Decision feedback-based demodulation and adaptive equalization for Doubly Differential PSK

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Abstract

We extended the decision feedback-based demodulation and equalization techniques for differential phase shift keying (DPSK) to the demodulation and equalization of doubly differential PSK (DDPSK). The decision feedback demodulator for DDPSK comprises two consecutive differential detectors which are based on some reference signals obtained from previously detected and received data. The proposed equalizer is a hybrid of this decision feedback demodulator and a linear feedback equalizer. By modifying the equalizer output based on the decision feedback demodulation before feeding back into its feedback part, the proposed equalizer can act like an equalizer with decision feedback structure. Computer simulation results demonstrate that the performances of the decision feedback demodulation and the equalizer are comparable to those of the coherent receiver and the decision feedback equalizer (DFE) with coherent detection, respectively.

I. Introduction

In an attempt to alleviate the need for frequency tracking loops in DPSK receivers, the DDPSK that employs second-order differential encoding of the data has been proposed [1]-[4]. The DDPSK modulation, together with second-order differentially coherent detection, has shown to attenuate the degrading effects on performance due to both phase and frequency offset. This robustness to asynchronous carrier, however, is achieved at the expense of performance. It has been observed that in the absence of frequency offset the performance degradation with second-order differentially coherent DDPSK can exceed $4dB$ as compared to conventional differentially coherent DPSK [2],[3]. To improve the performance of DDPSK systems, multiple-symbol-based detection schemes were proposed in [2],[3]. These methods can improve the receiver performance, but their implementation requires heavy computation.

In this paper, we first show that the decision feedback demodulation for DPSK [5]-[8] can be extended to the demodulation of DDPSK. It will be shown that the performance of this decision feedback demodulation for DDPSK is comparable to that of coherent detection of DDPSK signal, and that its implementation is considerably simpler than the one in [2]. In the second part of this paper, we consider the equalization of DDPSK signals, for which little work has been reported. The proposed equalizer is an extension of the equalizer in [10], originally proposed for DPSK systems. This equalizer is a hybrid of the decision feedback demodulator and the linear feedback equalizer. By modifying the equalizer output based on the decision feedback demodulation before feeding back into its feedback part, the proposed equalizer can perform like a decision feedback equalizer with coherent detection. This is demonstrated through computer simulation.

II. Decision Feedback-based Demodulation and Equalization for DPSK Systems

In this section, we will briefly review the decision feedback-based demodulation and equalization techniques for DPSK, introducing our notations. Fig.1 illustrate the decision feedback DPSK demodulator in [5]-[8]. At the transmitter, the information symbols $a(k)$ are differentially encoded into the symbols

$$d(k) = a(k)d(k-1) = d(0)\prod_{i=1}^{k} a(i). \qquad (1)$$

Assuming an AWGN channel, the received signal $r(k)$ which is the output of the matched filter sampled with symbol period T is written as

$$r(k) = e^{j\theta} d(k) + \eta(k) \qquad (2)$$

where θ is the phase offset and $\eta(k)$ represents additive white Gaussian noise. In the decision feedback demodulation, the phase reference at k is given by $\arg\{\tilde{d}_n(k)\}$ where $\tilde{d}_n(k) = \tilde{d}(k)/|\tilde{d}(k)|$ and

$$\tilde{d}(k) = r(k) + \sum_{i=1}^{M-1}\left\{\prod_{j=0}^{i-1}\tilde{a}(k-j)\right\} r(k-i). \qquad (3)$$

Here $\tilde{a}(k)$ is the decision and M is a positive integer which is usually set at about $3 \sim 5$. Note that if we disconnect the feedback part and remove the $\tilde{d}_n(k)$ calculation block, then this demodulator reduces to the conventional differentially coherent demodulator.

Fig. 2 illustrates the baseband model for the decision

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feedback-based equalizer [10], which is a combination of the linear feedback equalizer and the decision feedback demodulator. This equalizer is called the *modified linear feedback equalizer(MLFE).* The input to the equalizer is represented as

$$r(k) = e^{j\theta} \sum_{i=-\infty}^{\infty} h(i)d(k-i) + \eta(k) \quad (4)$$

where θ and $\eta(k)$ are those defined in (2) and $\{h(i)\}$ is the overall channel impulse response that includes the transmitter filter, the channel, and the receiver filter. The output of the linear feedback equalizer in Fig. 2 is inputted to the decision feedback demodulator to produce the decision $\tilde{a}(k)$. The error signal $e(k)$ is obtained for equalizer tap adaptation. The values $\tilde{d}_n(k)$, from which the phase reference is obtained, are fed back into the feedback part of the linear feedback equalizer. When the least mean square (LMS) algorithm is used for tap adaptation, this system is specified by the following equations:

$$x(k) = W_k^t Q_k, \quad (5a)$$

$$e(k) = \tilde{a}(k) - x(k)\tilde{d}_n^*(k-1), \quad (5b)$$

$$W_{k+1} = W_k + \mu e(k)\tilde{d}_n(k-1)Q_k^* \quad (5c)$$

where t means the transpose, $Q_k^t = \{r(k+N_f), \cdots, r(k), \tilde{d}_n(k-1), \cdots, \tilde{d}_n(k-N_b)\}$ with N_f+1 and N_b the number of feedforward and feedback taps, respectively. W_k is the (N_f+N_b+1) dimensional tap coefficient vector at k.

III. Decision Feedback Demodulation for DDPSK.

The DDPSK encoder and the second-order differentially coherent demodulator is illustrated in Fig. 3. The information symbols $a(k)$ are differentially encoded into

$$b(k) = a(k)b(k-1) = b(0)\prod_{i=1}^{k} a(i). \quad (6)$$

The transmitted symbols $d(k)$ are obtained through another differential encoding:

$$d(k) = b(k)d(k-1) = d(0)\prod_{i=1}^{k} b(i). \quad (7)$$

For AWGN channel the received signal $r(k)$ is represented as

$$r(k) = e^{j(w_d k+\theta)}d(k) + \eta(k) \quad (8)$$

where w_d is the carrier frequency offset. Note that in DDPSK, we consider both frequency and phase offset. The conventional second-order differentially coherent demodulation producing the decision $\tilde{a}(k)$ is expressed as

$$v(k) = r(k)r^*(k-1) \quad (9)$$

and

$$\tilde{a}(k) = dec\{v(k)v^*(k-1)\} \quad (10)$$

where $dec\{\bullet\}$ denotes a decision operator.

It is straightforward to extend the decision feedback demodulator for DDPSK in Fig. 1 to the case of DDPSK. We simply cascade two decision feedback detectors, as shown in Fig. 4. At the second stage of this demodulator, using M_1 previously detected data $\{\tilde{a}(k)\}$ and the outputs of the first differential detector $\{v(k)\}$, the reference phase at time k is estimated by $\arg\{\tilde{b}_n(k)\}$ where $\tilde{b}_n(k) = \tilde{b}(k)/|\tilde{b}(k)|$ and

$$\tilde{b}(k) = v(k) + \sum_{i=1}^{M_1-1}\left\{\prod_{j=0}^{i-1} a(k-j)\right\}v(k-i). \quad (11)$$

At the first stage, using the estimated phase reference phase signals $\{\tilde{b}_n(k)\}$ and received signal $\{r(k)\}$, the reference phase at time k is given by $\arg\{\tilde{d}_n(k)\}$ where $\tilde{d}_n(k) = \tilde{d}(k)/|\tilde{d}(k)|$ and

$$\tilde{d}(k) = r(k) + \sum_{i=1}^{M_2-1} c^i\left\{\prod_{j=0}^{i-1}\tilde{b}_n(k-j)\right\}r(k-i) \quad (12)$$

where c is a weighting factor. The estimate of $v(k)$ and the final decision $\tilde{a}(k)$ are given by

$$v(k) = r(k)\tilde{d}_n^*(k-1) \quad (13)$$

and

$$\tilde{a}(k) = dec\{v(k)\tilde{b}_n^*(k-1)\}. \quad (14)$$

To evaluate the performance of this decision feedback demodulator, and compare it with the conventional second-order differentially coherent demodulator, the bit error rate (BER) values for these receivers were empirically estimated through computer simulation. In this simulation $w_d = 0.5\pi$ and $\theta = 1.5\pi$ in (8) were assumed; 10^7 binary input data were generated; and the parameters for the proposed scheme: $M_1 = 20$, $M_2 = 5$, and $c = 0.6$. Fig. 5 shows the BER values. For comparison, the BER associated with the coherent demodulator with doubly differential decoding is also shown. The decision feedback demodulator performed considerably better than the conventional second-order differentially coherent demodulator in Fig. 3 (b). The proposed demodulator offers a power gain of about $4dB$ when BER is 10^{-2}, and is almost comparable to the coherent demodulation.

IV. The modified linear feedback equalizer for DDPSK.

The MLFE for DDPSK shown in Fig. 6 is a direct extension of the MLFE in Fig. 2. This equalizer is a hybrid of the decision feedback demodulation in Fig. 4 and a linear feedback equalizer. The received signal $r(k)$ is expressed as

$$r(k) = e^{j(w_d k+\theta)} \sum_{i=-\infty}^{\infty} h(i)d(k-i) + \eta(k) . \qquad (15)$$

This expression is the same as the one in (4) with the exception of the frequency offset w_d. If we use the least mean square (LMS) algorithm for tap adaptation, the system is specified by the following equations:

$$x(k) = W_k^t Q_k , \qquad (16.a)$$

$$e(k) = \tilde{a}(k) - x(k)\tilde{d}_n^*(k-1)\tilde{b}_n^*(k-1) , (16.b)$$

$$W_{k+1} = W_k + \mu e(k) Q_k^* \tilde{d}_n(k-1)\tilde{b}_n(k-1) . (16.c)$$

The performance of the MLFE is examined through computer simulation, and compared with that of the DFE with coherent demodulation. In our simulation, we considered two linear phase, finite impulse response (FIR) channels in [9]. They are channels A and B, specified by $\{h_0, h_1, h_2\} = \{0.304, 0.903, 0.304\}$ and $\{0.407, 0.815, 0.407\}$, respectively. Channel B has a spectral null in its frequency response and causes more severe ISI. Each equalizer has 15 taps with $N_f = 9$ and $N_b = 5$. we set $w_d = 0.5\pi$ and $\theta = 1.5\pi$. The step size for tap adaptations is set at $\mu = 0.005$; and the parameters for the decision feedback demodulator are: $M_1 = 20$, $M_2 = 5$, and $c = 0.6$. The tap coefficients were initially obtained by using 10^4 training data. The BER values were empirically estimated by processing 10^7 binary input data. Figs. 7 and 8 show the BER values for channels A and B, respectively. It is seen that for both channels, the MLFE which is based on differentially coherent detection is comparable to the DFE with coherent detection.

IV. Summary

A decision feedback demodulator and an adaptive equalizer for the detection of DDPSK signals have been proposed. The demodulator and equalizer, named as decision feedback demodulator and modified linear feedback equalizer respectively, are the extensions of demodulation and equalization techniques to the cases of DDPSK. Through computer simulations, it was seen that the performances of the decision feedback demodulation and the equalizer were comparable to those of the coherent receiver and the DFE/coherent, respectively, without dependency on carrier frequency and phase offset.

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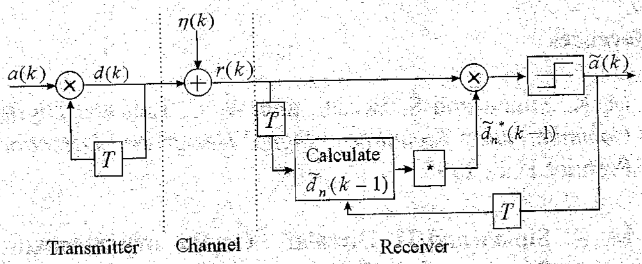

Fig. 1 The decision feedback demodulation for DPSK

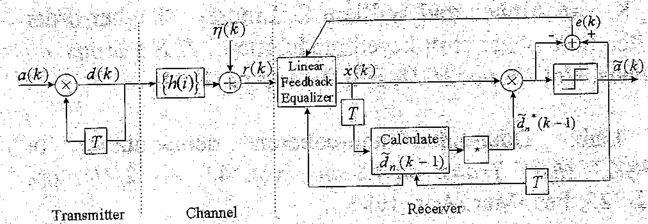

Fig 2. The modified linear feedback equalizer for DPSK

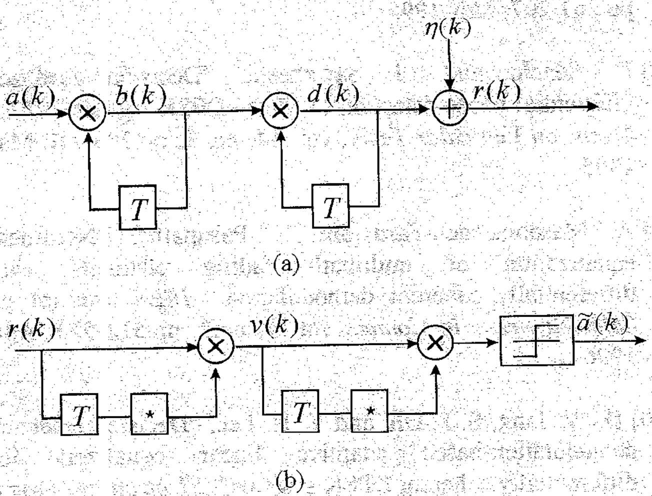

Fig 3. (a) Encoder for DDPSK, (b) Second order differentially coherent demodulator

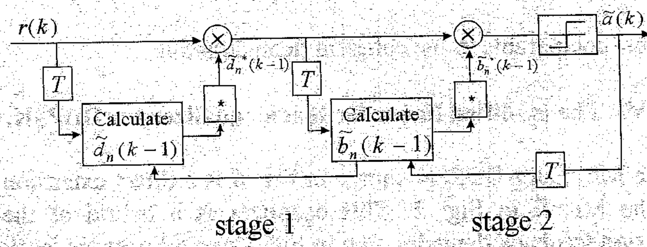

Fig 4. The decision feedback doubly differential demodulator

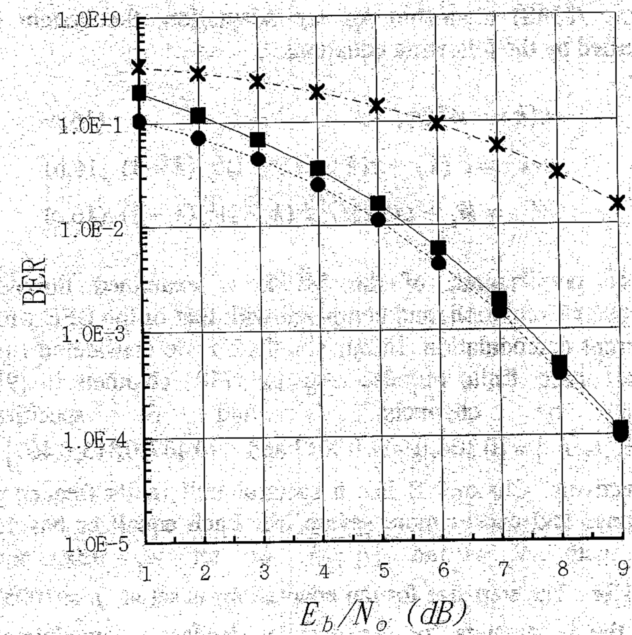

Fig. 5. Bit error probability versus E_b / N_o for AWGN channel (■ decision feedback demodulation, * conventional second-order differentially coherent demodulation , ● coherent demodulation)

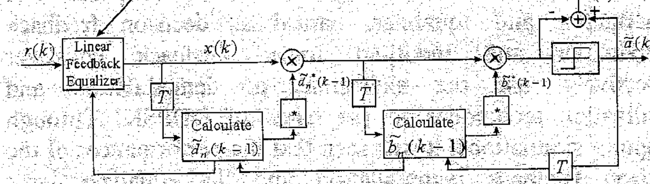

Fig 6. The modified linear feedback equalizer for second order differentially coherent PSK

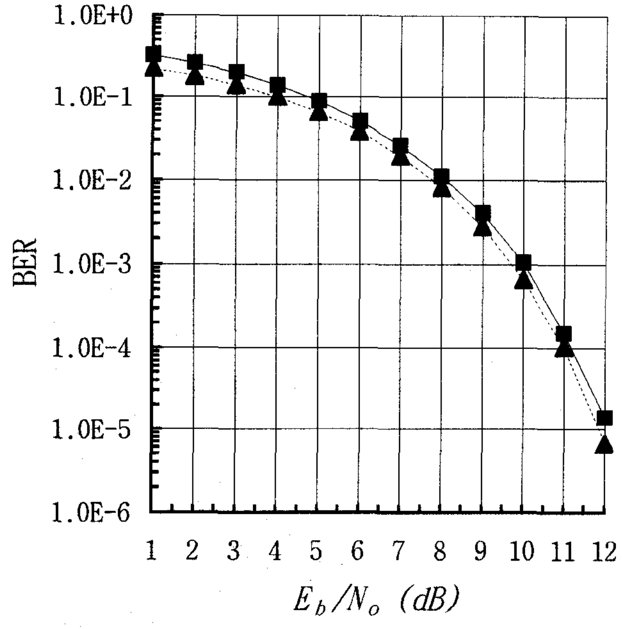

Fig. 7. Bit error probability versus E_b / N_o for channel A (■ MLFE, ▲ DFE/coherent)

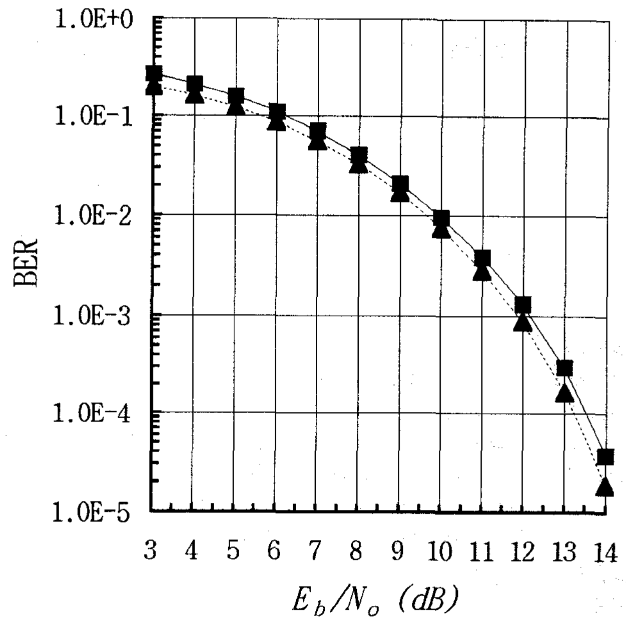

Fig. 8. Bit error probability versus E_b / N_o for channel B (■ MLFE, ▲ DFE/coherent)