at high SNR regime the optimal monobit receiver slightly out-

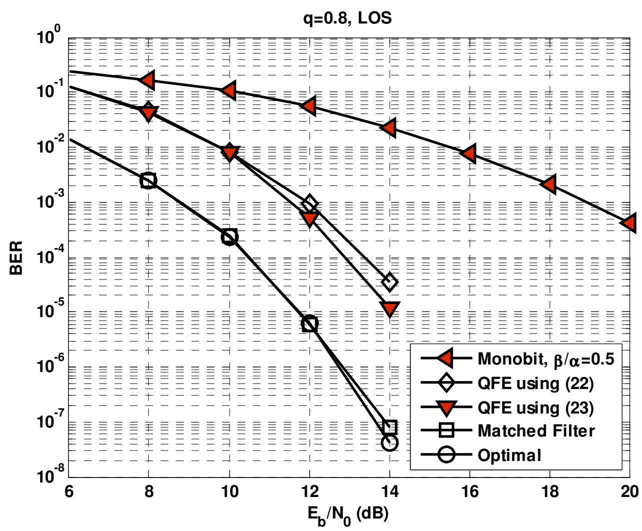

Fig. 5. Performance comparisons of the conventional, the QFE-based, the optimal, and the MF monobit receivers for q=0.8 in LOS scenarios.

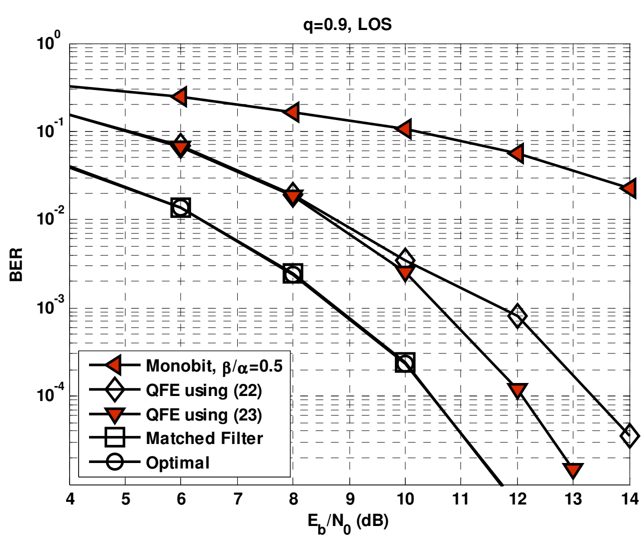

Fig. 6. Performance comparisons of the conventional, the QFE-based, the optimal, and the MF monobit receivers for q=0.9 in LOS scenarios.

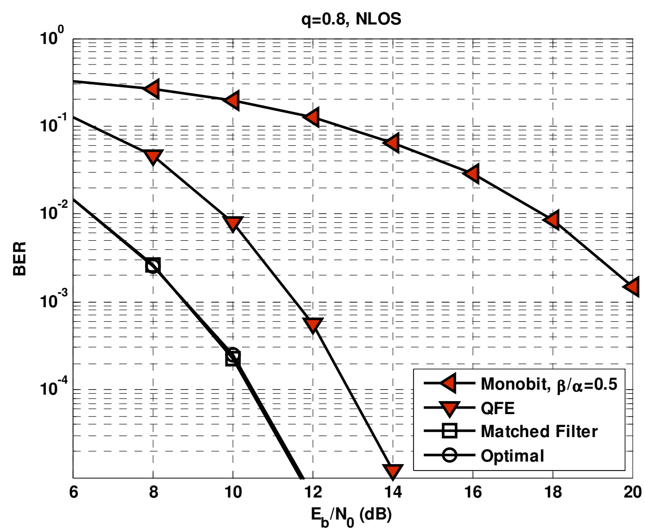

Fig. 7. Performance comparisons of the conventional, the QFE-based, the optimal, and the MF monobit receivers for q=0.8 in NLOS scenarios.

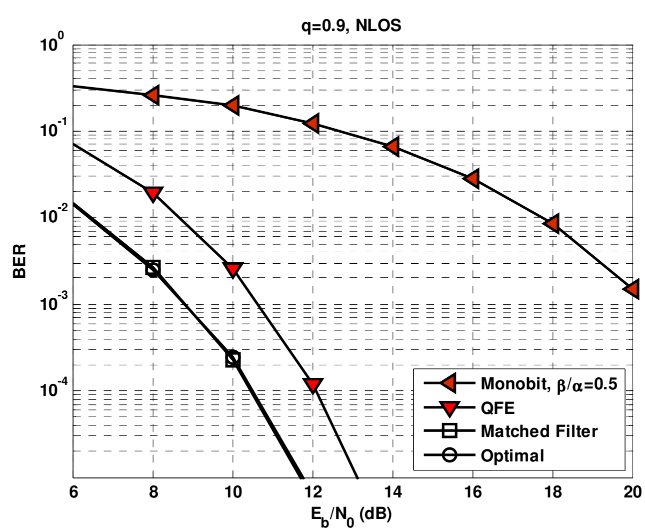

Fig. 8. Performance comparisons of the conventional, the QFE-based, the optimal, and the MF monobit receivers for q=0.9 in NLOS scenarios.

performs the MF one. For the QFE-based monobit WTR receiver, first its BER performance is much better than that of the conventional monobit WTR receiver, and is only 2~3dB worse than that of the optimal monobit receiver. Second, according to (19) increasing the value of q can improve the accuracy on estimating and hence can further enhance the performance of the QFE-based monobit WTR receiver. For example, as shown in Figs. 5 and 6, when q increases from 0.8 to 0.9, the BER performance of the receiver enhances about 1dB. Third, the simulation results shown in Figs. 5 and 6 indicate that calculating w_n^{QFE} with (23) can achieve a considerable performance improvement over calculating w_n^{QFE} with (22), especially for the high SNR situations. This is because (23) can effectively limit the range of w_n^{QFE} and avoid the large detection errors when $\hat{\varepsilon}_n^{\beta}$ takes an infinite value. Last but not the least, as shown in Figs. 4-8, unlike the conventional monobit WTR receiver, the QFE-based monobit WTR receiver is not sensitive to the delay spread of multipath channels and has almost the same performance in both LOS and NLOS channels. This is an important feature that the QFE-based monobit WTR receiver inherits from the optimal monobit receiver: close-to-zero weights are assigned to the noise-dominant samples and hence those samples can barely affect the decision statistics.

IV. CONCLUSIONS

In this paper, we have proposed a low complexity suboptimal monobit receiver, the QFE-based monobit receiver, to mitigate the performance degradation caused by severe quantization noise due to employing monobit ADCs, and applied the newly proposed receiver to a WTR IR-UWB system. Computer simulations demonstrate that in both LOS and NLOS intra-vehicle channels the QFE-based monobit WTR receiver can achieve a BER performance that is much better than the conventional monobit WTR receiver and is only 2~3dB worse than the optimal monobit receiver requiring full channel state information. Furthermore, in the QFE-based monobit receiver, because the calculations to obtain the optimal combining weights from the estimation results for the Q-function can be performed offline and the calculation results can be saved into a lookup table for online operations, the complexity to implement the proposed suboptimal detection approach is quite low. Therefore, the proposed suboptimal monobit receiver can be considered as a promising technology for IR-UWB applications requiring low system complexity and low power consumption.

ACKNOWLEDGMENT

This work is supported by the National Science Foundation under Grant No. 1002150 and 1002113, Natural Science and Engineering research Council of Canada under Grant No. STPGP 396417-10, and Faculty Summer Fellowship of University of Northern Iowa.

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