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Modified decision feedback differential detection for differentially encoded 16 APSK signals

Ruey-Yi Wei and Mao-Chao Lin

In 1996, a version of decision feedback differential detection (DFDD) was proposed for detecting differentially encoded 16-level amplitude/phase shift keying (16 DAPSK) signals. The authors propose two modified DFDD methods, each of which uses an amplitude-state indicator. Both DFDD methods provide an improved error performance compared with the original DFDD.

DFDD and 16 DAPSK: For differential phase shift keying (DPSK) signals, decision feedback differential detection (DFDD) is an efficient detection method which can narrow the performance gap between differential detection and coherent detection [1]. In 1996, a version of DFDD was proposed for the detection of 16 DAPSK signals [2]. However, the error performance cannot approach that of coherent detection by using an infinitely large number of observed symbols.

The 16 DAPSK signal [2] can be considered as a combination of 8 DPSK and 2 DASK. The received signal sample of 16 DAPSK for a time k is

$$z_k = \sqrt{2P}r_k e^{j\phi_k} e^{j\theta} + n_k \quad (1)$$

where P is the signal power, ϕ_k is the modulation phase, θ is an unknown and constant phase, n_k is a sample of zero-mean complex Gaussian noise, and $r_k \in \{R_l, R_h\}$. Let $a = R_h/R_l (> 1)$. The messages for time k are carried by the phase difference $\Delta\phi_k = \phi_k - \phi_{k-1}$ and the amplitude ratio $\Delta r_k = r_k/r_{k-1}$. The amplitude ratio Δr_k is 1 if the corresponding message bit is '0'. The amplitude ratio Δr_k is either a or $1/a$ if the corresponding message bit is '1'.

In [2], the decision rule for time k is given as follows. At the beginning, the signal reference $z_{r,k-1}$, based on the feedback of previously detected symbols, is calculated by

$$z_{r,k-1} = \frac{\sum_{l=1}^L z_{k-l} \prod_{i=1}^{l-1} \Delta r_{k-i}^{-1} e^{j\Delta\phi_{k-i}}}{\sum_{l=1}^L \prod_{i=1}^{l-1} \Delta r_{k-i}^{-2}} \quad (2)$$

Then, $\Delta\phi_k$ is determined as the $\Delta\Phi_k$ value that maximises:

$$\psi = \text{Re}\{z_k z_{r,k-1}^* \exp(-j\Delta\Phi_k)\} \quad (3)$$

Finally, Δr_k is chosen to be:

$$\Delta r_k = \begin{cases} a & \text{if } \Delta\hat{r}_k > \gamma_h \\ 1/a & \text{if } \Delta\hat{r}_k < \gamma_l \\ 1 & \text{if } \gamma_l \leq \Delta\hat{r}_k \leq \gamma_h \end{cases} \quad (4)$$

where γ_h and γ_l are thresholds, and

$$\Delta\hat{r}_k = \text{Re}\{z_k z_{r,k-1}^* \exp(-j\Delta\hat{\Phi}_k)\} / |z_{r,k-1}|^2 \quad (5)$$

The information bits can be recovered based on Δr_k and $\Delta\phi_k$. Simulation results of such a DFDD are presented in [2], which show

that $a = 2$ and $(\gamma_l, \gamma_h) = (0.68, 1.5)$ are the parameters that can result in the best error performance. However, even as L approaches infinity, the error performance of the DFDD in [2] is still worse than that of coherent detection by 1dB at a high signal-to-noise ratio.

Modified DFDD: Note that $\Delta r_k \in \{a, 1\}$ if $r_{k-1} = R_l$ and $\Delta r_k \in \{1/a, 1\}$ if $r_{k-1} = R_h$. Hence, the poor error performance of the DFDD in [2] is mainly due to the inappropriate decision rule as given in eqn. 4, which ignores the information of r_{k-1} and determines Δr_k to be one of $\{1, a, 1/a\}$. For example, if r_{k-1} is R_h , then it is apparently inappropriate to consider $1/a$ as a possible value of Δr_k . To solve this problem, we propose an amplitude-state indicator assigned with value of $\alpha_k \in \{0, 1\}$ for time k . When $\alpha_k = 0$, Δr_k is chosen to be:

$$\Delta r_k = \begin{cases} a & \text{if } \Delta\hat{r}_k > \gamma_h \\ 1 & \text{if } \Delta\hat{r}_k \leq \gamma_h \end{cases} \quad (6)$$

When $\alpha_k = 1$, Δr_k is chosen to be:

$$\Delta r_k = \begin{cases} 1/a & \text{if } \Delta\hat{r}_k > \gamma_l \\ 1 & \text{if } \Delta\hat{r}_k \leq \gamma_l \end{cases} \quad (7)$$

If $\Delta r_k = 1$, then $\alpha_{k+1} = \alpha_k$; if $\Delta r_k \neq 1$, then $\alpha_{k+1} \neq \alpha_k$. It is expected that α_k will be a good estimation of r_{k-1} in the modified DFDD. The phenomenon of error propagation by using the amplitude-state indicator is not serious. If the indicator is not correct, it is likely to turn to the correct value after only one more incorrect value.

Fig. 1 Error performance of modified DFDD using rules of eqns. 2, 3, 5, 6 and 7

- *— $L = 1$
- △— $L = 2$
- ×— $L = 3$
- $L = 5$
- +— $L = 10$
- ◇— coherent detection

Simulation shows that the error performance of the modified DFDD, using the rules of eqns. 2, 3, 5 and 7 with $a = 2$ and $(\gamma_l, \gamma_h) = (0.68, 1.5)$, is still not satisfactory. The problem is due to the incorrect selection of (γ_l, γ_h) for $a = 2$. In [2], it is said that if $z_{r,k-1}$ is a good estimation and $\Delta\phi_k$ is correctly decided, we should choose Δr_k as the ΔR_k value which can minimise:

$$\eta = |z_k - z_{r,k-1} e^{j\Delta\phi_k} \Delta R_k|^2 \quad (8)$$

where $\Delta R_k \in \{1/a, 1, a\}$. Apparently, a more appropriate selection of ΔR_k must be based on the amplitude-state indicator. That means, $\Delta R_k \in \{1/a, 1\}$ if $\alpha_k = 1$ and $\Delta R_k \in \{1, a\}$ if $\alpha_k = 0$. Minimising η is equivalent to minimising $\zeta = |\Delta R_k - \Delta\hat{r}_k|$ [2]. For $r_{k-1} = R_l$ ($\alpha_k = 0$), the optimal γ_h to minimise ζ is $\frac{1}{2}(1 + a)$. For $r_{k-1} = R_h$ ($\alpha_k = 1$), the optimal γ_l is $\frac{1}{2}(1 + 1/a)$. Hence, for $a = 2$, the optimum threshold should be $(\gamma_l, \gamma_h) = (0.75, 1.5)$.

In [2], the optimum threshold for $a = 2$ is around (0.68, 1.5), which is due to the inappropriate decision rule of eqn. 4.

Fig. 1 shows the simulation results of the modified DFDD using rules of eqns. 2, 3, 5, 6 and 7, with $a = 2$ and $(\gamma_b, \gamma_h) = (0.75, 1.5)$. The error performance of the modified DFDD with $L = 2$ is similar to that of the DFDD given in [2] with $L = 7$, and the error performance of the modified DFDD with $L = 3$ is better than that of the DFDD given in [2] with L approaching infinity. Moreover, the error performance of the modified DFDD with $L = 10$ is very close to that of coherent detection.

Fig. 2 Error performance of modified DFDD using rules of eqns. 3, 5, 6, 7, and 9

—*— $W = 0.5$
—△— $W = 1$
—×— $W = 2$
—□— $W = 6$
—◇— coherent detection

Motivated by the recursively generated signal references of the DFDD for DPSK signals given in [3, 4], the signal reference for 16 DPSK given in eqn. 2 can be calculated by a recursive form:

$$z_{r,k-1} = \frac{z_{k-1} + W z_{r,k-2} \Delta \bar{r}_{k-1}^{-1} e^{j\Delta \bar{\phi}_{k-1}}}{1 + W \Delta \bar{r}_{k-1}^{-2}} \quad (9)$$

The derivation of eqn. 9 is omitted here. Simulation results of the error performance for the modified DFDD using the rules of eqns. 3, 5, 6, 7 and 9 for $a = 2$ and various values of W are given in Fig. 2. When W becomes larger, its error performance approaches that of coherent detection. Note that although the DFDD with a larger value of W has a better error performance without increasing the complexity, the value of W should not be too large for a time-varying channel.

Finally, we would like to emphasise that the error performance of the modified DFDD can be further improved by using $a = 1.8$ instead of $a = 2$. In [2], the optimum value of a is ~ 2 , which is due to the inappropriate decision rule of eqn. 4.

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Multitone jamming rejection of FFH/BFSK linear-combining receiver over Rayleigh-fading channels

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A computationally efficient BER expression for an FFH/BFSK linear-combining receiver against multitone jamming and AWGN over Rayleigh-fading channels is derived based on a Taylor series expansion. The analytical expression, validated by simulation results, allows us to efficiently analyse the system performance with higher diversity levels, which is otherwise mathematically intractable.

Introduction: In a fast frequency-hopping binary frequency-shift-keying (FFH/BFSK) spread-spectrum (SS) communication system, the data signal with bit energy E_b and bit duration T_b is pseudo-randomly hopped over L hopping frequencies within the total SS bandwidth W_{SS} . The parameter $L \geq 1$ is called the diversity level, and the hopping duration is $T_h = T_b/L$. The bit error rate (BER) analysis of an FFH/BFSK receiver under the conditions of multitone jamming (MTJ) and additive white Gaussian noise (AWGN) over the fading channel has been studied in [1] for $L = 1$. However, the result in [1] cannot be generalised for the $L > 1$ case. In this Letter, we focus on the BER performance of an FFH/BFSK SS system with diversity combining against a single tone per band MTJ (i.e. there is only, at most, a jamming tone at one of the two possible orthogonal signalling frequencies in a specific FH band) and AWGN in a fading environment. We make use of the first three terms of a Taylor expansion to obtain the characteristic function of the system and derive a computationally efficient BER expression for any diversity level.

Fig. 1 Block diagram of FFH/BFSK linear-combining receiver

System model: For the FFH/BFSK linear-combining receiver shown in Fig. 1, the received signal is down-converted, bandpass filtered and dehopped to produce:

$$r(t) = \sqrt{2}a_s \cos(2\pi f_n t + \phi_s) + q_{nl} J(t) + w(t) \quad n = 1 \text{ or } 2 \quad (1)$$

where $\sqrt{2}a_s$ is the amplitude of the desired signal, $q_{nl} = 0$ or 1 is an indicator function representing the jamming state of the l th hop, $J(t) = \sqrt{2}a_j \cos(2\pi f_j t + \phi_j)$ is the multitone jammer, and $w(t)$ is the Gaussian noise process with zero mean and variance $\sigma_w^2 = N_0 B$, where N_0 is the one-sided power spectral density of AWGN and B is the bandwidth of the bandpass filters. We assume that there are N non-overlapping FH bands, each of bandwidth $2B$ Hz. The MTJ is assumed to have total power of P_{JT} which is distributed over Q equal-power interfering tones. The power of each jamming tone is $P_J = P_{JT}/Q$. An equivalent MTJ power spectral density is defined as $N_J \triangleq P_{JT}/W_{SS}$. The desired signal amplitude is modelled