Monobit Digital Receivers for QPSK: Design and Performance Analysis

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Abstract

Future communication system requires large bandwidth to achieve high data rate, which makes analog-to-digital (ADC) a key bottleneck due to its high complexity and large power consumption. Therefore, we consider the monobit receivers for QPSK in this paper. First, the optimal monobit receiver with Nyquist sampling and its performance are derived. Then a suboptimal but more practical receiver is obtained. The effect of the imbalances between the In-phase (I) and Quadrature (Q) branches is investigated too. To combat the performance loss, monobit receivers based on double training sequences and 8-sector phase quantization are proposed. Numerical simulations show that the low-complexity suboptimal receiver suffers only 3dB signal-to-noise-ratio (SNR) loss in AWGN channels and 1dB SNR loss in multipath static channels compared with the matched-filter based monobit receiver with full channel state information (CSI). It is demonstrated that the amplitude imbalance has no effect on monobit receivers. Receivers based on double training sequences can efficiently compensate for the SNR loss in AWGN channels without complexity increase. By doubling the complexity, receivers with 8-sector phase quantization can almost completely eliminate the SNR loss caused by IQ imbalances. The effect of imbalances on monobit receivers in dense multipath channels is slight.

Index Terms

Analog-to-digital conversion, deflection ratio, impulse radio, imbalance, monobit, QPSK modulation, ultra-wideband

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I. INTRODUCTION

In the future, communication systems are expected to provide high data rate up to several giga bit per second. To achieve this, extremely large bandwidth is needed. For instance, the ultra-wideband (UWB) communication occupies more than 1 GHz spectrum. The communication in the 60 GHz band [1] takes up even more. Due to the significant large bandwidth, it is a huge challenge to design a sophisticated digital receiver with implementation simplicity.

When the received signals of the high-rate high bandwidth systems are processed digitally, the analog-to-digital converter (ADC) becomes a key bottleneck. Since the power consumption of ADC is proportional to 2^b, where b is the bit number of the ADC [2], high-speed high-resolution ADC is power-hungry and costly. Therefore, single-bit or monobit ADC has attracted great attention in recent years, both in the aspect of receiver design, e.g., [3], [4], and the aspect of information theory, e.g. [5], [6]. As it can be simply realized by fast comparator, the monobit ADC can reach tens of Gsps sampling rate with very low power consumption; see, e.g., [7].

The monobit ADC is particularly suitable for UWB communication using impulse radio (IR), whose traditional reception methods mainly include coherent receiver, e.g., [8], autocorrelation based receiver, e.g. [9], [10], and noncoherent receiver, e.g. [11]. After the monobit sampling was introduced, a matched-filter based receiver has been proposed in [4] for BPSK modulation. However, it is not optimal under Nyquist sampling as proved in [12]. Moreover, the requirement for full resolution (FR) ideal received waveform makes it difficult to implement.

The optimal monobit receiver for BPSK was proposed in [12], which turns out to take the form of a linear combiner. By performing a Taylor's expansion of the optimal weights, a suboptimal receiver was also derived in [12]. Besides, some simple but useful techniques such as iterative demodulation and removal of small-weight points were shown to be effective. Compared with the receiver given by [4], the receiver proposed in [12] is easy to implement and only incurs a slight performance loss, even without the channel state information (CSI).

In order to achieve higher data rate, higher order modulations such as QPSK or QAM are considered. For standard uniform Phase Shift Keying (PSK) modulation with "phase-quantization", the capacity of noncoherent additive white Gaussian noise (AWGN) channel is calculated in [13]. However, the architecture of such "phase-quantization" is more complex than the traditional one. Furthermore, detailed receiver design is not discussed in [13].

In this paper, we study the design and performance of digital receivers for QPSK under the traditional receiver architecture, based on monobit Nyquist rate sampling. First, the maximum likelihood (ML) receiver is derived. To simplify the computation, its performance is analyzed in the form of deflection ratio (see, e.g., [14]). Secondly, the main ideas in [12] are extended here to obtain a suboptimal receiver for QPSK. The effect of the phase difference between the transmitter and the receiver is investigated. Besides, the interface with error-control decoder is also obtained. Compared with the matched-filter based monobit receiver with full CSI, it is shown that the suboptimal receiver without any prior information has only 3dB signal-to-noise-ratio (SNR) loss in AWGN channel and 1dB SNR loss in fading channel, under the assumption of perfect timing and no inter-symbol-interference (ISI). It can also be observed that the resulting practical receiver has only 3dB SNR loss even compared with the full-resolution matched filter (FRMF) in fading channel.

A limiting factor of practical systems is the imbalance between the In-phase (I) and Quadrature (Q) branches when received radio-frequency (RF) signal is down-converted to baseband. Basically, the IQ imbalances, including amplitude and phase imbalances, are any mismatch between the I and Q branches from the ideal case [15]. Compared to the heterodyne receiver, the direct conversion RF receiver considered in this paper is affected more seriously by the IQ imbalances [16]. Although this problem is well investigated in traditional receivers with high-resolution ADC, we are not aware of any published work dealing with the IQ imbalances under low-resolution sampling.

We will investigate the effect of the IQ imbalances on monobit receivers. Then, monobit receivers based on double training sequences are proposed to mitigate the SNR loss caused by IQ imbalances, without sacrificing the simplicity of implementation. To further improve monobit reception under IQ imbalances, the 8-sector phase quantization proposed in [13] is employed and the corresponding practical monobit receiver is obtained, at the price of doubling the complexity in the digital domain. It is shown that the amplitude imbalance has no effect on the monobit receivers. It can be demonstrated that the monobit receiver with 8-sector phase quantization can almost completely eliminate the SNR loss caused by IQ imbalances in AWGN and sparse multipath channels. Thanks to the diversity offered by dense multiphath, monobit receivers based on traditional architecture are more desirable in dense multipath channel compared with receivers with 8-sector phase quantization, for their slight performance loss but great simplicity.

The rest of the paper is organized as follows. The system model and receiver architecture are presented in the next section. The optimal and suboptimal monobit receivers without IQ imbalance are given in section III, along with discussion on several important practical issues such as channel estimation, effect of phase difference and interface with error-control decoder. Section IV discusses the effect of IQ imbalances and proposes monobit receivers based double training sequences or 8-sector phase quantization to combat the performance degradation. Application of the proposed receivers to large wideband signaling and numerical results are provided in section V. At last, section VI concludes the paper.

II. System Model And Receiver Architecture

The monobit digital receiver we study is depicted in the block diagram in Figure 1. The received baseband signal is composed of the I and Q components. Both of them are first filtered by an ideal low pass filter (LPF) of bandwidth B, then sampled and quantized by a monobit ADC at Nyquist rate $2B$. The digitized signals are processed by the digital signal processing (DSP) unit.

In order to get better bit error rate (BER) performance, the Gray coded QPSK modulation is employed here [17]. Thus, the transmitted signal can be written as follows

$$s\left(t\right)=\sqrt{2}\Re\left\{\sum_{k=0}^{\infty}e^{j\left(g\left(d_{k1},d_{k0}\right)+2\pi f_c t\right)}p_{\text{tr}}\left(t-kT_s\right)\right\} \tag{1}$$

where k is the symbol index, T_s is symbol duration, f_c is the carrier frequency and $d_{k0}, d_{k1} \in \{+1,-1\}$ are the binary data of the kth QPSK symbol. $g\left(d_{k1}, d_{k0}\right)$ is the QPSK modulation function according to Gray coding rules, that is $g\left(1,1\right)=0$, $g\left(1,-1\right)=\pi/2$, $g\left(-1,1\right)=-\pi/2$ and $g\left(-1,-1\right)=\pi$. p_{tr} is the spectral shaping pulse. We assume that d_{k0} and d_{k1} are equally likely to be ±1, and they can be either uncoded or coded.

The channel is modeled as a linear time-invariant system with a finite impulse response $h\left(t\right)$. In the case of wireless time-varying channel, we assume that within the coherence interval the channel can be modeled as time-invariant. The received RF signal can be written as $r_{\text{rf}}\left(t\right)=s\left(t\right)\star h\left(t\right)+n\left(t\right)$, where $\star$ denotes convolution, and $n\left(t\right)$ is AWGN with double-sided power spectral density $N_0/2$. We assume that there is no interference here.

At the receiver, the received RF signal is first down-converted to baseband. Then, both the I and Q components of the received baseband signal are filtered by an ideal LPF respectively. The bandwidth of the ideal LPF is B, and its impulse response is $p_{\text{rec}}(t) = \sin(2\pi Bt) / \left(\pi t\sqrt{N_0 B}\right)$. The gain of the LPF, $1/\sqrt{N_0 B}$, is chosen so that the noise variance after sampling will be normalized to one. We define $p_{\text{ref}}(t) = p_{\text{tr}}(t) \star h(t) \star p_{\text{rec}}(t)$ as the reference signal. It can be assumed that the frequency of the local carrier is perfectly locked to the carrier of the incoming signal. If there is no imbalance between the I and Q branches, the filtered baseband signal can be given as

$$r_{\text{b}}(t) = \sum_{k=0}^{\infty} e^{j(g(d_{k1},d_{k0})-\varphi)} p_{\text{ref}}(t - kT_s) + n_{\text{b}}(t) \tag{2}$$

where φ is the carrier phase difference of the transmitter and the receiver, and $n_{\text{b}}(t)$ is the baseband-equivalent noise. The phase difference φ is an unknown constant with uniform distribution on $[0, 2\pi)$.

Due to analog component imperfections, the I and Q branches of the receiver usually do not have equal amplitude or exact 90° phase difference, leading to the amplitude and phase imbalance respectively. Similar to [15] and [18], the received baseband signal distorted by the IQ imbalance can be modeled as

$$r_{\text{d}}(t) = \mu r_{\text{b}}(t) + \upsilon r_{\text{b}}^*(t) \tag{3}$$

where μ and υ, characterizing the imbalance between the I and Q branches, are given as follows

$$\begin{aligned} \mu &= \cos(\theta/2) - j\alpha\sin(\theta/2) \\ \upsilon &= \alpha\cos(\theta/2) + j\sin(\theta/2) \end{aligned} \tag{4}$$

where θ denotes the phase deviation between the I and Q branches from the ideal 90°, and α denotes the amplitude imbalance given as

$$\alpha = \frac{a_I - a_Q}{a_I + a_Q} \tag{5}$$

where a_I and a_Q are the gain amplitude on the I and Q branches, respectively. When stated in dB, the amplitude imbalance is computed as $10\log(1+\alpha)$.

We choose the filter bandwidth B and the sampling period T to be $T = 1/(2B) = T_s/N$, so that the Nyquist rate sampling of the filtered signal is used and every pulse is sampled by N points. Within the kth symbol, we denote the lth samples of the I and Q branches as $r_{I,k,l}$ and

$r_{Q,k,l}$ respectively. Then we have

$$r_{I,k,l} = \begin{cases} 1, & r_{\mathrm{d},I}\left(kT_s + lT\right) > 0 \\ -1, & r_{\mathrm{d},I}\left(kT_s + lT\right) \le 0 \end{cases} \qquad l = 0, ..., N-1 \tag{6}$$

and

$$r_{Q,k,l} = \begin{cases} 1, & r_{\mathrm{d},Q}\left(kT_s + lT\right) > 0 \\ -1, & r_{\mathrm{d},Q}\left(kT_s + lT\right) \le 0 \end{cases} \qquad l = 0, ..., N-1 \tag{7}$$

where $r_{\mathrm{d},I}\left(t\right)$ and $r_{\mathrm{d},Q}\left(t\right)$ denote the received signal of the I and Q branches respectively. We assume the maximum delay spread is significantly smaller than symbol duration T_s so that ISI is negligible.

Define $\mathbf{r}_k = \left[r_{I,k,0}, r_{Q,k,0}, ..., r_{I,k,N-1}, r_{Q,k,N-1}\right]^T$. The digital signal processing unit of the receiver is to detect d_{k0}, d_{k1} based on $\mathbf{r}_k$.

III. Monobit Receivers Without IQ Imbalance

In this section, we first derive the optimal monobit detector, assuming that there is no IQ imbalance at the receiver. However, the precise reference signal $p_{\mathrm{ref}}\left(t\right)$ and phase difference φ required by such receiver are not available in practice. Hence, we extend the main ideas in [12] to QPSK modulation, and obtain a practical monobit receiver for QPSK. The performance of these monobit receivers, the effect of the phase difference and the interface with error-control decoder are also discussed.

A. Optimal Monobit Receiver

In communication systems, we commonly assume that d_{k0} and d_{k1} are equally likely to be ± 1. This implies that the optimal detector, based on the digital samples $\mathbf{r}_k$, is the maximum-likelihood (ML) detector, which minimizes the symbol error probability. Define

$$\epsilon_{I,l} = Q\left(p_{\mathrm{ref}}\left(lT\right)\cos\left(\varphi\right)\right), \quad \epsilon_{Q,l} = Q\left(p_{\mathrm{ref}}\left(lT\right)\sin\left(-\varphi\right)\right) \tag{8}$$

where $Q\left(\cdot\right)$ is the Q function $Q\left(x\right) = \frac{1}{\sqrt{2\pi}}\int_x^\infty e^{-\frac{t^2}{2}}dt$. The parameters $\epsilon_{I,l}$ and $\epsilon_{Q,l}$ can be viewed as the error probability for binary transmission of the lth "chip" $p_{\mathrm{ref}}\left(lT\right)\cos\left(\varphi\right)$ and $p_{\mathrm{ref}}\left(lT\right)\sin\left(-\varphi\right)$ through AWGN channel, respectively. Thanks to the memoryless property of the AWGN, the log-likelihood function of the kth symbol, denoted as $\Lambda^{(\text{opt})}(d_{k1}, d_{k0})$, is given as follows

$$\begin{aligned}\Lambda^{(\text{opt})}(d_{k1}, d_{k0}) = \sum_{l=0}^{N-1} \Bigg\{ &\log\left(1 + \left(\frac{d_{k1}+d_{k0}}{2} r_{I,k,l} + \frac{d_{k1}-d_{k0}}{2} r_{Q,k,l}\right)(1-2\epsilon_{I,l})\right) \\ &+ \log\left(1 - \left(\frac{d_{k1}-d_{k0}}{2} r_{I,k,l} - \frac{d_{k1}+d_{k0}}{2} r_{Q,k,l}\right)(1-2\epsilon_{Q,l})\right)\Bigg\} - 2N\log 2\end{aligned} \tag{9}$$

Then the ML detector can be derived as

$$\left(\hat{d}_{k1}, \hat{d}_{k0}\right) = \arg\max_{d_{k1}, d_{k0} = \pm 1} \Lambda^{(\text{opt})}(d_{k1}, d_{k0}) \tag{10}$$

Note that there is logarithmic operation in (9), the complexity of such receiver is extremely high, even implemented by the lookup table.

Since the Gray coding is employed, we can assume that the BER is approximately equal to symbol error probability. Unfortunately, the complexity of calculating the symbol error probability of the optimal monobit receiver is still extremely high. Hence, we use deflection ratio as the performance criteria, not only for its computation simplicity, but also the equivalence between the ML receiver and the optimum receiver in terms of deflection [14].

Define $\lambda_k = \Lambda^{(\text{opt})}(d_{k1} = 1, d_{k0} = 1)$ as the decision statistic of the ML detector. According to [14], the deflection ratio under QPSK modulation with monobit sampling is given as

$$D = \frac{\left[E(\lambda_k | d_{k1} = 1, d_{k0} = 1) - E(\lambda_k | d_{k1} = 1, d_{k0} = -1)\right]^2}{Var(\lambda_k)} \tag{11}$$

After some manipulations, the deflection ratio of the optimal monobit receiver is given as follows

$$D^{(\text{opt})} = 2\frac{\left[\sum\limits_{l=0}^{N-1}\left((1-\epsilon_{I,l}-\epsilon_{Q,l})\log\frac{1-\epsilon_{I,l}}{\epsilon_{I,l}} + (\epsilon_{I,l}-\epsilon_{Q,l})\log\frac{1-\epsilon_{Q,l}}{\epsilon_{Q,l}}\right)\right]^2}{\sum\limits_{l=0}^{N-1}\left(\epsilon_{I,l}(1-\epsilon_{I,l}) + \epsilon_{Q,l}(1-\epsilon_{Q,l})\right)\left(\log^2\frac{1-\epsilon_{I,l}}{\epsilon_{I,l}} + \log^2\frac{1-\epsilon_{Q,l}}{\epsilon_{Q,l}}\right)} \tag{12}$$

From another perspective, the decision statistic λ_k can be treated as a Gaussian random variable using a central limit argument, when N is large. Thus, the BER performance can be approximately estimated as $Q\left(\sqrt{D}\right)$, which makes the deflection ratio a simpler performance criterion. It can be observed that the reception performance will be better if the deflection ratio is bigger.

With the information of the received reference waveform $p_{\text{ref}}(t)$ and the phase difference φ (and hence the channel state information), we can evaluate the deflection ratio of the optimal monobit receiver to obtain a general point of view about its performance.

B. Suboptimal Monobit Receiver

To get the knowledge of $p_{\text{ref}}(t)$, a reference signal estimation based on training sequences is proposed in [19] for BPSK transmission. However, it requires a large lookup table and exhaustive search. The method proposed in [12] only employs bit-level addition and shift operation to recover the reference signal from monobit sampling results. It can be simply realized online. Herein, the main ideas to obtain the suboptimal receiver in [12] are extended to QPSK modulation here, and a practical monobit receiver for QPSK is derived.

The first technique to be used is the Taylor's expansion. Define $w_{I,l} = 1 - 2\epsilon_{I,l}$ and $w_{Q,l} = 1 - 2\epsilon_{Q,l}$. When SNR is small, $\epsilon_{I,l} \approx 0.5$ and $\epsilon_{Q,l} \approx 0.5$, which lead to $w_{I,l} \approx 0$ and $w_{Q,l} \approx 0$. Thus, we can perform a first order Taylor's expansion of the $\log$ functions in (9), according to $\log(1+x) \approx 1 + x$ when $|x| \ll 1$. Considering the constant $-2N\log 2$ in (9) has no effect on the performance of demodulation in the uncoded case, we ignore this constant and derive the linear approximation of (9) as follows

$$\begin{aligned}\Lambda(d_{k1}, d_{k0}) = \sum_{l=0}^{N-1} \Bigg\{ & w_{I,l}\left(\frac{d_{k1}+d_{k0}}{2} r_{I,k,l} + \frac{d_{k1}-d_{k0}}{2} r_{Q,k,l}\right) \\ & - w_{Q,l}\left(\frac{d_{k1}-d_{k0}}{2} r_{I,k,l} - \frac{d_{k1}+d_{k0}}{2} r_{Q,k,l}\right)\Bigg\}\end{aligned} \tag{13}$$

Therefore, in the suboptimal detector given in (10), we replace $\Lambda^{(\text{opt})}(d_{k1}, d_{k0})$ by $\Lambda(d_{k1}, d_{k0})$. Compared with the ML receiver, the suboptimal receiver in (13) only needs addition, which greatly reduces the implementation complexity.

Due to the monobit quantization, the receiver can not obtain the precise reference signal $p_{\text{ref}}(t)$ and the phase difference φ, or equivalently $\epsilon_{I,l}$, $\epsilon_{Q,l}$. Hence, we need to estimate $\epsilon_{I,l}$ and $\epsilon_{Q,l}$, to further estimate $w_{I,l}$ and $w_{Q,l}$. Assume that a sequence of training symbols (say N_t symbols) are used for estimation. Without loss of generality, all symbols in the training sequence are assumed to be $(d_1 = 1, d_0 = 1)$. Then the ML estimates of $w_{I,l}$ and $w_{Q,l}$ can be given as

$$\hat{w}_{I,l} = \frac{1}{N_t}\sum_{k=0}^{N_t-1} r_{I,k,l}, \quad \hat{w}_{Q,l} = \frac{1}{N_t}\sum_{k=0}^{N_t-1} r_{Q,k,l}, \quad 0 \le l < N-1 \tag{14}$$

Replacing $w_{I,l}$ and $w_{Q,l}$ in (13) by $\hat{w}_{I,l}$ and $\hat{w}_{Q,l}$ respectively, the practical monobit receiver without prior CSI is derived.

It is reasonable to plug $\hat{w}_{I,l}$ and $\hat{w}_{Q,l}$ into (9) and obtain a ML receiver without prior CSI. However, the $\log$ operation in the ML receiver will significantly increase the complexity.

Furthermore, the robustness of such receiver is poor when SNR is relatively large, since a small estimation error will lead to a large error in detection. For these reasons, we will only focus the suboptimal but more robust monobit receiver when perfect CSI is unavailable.

In [12], iteration is proved to be efficient for BPSK demodulation. The main idea of such iterative demodulation is using previous decided symbols to refine the weight estimation. This can also be applied to the QPSK reception. First, the iterative demodulation algorithm estimates the wights $\hat{w}_{I,l}$ and $\hat{w}_{Q,l}$ according to (14) using only training symbols. Then the algorithm detects the data symbols based on the estimated weights. After that, these detected data symbols are used to refine the weight estimation, as additional training symbols. The updated weights are further used to demodulate the data symbols again. This goes back and forth till the symbol decisions will not change any more.

Removing the samples with small amplitude can greatly improve the reception performance without increasing implementation complexity, since monobit quantizer is sensitive to additive noise when the signal amplitude is small. The suboptimal monobit receiver for QPSK can employ this skill too, by setting the corresponding weights to zero. Different from the situation in BPSK, the samples either in the I or Q branch should be removed or not according the same threshold. Although the optimal threshold is hard to obtain, the performance is not sensitive to the threshold, which greatly increases the robustness of the receiver.

C. *Performance of Suboptimal Monobit Receiver*

From (14), we can obtain the mean and variance of the weight $\hat{w}_{I,l}$ as follows

$$E\left\{\hat{w}_{I,l}\right\} = 1 - 2\epsilon_{I,l}, \quad Var\left\{\hat{w}_{I,l}\right\} = \frac{4\epsilon_{I,l}\left(1-\epsilon_{I,l}\right)}{N_t} \tag{15}$$

The mean and variance of $\hat{w}_{Q,l}$ can be obtained similarly. With these knowledge, the deflection ratio of the suboptimal monobit receiver can be calculated as follows

$$D = \frac{\left(\sum\limits_{l=0}^{N-1} M\left(\epsilon_{I,l}\right) + M\left(\epsilon_{Q,l}\right)\right)^2}{\sum\limits_{l=0}^{N-1}\left\{M\left(\epsilon_{I,l}\right) + M\left(\epsilon_{Q,l}\right) + 4\left(V\left(\epsilon_{I,l}\right) + V\left(\epsilon_{Q,l}\right)\right)/N_t - 0.5\left(M\left(\epsilon_{I,l}\right) + M\left(\epsilon_{Q,l}\right)\right)^2\right\}} \tag{16}$$

where $M\left(x\right) = \left(1-2x\right)^2$ and $V\left(x\right) = x\left(1-x\right)$. We remark that the deflection ratio of the suboptimal receiver increases with the number of the training symbols, for a specific reference waveform $p_{\text{ref}}\left(t\right)$ and a specific phase difference φ.

When iterative demodulation is employed, the weights are updated during each iteration. To quantify the possible performance gain offered by the iterative demodulation, the deflection ratio after the nth iteration, denoted as D_n, is calculated. It turns out that D_n can be obtained by simply replacing N_t in (16) with $N_{t,n}^{eq}$, which can be regarded as the equivalent number of training symbols. Through some computation, we can derive $N_{t,n}^{eq}$ as follows

$$N_{t,n}^{\text{eq}} = \begin{cases} \frac{(N_t+N_d-N_{e,n-1})^2}{N_t+N_d}, & n \geq 1 \\ N_t, & n = 0 \end{cases} \tag{17}$$

where N_d is the number of data symbols and $N_{e,n-1}$ denotes the number of decision errors after $(n-1)$th iteration. The derivation details can refer to [12].

D. Effect of Phase Difference

From (16), it can be observed that the deflection ratio only depends on $\epsilon_{I,l}$ and $\epsilon_{Q,l}$ when $N_{t,n}^{\text{eq}}$ is fixed. For a specific reference signal $p_{\text{ref}}(t)$, the parameters $\epsilon_{I,l}$ and $\epsilon_{Q,l}$ are only determined by the phase difference φ, which is constant but unknown to the receiver. As a result, the reception performance is affected by the phase difference.

The deflection ratios both in AWGN and fading channels, normalized by the maximum over $\varphi \in [0, 90°]$, are presented in Figure 2. It is observed that the deflection ratio will be bigger if the amplitudes of the I and Q branches are closer to each other, such as the situation when $\varphi = 45°$. This leads to a better reception performance. The deflection ratio decreases significantly when the amplitude difference between the two branches is large, e.g. $\varphi = 0°$, $\varphi = 90°$. Fortunately, the impact of the phase difference is much weaker in fading channel, thanks to the diversity offered by multipath. From this point of view, the suboptimal monobit receiver is more suitable in fading channel.

E. Interface with Error-Control Decoder

Error-control coding is widely used in practical communication systems against noise and fading introduced by the channel. Some powerful modern codes such as turbo, low-density parity-check (LDPC) or convolutional codes usually involve iterative decoding via a message passing algorithm [20], [21], [22]. Messages in the form of log-likelihood ratio (LLR) need to be fed to the decoder, as well as exchanged inside the decoder.

In the coded case, the LLRs of d_{k0} and d_{k1} can be derived from the estimated symbol log-likelihood function given by (13). Note that $\Lambda^{(\text{opt})}(d_{k1}, d_{k0}) \approx \Lambda(d_{k1}, d_{k0}) - 2N \log 2$ and $\Lambda^{(\text{opt})}(d_{k1}, d_{k0}) = \log P(\mathbf{r}_k | d_{k1}, d_{k0})$, we arrive at

$$P(\mathbf{r}_k | d_{k1}, d_{k0}) \approx e^{\Lambda(d_{k1}, d_{k0}) - 2N \log 2} \tag{18}$$

According to probability theories, it is clear that

$$P(\mathbf{r}_k | d_{k0} = \pm 1) = \sum_{d_{k1}} P(d_{k1}) P(\mathbf{r}_k | d_{k1}, d_{k0} = \pm 1) \tag{19}$$

Without considering the iteration between the decoder and the demodulator, we can assume that $P(d_{k1} = 1) = P(d_{k1} = -1)$. Thus, the LLR of the data d_{k0} can be given as follows

$$\Lambda(d_{k0}) = \log \frac{e^{\Lambda(d_{k1}=+1, d_{k0}=+1)} + e^{\Lambda(d_{k1}=-1, d_{k0}=+1)}}{e^{\Lambda(d_{k1}=+1, d_{k0}=-1)} + e^{\Lambda(d_{k1}=-1, d_{k0}=-1)}} \tag{20}$$

Similarly, the LLR of the data d_{k1} is given as

$$\Lambda(d_{k1}) = \log \frac{e^{\Lambda(d_{k1}=+1, d_{k0}=+1)} + e^{\Lambda(d_{k1}=+1, d_{k0}=-1)}}{e^{\Lambda(d_{k1}=-1, d_{k0}=+1)} + e^{\Lambda(d_{k1}=-1, d_{k0}=-1)}} \tag{21}$$

It can be observed that the calculation of the LLR employs exponential and logarithmic operations. Both of them will greatly increase the implementation complexity. Therefore, the max-log approximation well known in coding theory is used here to reduce the complexity. Thus, the suboptimal LLRs after approximation can be given as follows

$$\begin{aligned}\Lambda^{(\text{sub})}(d_{k0}) = & \max(\Lambda(d_{k1} = 1, d_{k0} = 1), \Lambda(d_{k1} = -1, d_{k0} = 1)) \\ & - \max(\Lambda(d_{k1} = 1, d_{k0} = -1), \Lambda(d_{k1} = -1, d_{k0} = -1))\end{aligned} \tag{22}$$

$$\begin{aligned}\Lambda^{(\text{sub})}(d_{k1}) = & \max(\Lambda(d_{k1} = 1, d_{k0} = 1), \Lambda(d_{k1} = 1, d_{k0} = -1)) \\ & - \max(\Lambda(d_{k1} = -1, d_{k0} = 1), \Lambda(d_{k1} = -1, d_{k0} = -1))\end{aligned} \tag{23}$$

Substituting the estimated symbol log-likelihood functions given by (13) into (22) and (23), the LLRs of binary data d_{k0} and d_{k1} can be derived from the monobit samples and training symbols.

When iterative demodulation is implemented, the weights are updated after each iteration using the data sequence. Note that there might be some decision errors, the weight $w_{I,l,n}^{(\text{iter})}$ is no longer the unbiased estimate of $1 - 2\epsilon_{I,l}$. The situation of $w_{Q,l,n}^{(\text{iter})}$ is similar. Fortunately, such estimation error and its effect on iterative decoding are negligible when $N_t + N_d$ is large.

IV. MONOBIT RECEIVERS WITH IQ IMBALANCE

IQ imbalance, caused by analog component imperfections, is a serious issue degrading the reception performance. However, it is unwise to compensate such impairment in the analog domain due to power and area costs. Therefore, economic schemes in the digital domain are desirable. In this section, we first investigate the effect of the IQ imbalance. To mitigate the performance loss without increasing the complexity, monobit receivers based on double training sequences are proposed. Finally, monobit receiver with 8-sector phase quantization is proposed to counter the IQ imbalance, at the price of doubling the implementation complexity.

A. Effect of IQ Imbalances

For an arbitrary branch of the receiver, both the signal and the noise are scaled by the same positive factor, $1+\alpha$ or $1-\alpha$. Since the monobit sampling is insensitive to the amplitude, the quantization results under amplitude imbalance are the same as the ones without amplitude imbalance. Thus, we can conclude that the amplitude imbalance has no impact on the monobit receivers. To simplify the analysis in the following, α is set to be 0.

In a noise-free channel, when $\alpha = 0$ and $\theta = 0$, which means there is no IQ imbalance at the receiver, we have the following relations

$$\begin{aligned} r_{\mathrm{d},I}\left(t|d_{k1}=1, d_{k0}=1\right) &= r_{\mathrm{d},Q}\left(t|d_{k1}=1, d_{k0}=-1\right) \\ r_{\mathrm{d},Q}\left(t|d_{k1}=1, d_{k0}=1\right) &= -r_{\mathrm{d},I}\left(t|d_{k1}=1, d_{k0}=-1\right) \end{aligned} \tag{24}$$

where $r_{\mathrm{d},I}\left(t|d_{k1}, d_{k0}\right)$ and $r_{\mathrm{d},Q}\left(t|d_{k1}, d_{k0}\right)$ denote the signal of the I and Q branches respectively, when the symbol (d_{k1}, d_{k0}) is transmitted. These relations are the bases of the monobit receivers given in last section. In this case, using symbol $(1, 1)$ for training is enough.

When there is phase imbalance at the receiver, relations given in (24) does not hold, which makes (9) no longer the optimal monobit receiver. Thus, the corresponding practical receiver will suffer some performance loss. One particularly bad situation is reception when φ is around 0°. In such case, receiver given by (13) will confuse the symbol $(1, 1)$ with $(1, -1)$ or $(-1, 1)$ with high probability. Hence, monobit receivers combating phase imbalance are desirable.

B. Monobit Receivers Based on Double Training Sequences

It is observed that the noise of the I branch is not independent of the noise in the Q branch if phase imbalance exists. Fortunately, such dependency is weak since θ is usually small. Therefore, we assume they are independent in the following to simplify the analysis. For the same reason, α is set to be 0. Define

$$\begin{aligned}\epsilon_{I,l}^{0} &= Q\left(p_{\text{ref}}\left(lT\right)\cos\left(\varphi+\theta/2\right)\right), \quad \epsilon_{Q,l}^{0} = Q\left(-p_{\text{ref}}\left(lT\right)\sin\left(\varphi-\theta/2\right)\right) \\ \epsilon_{I,l}^{1} &= Q\left(p_{\text{ref}}\left(lT\right)\sin\left(\varphi+\theta/2\right)\right), \quad \epsilon_{Q,l}^{1} = Q\left(p_{\text{ref}}\left(lT\right)\cos\left(\varphi-\theta/2\right)\right)\end{aligned} \tag{25}$$

The log-likelihood function of the kth symbol with phase imbalance is given as follows

$$\begin{aligned}\Lambda_{\text{d}}^{(\text{opt})}\left(d_{k1},d_{k0}\right) = \sum_{i=I,Q}\sum_{l=0}^{N-1}&\left\{\log\left(1+\frac{d_{k1}+d_{k0}}{2}r_{i,k,l}\left(1-2\epsilon_{i,l}^{0}\right)\right)\right. \\ &\left.+\log\left(1+\frac{d_{k1}-d_{k0}}{2}r_{i,k,l}\left(1-2\epsilon_{i,l}^{1}\right)\right)\right\}-2N\log 2\end{aligned} \tag{26}$$

Replacing $\Lambda^{(\text{opt})}\left(d_{k1},d_{k0}\right)$ in (10) by $\Lambda_{\text{d}}^{(\text{opt})}\left(d_{k1},d_{k0}\right)$, the ML monobit detector under phase imbalance is derived.

Define $w_{I,l}^{0}=1-2\epsilon_{I,l}^{0}$, $w_{Q,l}^{0}=1-2\epsilon_{Q,l}^{0}$, $w_{I,l}^{1}=1-2\epsilon_{I,l}^{1}$ and $w_{Q,l}^{1}=1-2\epsilon_{Q,l}^{1}$. Similar to the monobit ML receiver without phase imbalance, we can perform a Taylor's expansion of (26), discard the constant $2N\log 2$ and obtain a suboptimal but practical monobit receiver under phase imbalance as follows

$$\Lambda_{\text{d}}^{(\text{sub})}\left(d_{k1},d_{k0}\right) = \sum_{i=I,Q}\sum_{l=0}^{N-1}\left\{\frac{d_{k1}+d_{k0}}{2}r_{i,k,l}\left(1-2\epsilon_{i,l}^{0}\right)+\frac{d_{k1}-d_{k0}}{2}r_{i,k,l}\left(1-2\epsilon_{i,l}^{1}\right)\right\} \tag{27}$$

To estimate the weights $w_{I,l}^{0}$, $w_{Q,l}^{0}$, $w_{I,l}^{1}$ and $w_{Q,l}^{1}$, we employ double training sequences. The first training sequence consists of N_t^0 symbols of $(1,1)$ and the second sequence consists of N_t^1 symbols of $(1,-1)$. Let $N_t = N_t^0+N_t^1$ to maintain the system efficiency. Usually, we have $N_t^0=N_t^1=N_t/2$. Thus, the weights can be estimated as follows

$$\begin{aligned}\hat{w}_{I,l}^{0} &= \frac{1}{N_t^0}\sum_{k=0}^{N_t^0-1}r_{I,k,l}, \quad \hat{w}_{Q,l}^{0} = \frac{1}{N_t^0}\sum_{k=0}^{N_t^0-1}r_{Q,k,l} \\ \hat{w}_{I,l}^{1} &= \frac{1}{N_t^1}\sum_{k=0}^{N_t^0-1}r_{I,k,l}, \quad \hat{w}_{Q,l}^{1} = \frac{1}{N_t^1}\sum_{k=0}^{N_t^0-1}r_{Q,k,l}\end{aligned} \tag{28}$$

Substituting the estimated weights into (27), the practical monobit receiver under phase imbalance without prior CSI is obtained.

To evaluate the performance of the monobit receiver given by (27), the deflection ratio of such receiver is calculated as follows

$$D_{\text{d}}^{(\text{sub})} = \frac{4\left[\sum\limits_{l=0}^{N-1}\left(1-2\epsilon_{I,l}^{0}\right)\left(\epsilon_{I,l}^{1}-\epsilon_{I,l}^{0}\right)+\left(1-2\epsilon_{Q,l}^{0}\right)\left(\epsilon_{Q,l}^{1}-\epsilon_{Q,l}^{0}\right)\right]^2}{Var\left(\lambda_{\text{d}}^{(\text{sub})}\right)} \tag{29}$$

where $\lambda_{\mathrm{d}}^{(\mathrm{sub})} = \Lambda_{\mathrm{d}}^{(\mathrm{sub})}(1,1)$ is the decision statistic and its variance is given as follows

$$Var\left(\lambda_{\mathrm{d}}^{(\mathrm{sub})}\right) = \sum_{l=0}^{N-1}\left\{4\left[\epsilon_{I,l}^{0}\left(1-\epsilon_{I,l}^{0}\right)+\epsilon_{Q,l}^{0}\left(1-\epsilon_{Q,l}^{0}\right)\right]/N_t^0+\left(w_{I,l}^{0}\right)^2+\left(w_{Q,l}^{0}\right)^2 \right. \\ \left. -0.5\left[\left(w_{I,l}^{0}\right)^4+\left(w_{I,l}^{0}w_{I,l}^{1}\right)^2+\left(w_{Q,l}^{0}\right)^4+\left(w_{Q,l}^{0}w_{Q,l}^{1}\right)^2\right]\right\} \tag{30}$$

Analogously, the decision statistic can also be defined as $\lambda_{\mathrm{d}}^{(\mathrm{sub})} = \Lambda_{\mathrm{d}}^{(\mathrm{sub})}(1,-1)$ and the corresponding result is similar. It can be observed that if we increase both N_t^0 and N_t^1, the deflection ratio increases and the performance gets better.

To improve the reception performance, we need to increase the equivalent number of training sequences without sacrificing system efficiency. Considering the phase imbalance θ is usually small in practice, we can have

$$w_{I,l}^{0} \approx A w_{Q,l}^{1}, \quad w_{Q,l}^{0} \approx B w_{I,l}^{1} \tag{31}$$

where A and B, called sign factors, represent the sign relations between weights. They are asserted to be either 1 or -1, determined by θ and φ. Due to the absence of the precise θ and φ, the sign factors need to be estimated from training sequences. Note that calculating the ML estimates of the sign factors is extremely complex, a simple but effective estimation method is proposed as follows

$$\hat{A} = \mathrm{sgn}\left(\sum_{l=0}^{N-1}\hat{w}_{I,l}^{0}\hat{w}_{Q,l}^{1}\right), \quad \hat{B} = \mathrm{sgn}\left(\sum_{l=0}^{N-1}\hat{w}_{Q,l}^{0}\hat{w}_{I,l}^{1}\right) \tag{32}$$

Once the sign factors have been estimated, the weights estimated from training sequences can be combined as follows

$$\hat{w}_{I,l}^{(\mathrm{cw})} = \frac{1}{2}\hat{w}_{I,l}^{0} + \frac{\hat{A}}{2}\hat{w}_{Q,l}^{1}, \quad \hat{w}_{Q,l}^{(\mathrm{cw})} = \frac{1}{2}\hat{w}_{Q,l}^{0} + \frac{\hat{B}}{2}\hat{w}_{I,l}^{1} \tag{33}$$

If the sign factors are estimated correctly, the combinational weights will have smaller variances compared with the original weights. Representing the original weights with the combinational weights and substituting them into (27), we can obtain a monobit receiver with combinational weights as follows

$$\Lambda_{\mathrm{d}}^{(\mathrm{cw})}(d_{k1},d_{k0}) = \sum_{l=0}^{N-1}\left\{\frac{d_{k1}+d_{k0}}{2}\left(\hat{w}_{I,l}^{(\mathrm{cw})}r_{I,k,l}+\hat{w}_{Q,l}^{(\mathrm{cw})}r_{Q,k,l}\right) \right. \\ \left. +\frac{d_{k1}-d_{k0}}{2}\left(\hat{B}\hat{w}_{Q,l}^{(\mathrm{cw})}r_{I,k,l}+\hat{A}\hat{w}_{I,l}^{(\mathrm{cw})}\right)\right\} \tag{34}$$

To evaluate the possible performance gain offered by such receiver, its deflection ratio, denoted as $D_{\text{d}}^{(\text{cw})}$, is calculated. The decision statistic is defined as $\lambda_k = \Lambda_{\text{d}}^{(\text{cw})}(1,1)$. After some manipulations, the numerator and denominator of $D_{\text{d}}^{(\text{cw})}$ are given as

$$\begin{aligned} D_{\text{d,num}}^{(\text{cw})} &= \left(w_{I,l}^0\right)^2 + \left(w_{Q,l}^0\right)^2 + (A+B)\, w_{I,l}^1 w_{Q,l}^1 \\ &\quad + A w_{I,l}^0 w_{Q,l}^1 + B w_{I,l}^1 w_{Q,l}^0 + w_{I,l}^0 w_{I,l}^1 + w_{Q,l}^0 w_{Q,l}^1 \end{aligned} \tag{35}$$

$$\begin{aligned} D_{\text{d,den}}^{(\text{cw})} &= \left[V\left(\epsilon_{I,l}^0\right) + V\left(\epsilon_{Q,l}^0\right)\right]/N_t^0 + \left[V\left(\epsilon_{I,l}^1\right) + V\left(\epsilon_{Q,l}^1\right)\right]/N_t^1 \\ &\quad + 0.5\left(w_{I,l}^0 + A w_{Q,l}^1\right)^2 \left[V\left(\epsilon_{I,l}^0\right) + V\left(\epsilon_{I,l}^1\right)\right] \\ &\quad + 0.5\left(w_{Q,l}^0 + B w_{I,l}^1\right)^2 \left[V\left(\epsilon_{Q,l}^0\right) + V\left(\epsilon_{Q,l}^1\right)\right] \end{aligned} \tag{36}$$

and the deflection ratio is given by $D_{\text{d}}^{(\text{cw})} = D_{\text{d,num}}^{(\text{cw})}/D_{\text{d,den}}^{(\text{cw})}$. Through some simple comparison, we can find that $D_{\text{d}}^{(\text{cw})} > D_{\text{d}}^{(\text{sub})}$, which means that the receiver with combinational weights can outperform the receiver given by (27). This will also be proved in the numerical results.

C. Monobit Receiver with 8-Sector Phase Quantization

To further improve monobit reception, more complex and sophisticated strategies are need. Compared with the monobit quantization under traditional receiver architecture, the 8-sector phase quantization proposed in [13] can provide more precise phase information of the received signal, at the price of two extra analog adders and monobit ADCs. By adding the I+Q and I-Q branches to the traditional monobit receiver, the 8-sector phase quantization can simply be implemented. Next, the receivers with 8-sector phase quantization are obtained.

Define $\epsilon_I = [\epsilon_{I,0}, ..., \epsilon_{I,N-1}]^T$ and $\epsilon_Q = [\epsilon_{Q,0}, ..., \epsilon_{Q,N-1}]^T$. To simplify the notation, we can let $\Lambda^{(\text{opt})}(d_{k1}, d_{k0}|\epsilon_I, \epsilon_Q) = \Lambda^{(\text{opt})}(d_{k1}, d_{k0})$ and $\Lambda(d_{k1}, d_{k0}|\epsilon_I, \epsilon_Q) = \Lambda(d_{k1}, d_{k0})$, where $\Lambda^{(\text{opt})}(d_{k1}, d_{k0})$ and $\Lambda(d_{k1}, d_{k0})$ are given by (9) and (13) respectively. If there is no IQ imbalance at the receiver, the optimal monobit receivers under 8-sector phase quantization is given as follows

$$\Lambda_{\text{phq}}^{(\text{opt})}(d_{k1}, d_{k0}) = \Lambda^{(\text{opt})}(d_{k1}, d_{k0}|\epsilon_I, \epsilon_Q) + \Lambda^{(\text{opt})}(d_{k1}, d_{k0}|\epsilon_{I-}, \epsilon_{I+}) \tag{37}$$

where ϵ_{I-} and ϵ_{I+}, defined similarly as ϵ_I and ϵ_Q, are parameter vectors of the I-Q and I+Q branches respectively. The corresponding suboptimal monobit receiver can be given as

$$\Lambda_{\text{phq}}^{(\text{sub})}(d_{k1}, d_{k0}) = \Lambda(d_{k1}, d_{k0}|\epsilon_I, \epsilon_Q) + \Lambda(d_{k1}, d_{k0}|\epsilon_{I-}, \epsilon_{I+}) \tag{38}$$

For such receiver, the weights of all four branches need to be estimated based on the training sequence. Besides, the iterative demodulation and removal of small-weights points are still useful.

When there is IQ imbalance at the receiver, the ML monobit receiver is much more complex. Thanks to the phase information offered by phase quantization, the effect of the IQ imbalance is much weaker. As we will see in the numerical results, the suboptimal receiver proposed in (38) is enough to combat the SNR loss caused by IQ imbalance.

V. Numerical Results

A. Simulation Parameters

The receivers obtained in this paper are particularly suitable to process large wideband signals, such as UWB or 60G signals. The main reason is that the Nyquist sampling rates of these signals are up to multi giga hertz. At such high rate, high-precision ADC is extremely costly and power-hungry. Consequently, monobit receivers are very attractive.

In wideband communication system, we usually use raised cosine pulse, which is given as

$$p_{\text{tr}}\left(t\right) = \text{sinc}\left(t/\tau\right) \frac{\cos\left(\pi\beta t/\tau\right)}{1-4\beta^2 t^2/\tau^2} \tag{39}$$

where τ is the time constant that controls the pulse duration, and β is the roll-off factor.

In the following, simulation results are provided to evaluate the performance of the monobit receivers proposed in this paper. The raised cosine pulse was adopted with $\tau = 0.5$ns and $\beta = 1$. The bandwidth of the low-pass filter was B = 5GHz, and the sampling rate was Nyquist rate T = 100ps. We assume that there is no ISI and the timing is perfect. The number of data symbols was $N_d = 1000$, and the length of the training sequence was $N_t = 100$. The training overhead amounted to 10 percent of the total transmission duration. The SNR is defined as $E_b/N_0 = \sum_{l=0}^{N-1} p_{\text{ref}}^2\left(lT\right)$. For the simulations in fading channels, we used standard CM1 channel model [23] in UWB for the dense multipath case, and CM14 channel model [24] in 60G for the sparse multipath case. Both of the two channel models were simulated with 100 realizations.

B. Receiver Considered

To simplify the notation in the next discussion, we use abbreviations. The first part of the abbreviation is either FR, MB or PQ, indicating whether full-resolution ADC, monobit sampling or 8-sector phase quantization is used. The second part is either E or F indicating whether estimated or full CSI is used in obtaining the weighting signal for detection. The third part is one of ML, MF, TE, indicating the type of weighting method used, corresponding to the optimal weights in (9), the matched-filter weights, the suboptimal weights in (13) obtained from Taylor's expansion. For the simulations with the IQ imbalances, the abbreviations DT and CW indicate the weights in (28) and the combinational weights in (33) respectively, based on the double training sequences. Finally, the suffix IR indicates the iterative demodulation with removing small-weight samples, and SI indicates the sign factors are available at the receiver. Use such notation, the receivers that we will consider are as follows:

1) FR-F-MF: the optimal receiver with full-resolution sampling, full CSI, and matched-filter weights.

2) MB-F-ML: the optimal receiver with monobit sampling, full CSI, and optimal weights in (9).

3) MB-F-MF: the monobit receiver with full CSI and the matched filter weights.

4) MB-E-TE: the monobit receiver with estimated CSI, Taylor's expansion approximated weights.

5) MB-E-TE-IR: MB-E-TE receiver with removal of small-weight points and iterative demodulation.

6) MB-F-MF-SI: the monobit receiver with full CSI, the sign factor information, and matched filter weights.

7) MB-E-DT: the monobit receiver with estimated CSI, Taylor's expansion approximated weights based on double training sequences.

8) MB-E-DT-IR: MB-E-DT receiver with iterative demodulation and removal of small amplitude samples.

9) MB-E-CW: the monobit receiver with estimated CSI, combinational weights based on double training.

10) MB-E-CW-IR: MB-E-CW receiver with iterative demodulation and removal of small-weight points.

C. *Numerical Results*

Figure 3 compares the performance of different receivers in AWGN channel without IQ imbalances. Given the perfect reference signal (which is not possible in practice though), the MB-F-MF and MB-F-ML receivers have similar performance in entire SNR range. Both of them have about 5dB SNR loss to the full-resolution matched filter when the BER is around 10^{-3}. Compared to the MB-E-TE receiver, the MB-E-TE-IR receiver can provide about 3dB performance gain, with only 3dB SNR gap from the monobit receiver with full CSI. By doubling the processing complexity in digital domain, the PQ-E-TE-IR receiver has 2dB SNR loss compared with the full-resolution matched filter.

The performance under CM1 channel is shown in Figure 4. Similar to the AWGN channel, the MB-F-MF and MB-F-ML receivers have almost the same performance, which is about 2dB SNR loss to the FR-F-MF receiver. The MB-E-TE-IR receiver still outperforms the MB-E-TE receiver about 3dB performance gain, and has only 3dB SNR loss compared with the FR-F-MF receiver. It is also observed that the MB-E-TE-IR receiver can perform as well as the PQ-E-TE-IR receiver with less than 1dB SNR loss, thanks to the diversity of the dense multipath. The performance of different receivers in CM14 channel is given in Figure 5. In such sparse multipath channel, the PQ-E-TE-IR receiver can provide much better performance than the MB-E-TE-IR receiver, with double the complexity. It has only 2dB SNR loss to the full-resolution matched filter.

Figure 6 shows the effect of the phase difference on the practical monobit receivers in AWGN and CM1 channels without IQ imbalances. The numbers suffixed to the abbreviations of the receivers indicate the degree of the phase difference, e.g. $\varphi = 0°$. It can be observed that the MB-E-TE-IR receiver is greatly affected by the phase difference in AWGN channel. The reception performance is much better when the amplitudes of the I and Q branches are closer to each other ($\varphi = 45°$), which is consistent with the results derived from the deflection ratio in Figure 2. On the other hand, the PQ-E-TE-IR receiver can avoid being affected by the phase difference, due to the more precise phase information of the received signal it has. It also shows that the effect of the phase difference on the MB-E-TE-IR receiver in dense multipath channel is weak.

Figure 7 presents the performance of different LLR approximations of the MB-E-TE-IR receiver in AWGN and CM1 channels in the coded case. We employed convolutional code with 1/2 rate. The abbreviation LLR-Opt, LLR-Sub or LLR-Hard indicates the optimal LLR in (20)-(21), the approximate LLR in (22)-(23) or the hard decoding, respectively. It shows that the soft decoding using LLR can offer much better performance than the hard decoding, both in AWGN channel and fading channel. It is observed that the decoding using approximate LLR has almost the same performance as the one using optimal LLR. Thus, the approximate LLR with lower complexity is quite enough for practical communication systems.

Figure 8 gives the performance of different receivers in AWGN channel with amplitude imbalance $\alpha = 0.1$ and phase imbalance $\theta = 2.5°$. When double training sequences are used, we set $N_t^0 = N_t^1 = N_t/2 = 50$ to maintain the system efficiency. It shows that monobit receiver without the information of the sign factors, such as MB-F-MF or MB-E-TE-IR, has a 10^{-2} error floor at high SNR region. After estimated the sign factors, the MB-E-DT-IR and MB-E-CW-IR receivers eliminate such error floor. However, both of them have a BER upturn when SNR is in 25-40dB. This is caused by the side-lobes of the pulse and the IQ imbalance. The MB-E-CW-IR receiver outperforms the MB-E-DT-IR receiver as analyzed before. It is observed that the PQ-E-TE-IR receiver can completely eliminate the effect of the IQ imbalance.

Figure 9 presents the performance of different receivers with IQ imbalances in CM1 channel. The parameters of the IQ imbalance are the same as the ones in AWGN channel. Thanks to the diversity offered by the dense multipath, all receivers have almost no SNR loss compared with the performance without IQ imbalances in Figure 4, except for the MB-E-DT-IR receivers whose performance is limited by the equivalent number of the training sequence. The MB-E-CW-IR and MB-E-TE-IR receivers have almost the same performance. The PQ-E-TE-IR receiver has only about 1dB SNR gain to the MB-E-TE-IR or MB-E-CW-IR receiver, which is uneconomical compared to its complexity increasing. It can be observed that the effect of the IQ imbalance at the receiver in dense multipath channel is negligible. The performance of different receivers with IQ imbalances in CM14 channel is given in Figure 10. In such sparse multipath channel, we can observe that the PQ-E-TE-IR receiver can provide considerable performance gain. As a result, the tradeoff between the performance and the complexity need to be made.

VI. Conclusions

We have derived the optimal monobit receiver for QPSK and its performance in the form of deflection ratio. To reduce the implementation complexity, we extended the main ideas of [12] here to obtain a suboptimal monobit receiver for QPSK. We investigated the effect of the phase difference between the transmitter and the receiver. The interface with error-control decoder was also derived for such receiver. The simulation results showed that such receiver greatly reduce the complexity with about 3dB SNR loss in AWGN channel and 1dB SNR loss in fading channel, compared with the matched-filter based monobit receiver with full CSI.

We have also investigated the effect of the IQ imbalances at the receiver. Monobit receivers based on double training sequences are proposed to counter the performance loss caused by IQ imbalances, without increasing the complexity. Moreover, monobit receiver with 8-sector phase quantization is proposed to completely eliminate the effect of IQ imbalances. It is proved that the amplitude imbalance has no effect on monobit receivers. We noticed that the proposed monobit receiver can efficiently compensate for the SNR loss in AWGN channel, especially when SNR is high. The SNR loss of all these receivers in fading channel is acceptable, thanks to the diversity offered by the multipath.

For cost and complexity consideration, the digital receivers with monobit sampling are strong candidates for future communication systems with significantly large bandwidth, such as UWB communication or communication in 60G band. There are several open issues to be addressed, such as evaluating the performance of the monobit receiver under QAM modulation, the impact of the IQ imbalances at the transmitter, and monobit reception in frequency asynchronous system.

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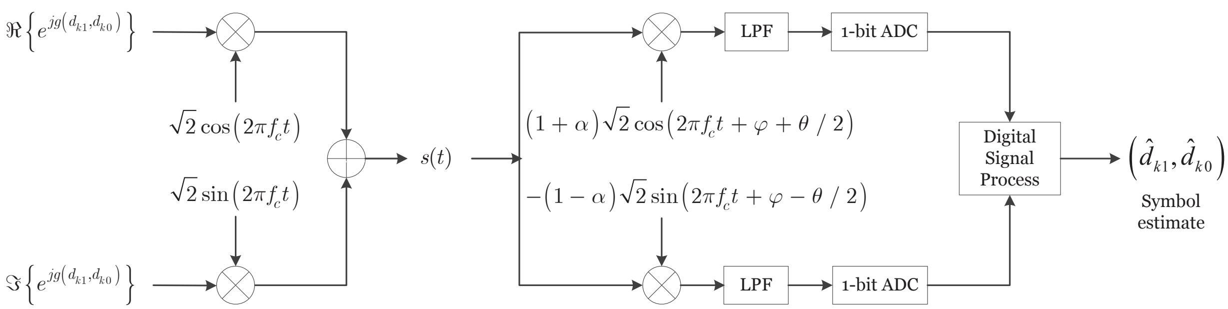

Fig. 1. Monobit receiver architecture for QPSK

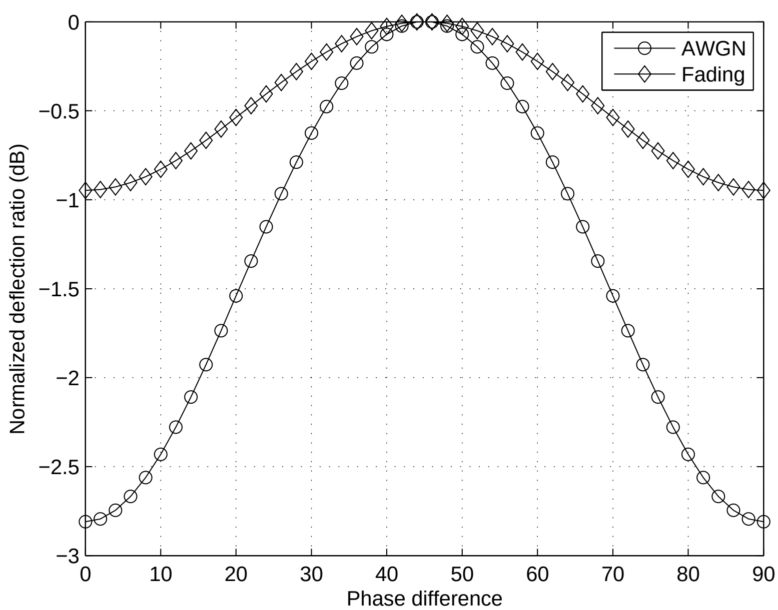

Fig. 2. Normalized deflection ratio under difference phase difference φ

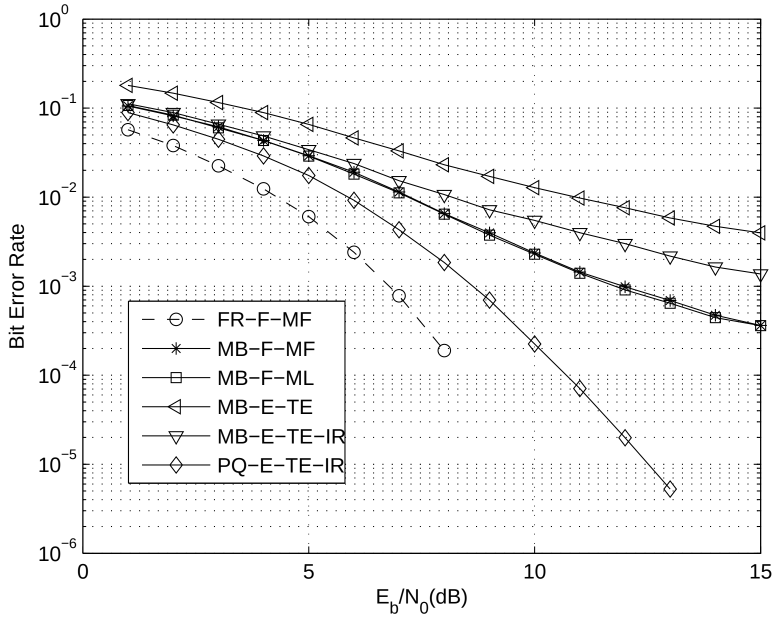

Fig. 3. Comparison of performance of optimal, suboptimal, and full-resolution, phase-quantization, monobit receivers in AWGN channel under Nyquist sampling without IQ imbalances

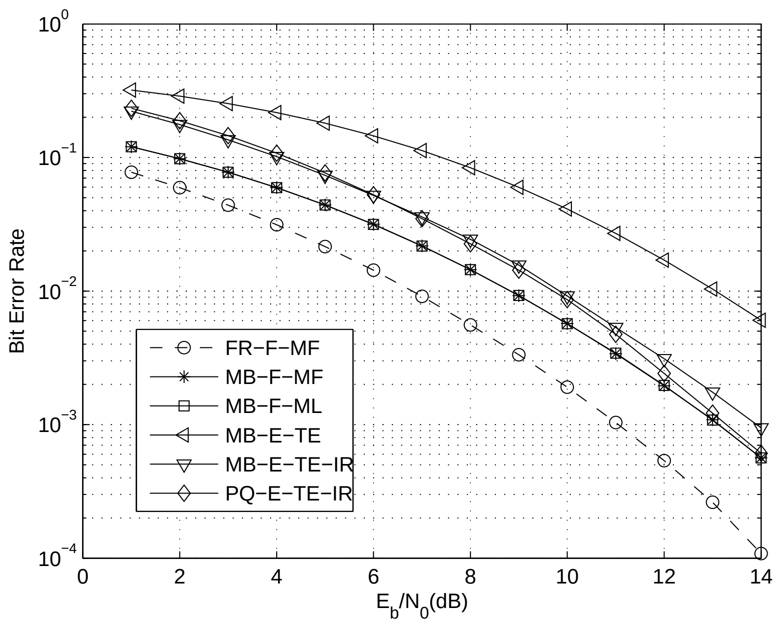

Fig. 4. Comparison of performance of optimal, suboptimal, and full-resolution, phase-quantization, monobit receivers in dense multipath channel under Nyquist sampling without IQ imbalances

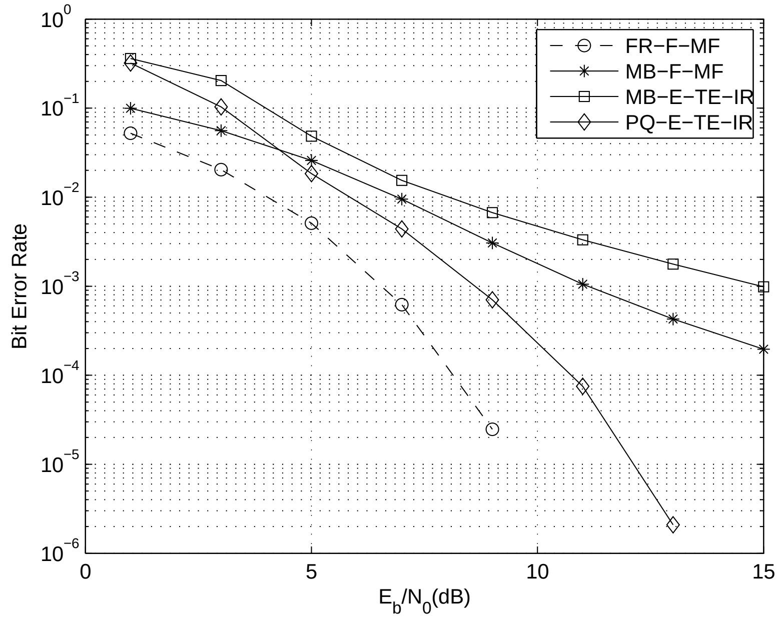

Fig. 5. Comparison of performance of suboptimal, and full-resolution, phase-quantization, monobit receivers in sparse multipath channel under Nyquist sampling without IQ imbalances

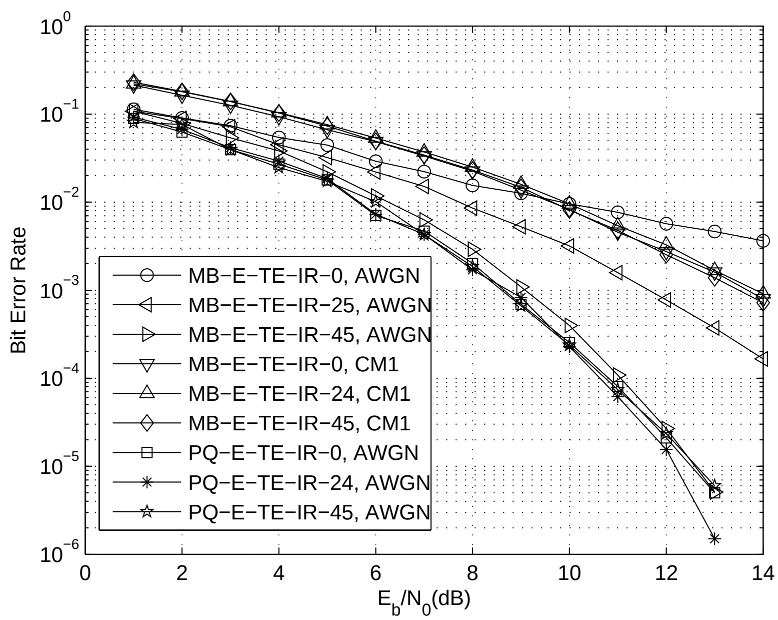

Fig. 6. Effect of phase difference on suboptimal monobit receivers under monobit, phase-quantization sampling without IQ imbalances

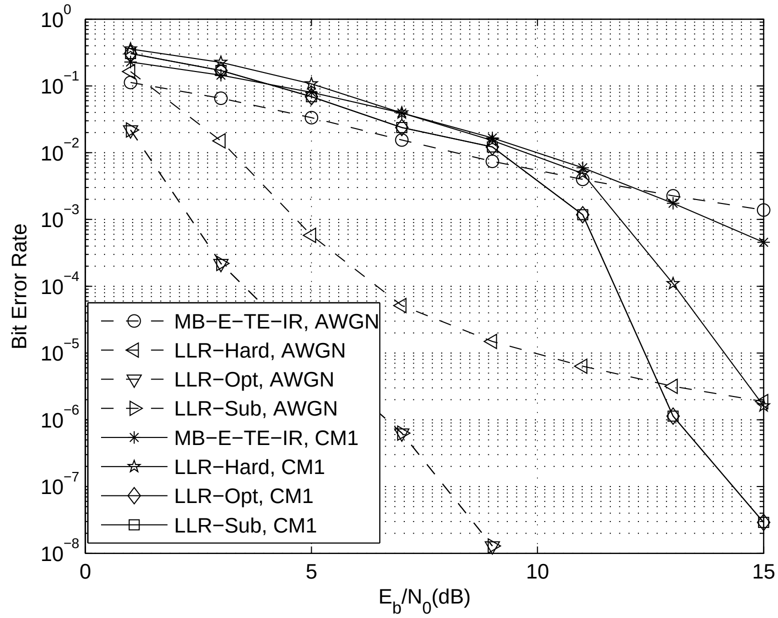

Fig. 7. Comparison of different LLR approximations for suboptimal monobit receiver in AWGN and CM1 channels without IQ imbalances

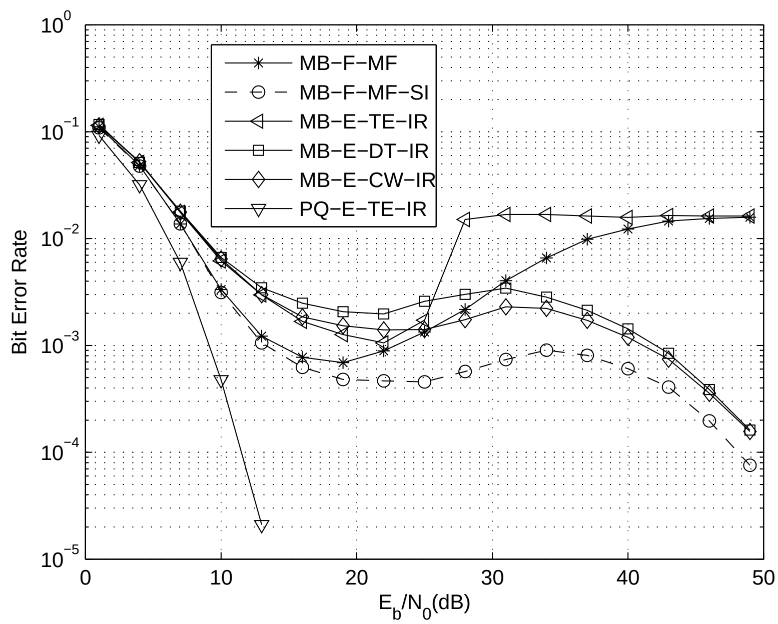

Fig. 8. Comparison of performance of different monobit receivers in AWGN channel with IQ imbalances

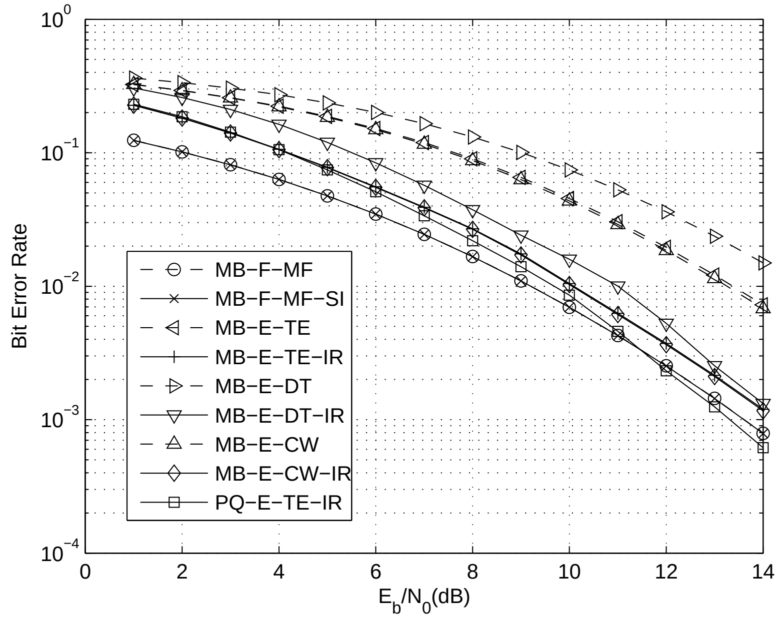

Fig. 9. Comparison of performance of different monobit receivers in dense multipath channel with IQ imbalances

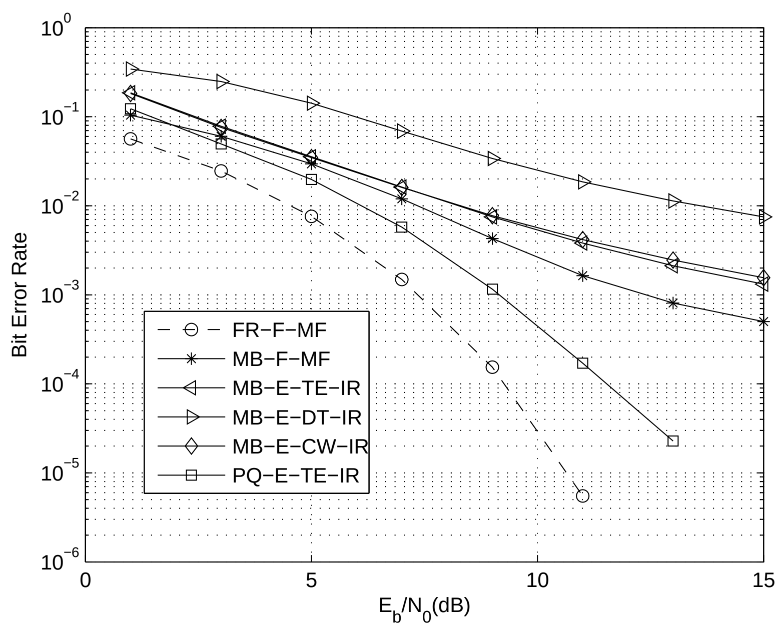

Fig. 10. Comparison of performance of different monobit receivers in sparse multipath channel with IQ imbalances