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We investigate two erasure insertion schemes for a coded slow frequency-hopping system with multiple antennas at both the transmitter and the receiver. The demodulated symbols may become unreliable due to fading, thermal noise, and partial band interference (PBI). Since error correcting codes can correct more erasures than errors, it is often beneficial to erase these unreliable symbols rather than feeding them to the decoder. In this paper, we first extend the Bayesian erasure insertion, originally proposed in [10] for FSK modulation, to the case when differential space-time modulation is used. We then propose a simple, suboptimal erasure insertion scheme based on a likelihood ratio threshold test (LRTT). Simulation results show that substantial improvements in decoding error probability can be achieved by the proposed schemes over the non-erasure-insertion scheme, especially when the partial-band interference becomes severe.

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Erasure Insertion for Coded MIMO Slow Frequency-Hopping Systems in Presence of PBI

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Abstract— **We investigate two erasure insertion schemes for a coded slow frequency-hopping system with multiple antennas at both the transmitter and the receiver. The demodulated symbols may become unreliable due to fading, thermal noise, and partial band interference (PBI). Since error correcting codes can correct more erasures than errors, it is often beneficial to erase these unreliable symbols rather than feeding them to the decoder. In this paper, we first extend the Bayesian erasure insertion, originally proposed in [10] for FSK modulation, to the case when differential space-time modulation is used. We then propose a simple, suboptimal erasure insertion scheme based on a likelihood ratio threshold test (LRTT). Simulation results show that substantial improvements in decoding error probability can be achieved by the proposed schemes over the non-erasure-insertion scheme, especially when the partial-band interference becomes severe.**

I. Introduction

Ad hoc networks are an active area of research because of their capability to self-organize without a fixed infrastructure. However, due to the lack of central control, some commonly used multiple-access schemes, such as traditional TDMA or FDMA, may be difficult to implement in ad hoc networks. In addition, DS-CDMA may suffer from the lack of accurate power control. Consequently, frequency-hopping (FH) multiple-access has been widely used in packet radio networks [1] [2], especially for military applications, due to some of its desirable properties such as the anti-jamming capability, low probability of intercept, operability in non-continuous spectrum, and reduced sensitivity to near-far problems.

Multiple antennas have been shown to be an effective approach to improve link reliability and spectral efficiency [3]. The optimum performance attainable with multiple antennas requires perfect channel state information (CSI). However, the desired CSI is not easily achieved in FH systems because channels at different carrier frequencies may be low-correlated or independent, and the channel condition within a dwell can be time-varying. These considerations, together with the fact that there are multiple channels in a MIMO system, pose serious difficulties for implementing the optimal MIMO receiver. Previous studies on FH systems focus on single transmit and receive antenna and frequency-shift-keying (FSK) modulation with noncoherent detection. However, the performance of these systems are limited by the low spectral-efficiency of FSK modulation. In this paper, we study a FH system with multiple transmit and receive antennas. using differential space-time modulation (DSTM) instead of FSK. Given the knowledge of the hopping pattern, CSI is not required for demodulation at the receiver and the spectral-efficiency can be much higher than FSK modulation, as explained in the next section. Although we evaluate the specific DSTM scheme proposed in [4] in this paper, the idea and results can be extended to other more general DSTM schemes (e.g. [5] [6]).

In a coded communication system, the knowledge of the reliability of a symbol estimate is sometimes called "side information" [7]. Such information is beneficial because an error-correction code with minimum distance d can be decoded correctly if $2n_r + n_e < d$ [8,pp.452]. In other words, the number of erasures that can be tolerated is twice as many as the number of errors for a simple bounded distance decoder. Therefore, if a symbol estimate is subject to a high error probability, it may be preferable to output an erasure instead. This can also be viewed as a hard-decision based joint demodulation-decoding, which may have lower complexity and reduced latency than soft-decision based schemes.

The side information in a FH system can be obtained by using test symbols or concatenated coding [7], both requiring additional redundancy embedded in symbols from the same dwell. Two non-redundant erasure insertion schemes were proposed in [9] and [10]. In [10], an erasure insertion criterion based on the Bayesian decision theory is proposed to minimize a bound on the decoding error probability.

In this paper, we extend the Bayesian erasure insertion rule in [10] to a FH systems with DSTM in presence of PBI. We further propose an erasure insertion rule based on a likelihood ratio threshold test (LRTT), which is analogous to the Viterbi test [9] for FSK modulation. The LRTT receiver, when the DSTM from [4] is used, can be implemented with much lower complexity than the Bayesian receiver. Simulation results show that the receiver with erasure insertion can achieve significantly lower decoding error probability than the one without erasure insertion and the improvement in performance becomes more pronounced when the PBI becomes more severe.

The organization of the paper is as follows. System model is presented in Section II. The Bayesian and the LRTT erasure insertion schemes are discussed in Section III. Simulation

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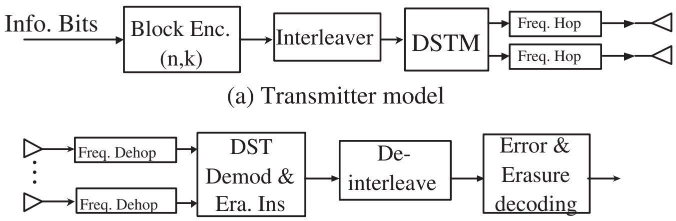

Fig. 1. System model

results are presented in Section IV and Section V concludes the paper.

II. System Model

The models for the transmitter and receiver are depicted in Fig.1. A block channel encoder with rate $\frac{k}{n}$ is used and the coded symbols are n-ary where $n = 2^{2b}$. We denote the set of coded symbols to be $\mathcal{A} \triangleq \{a_0, ..., a_{n-1}\}$. The DSTM maps each coded symbol to a space-time signal. The transmission can be illustrated in a time-frequency plane as in Fig.2, where each time-frequency slot has a time duration of T_s seconds and occupies a bandwidth of Δf. Symbols in the same dwell are transmitted on the same subcarrier. The subcarrier frequency is hopped to another value after each dwell. The next dwell is transmitted using the new subcarrier and so on. The subcarrier frequency sequence, or the hopping pattern, is assumed to be pseudo-random. The coded symbols are interleaved in such a way that symbols from the same codeword are transmitted in different dwells [7].

We use the DSTM proposed by [4] in this paper. Specifically, we consider two transmit antennas and $T_s = 2T$. Without loss of generality, we consider the dwell $[0, N_{dw}T_s]$, where N_{dw} is the number of slots within each dwell. In $[tT_s, (t+1)T_s]$ $(t = 0, \ldots, N_{dw} - 1)$, the transmitted signal can be represented as a 2×2 matrix in the form of

$$\begin{bmatrix} x_{2t,1} & x_{2t+1,1} \\ x_{2t,2} & x_{2t+1,2} \end{bmatrix} = \begin{bmatrix} x_{2t,1} & -x^*_{2t,2} \\ x_{2t,2} & x^*_{2t,1} \end{bmatrix}$$

where $x_{\tau,i}$ is the signal transmitted on the ith antenna in $[\tau T, (\tau+1)T]$ $(i = 1, 2,\ \tau = 2t, 2t+1)$. The redundant transmission of $x_{2t+1,1}$ and $x_{2t+1,2}$ provides diversity as defined in [4]. Here we implicitly assume $\Delta f \cdot T \geq 1$ so that $x_{2t,i}$ and $x_{2t+1,i}$ can be separated. Although this leads to $\Delta f \geq \frac{2}{T_s}$ (as opposed to the minimum subcarrier spacing $\Delta f \geq \frac{1}{T_s}$), it still offers better bandwidth efficiency than n-ary FSK modulation for $n > 2$, which requires $\frac{\Delta f}{n} \geq \frac{1}{T_s}$ or $\Delta f \geq \frac{n}{T_s}$. The signal $x_{2t,1}$ and $x_{2t,2}$ are generated as described below. Suppose an n-ary coded symbol $c_t \in \mathcal{A}$ is transmitted in $[tT_s, (t+1)T_s]$. It is first mapped to a pair $(s_{2t,1}, s_{2t,2})$, each element of which belongs to a 2^b-PSK constellation, i.e., $(s_{2t,1}, s_{2t,2}) = \mathcal{M}(c_t)$. The value of $x_{2t,1}$ and $x_{2t,2}$ depend on both c_t and also $(x_{2t-2,1}, x_{2t-2,2})$, the signal transmitted in the previous slot. That is

$$\begin{bmatrix} x_{2t,1} \\ x_{2t,2} \end{bmatrix} = \frac{1}{2} \begin{bmatrix} x_{2t-2,1} & -x^*_{2t-2,2} \\ x_{2t-2,2} & x^*_{2t-2,1} \end{bmatrix} \begin{bmatrix} 1 & 1 \\ -1 & 1 \end{bmatrix} \begin{bmatrix} s_{2t,1} \\ s_{2t,2} \end{bmatrix} \quad (1)$$

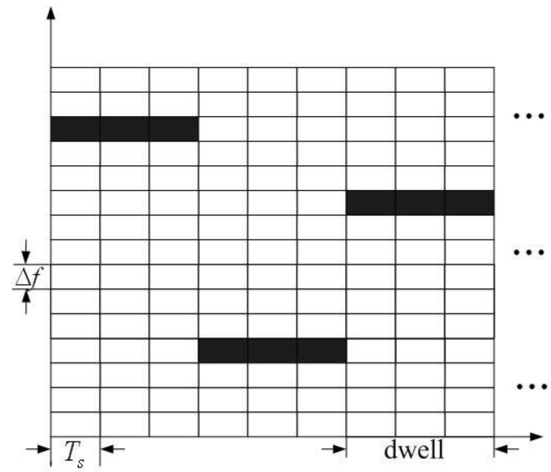

Fig. 2. An illustration of one user's transmission in time-frequency plane

In order to ensure that the received signal can be demodulated differentially, $x_{0,1}$ and $x_{0,2}$ are deterministic and both set to $1/\sqrt{2}$. As a consequence, at arbitrary t, the total transmit power $|x_{2t,1}|^2 + |x_{2t,2}|^2$ is always 1. We further assume c_t takes values in $\mathcal{A}$ with equal probability, which is valid when the coded symbols are interleaved and detected one-by-one.

The receiver is assumed to have N antennas. In $[tT_s, (t+1)T_s]$, the receiver processes two data samples on the jth antenna, which are given by

$$r_{2t,j} = h_{1,j}(t)\, x_{2t,1} + h_{2,j}(t)\, x_{2t,2} + w_{2t,j} \quad (2)$$
$$r_{2t+1,j} = h_{1,j}(t)\, x_{2t+1,1} + h_{2,j}(t)\, x_{2t+1,2} + w_{2t+1,j} \quad (3)$$

We assume the channel is constant during $[tT_s, (t+1)T_s]$ and $h_{i,j}(t)$ is the fading coefficient from transmit antenna i to receive antenna j. The correlation function of $\{h_{i,j}(t)\}$ is $J_0(2\pi f_d T_s \tau)$ $(\tau = 0, 1, \ldots)$, as in Jakes' model. We further assume $\{h_{i,j}(t)\}$ are independent random processes for different (i, j) pair. The coefficient $h_{i,j}(t)$ has the circularly symmetric complex Gaussian distribution $\mathcal{CN}(0, 1)$. We normalize the average channel gain $E|h_{i,j}(t)|^2 = 1$. Therefore, the average received power is also 1.

The additive term $w_{\tau,j}$ consists of thermal noise and possibly partial-band interference (PBI). The PBI is only present in a fraction ρ of the total bandwidth. When it is present, the PBI is modelled as $\mathcal{CN}\left(0, \frac{1}{\rho}\sigma_I^2\right)$ random variable. This implies that the average power of PBI is proportional to σ_I^2 and independent of ρ. The PBI terms at different receive antennas are assumed to be uncorrelated, if that dwell is hit by PBI. In general, the PBI-bands could be time-varying but we assume such variation is much slower than the hopping rate. Therefore we can assume that the PBI is either present or absent in a complete dwell. The probability of a dwell contaminated by the PBI equals ρ when a pseudo-random hopping pattern is used. The thermal noise exists over the entire available frequency band and is modelled as $\mathcal{CN}\left(0, \sigma_N^2\right)$ random variables, which are uncorrelated in both time and space, and independent of the PBI process.

III. Erasure Insertion

A. ML Demodulation

At the receiver, data samples received during the current and the previous time slot are used for differential demodulation. That is, the receiver forms the matrix

$$\mathbf{Y}_t = [\mathbf{y}_{t,1}, \ldots, \mathbf{y}_{t,N}] \tag{4}$$

where the nth column is

$$\mathbf{y}_{t,n} = [r_{2t-2,n}, r_{2t-1,n}, r_{2t,n}, r_{2t+1,n}]^T \tag{5}$$

contains the data received on the nth receive antenna during $[(t-1)T_s, (t+1)T_s]$. The ML demodulator chooses $l_{ML} = \arg\max_l f\left(\mathbf{Y}_t|a_l\right)$. We show that, in the presence of PBI, the likelihood function is given by

$$\begin{aligned} f\left(\mathbf{Y}_t|a_l\right) \quad = \quad & \frac{1-\rho}{\pi^{4N}\xi_N^{2N}} \exp\left(-\operatorname{tr} \mathbf{Y}_t^H \mathbf{C}_{y|l,N}^{-1} \mathbf{Y}_t\right) \\ & + \frac{\rho}{\pi^{4N}\xi_P^{2N}} \exp\left(-\operatorname{tr} \mathbf{Y}_t^H \mathbf{C}_{y|l,P}^{-1} \mathbf{Y}_t\right) \end{aligned} \tag{6}$$

where

$$\xi_N = \left(1+\sigma_N^2\right)^2 - R_h^2(1) \tag{7}$$

$$\mathbf{C}_{y|l,N} = \begin{bmatrix} \mathbf{I}_2 & R_h^*(1)\,\boldsymbol{\Lambda}(l) \\ R_h(1)\,\boldsymbol{\Lambda}^H(l) & \mathbf{I}_2 \end{bmatrix} + \sigma_N^2 \mathbf{I}_4 \tag{8}$$

$$\boldsymbol{\Lambda}(l) = \begin{bmatrix} s_{l,2}^* + s_{l,1}^* & s_{l,1} - s_{l,2} \\ s_{l,2}^* - s_{l,1}^* & s_{l,2} + s_{l,1} \end{bmatrix} \tag{9}$$

where $\mathbf{I}_M$ is the $M \times M$ identity matrix, $R_h(1) = J_0(2\pi f_d T_s)$ is the correlation between channel coefficients in two consecutive time slots. ξ_P and $\mathbf{C}_{y|l,P}$ can be obtained similarly by replacing σ_N^2 by $\sigma_P^2 \triangleq \sigma_N^2 + \frac{1}{\rho}\sigma_I^2$.

B. Bayesian erasure insertion

Although the $a_{l_{ML}}$ is the optimal symbol estimate, as argued in Section I, sometimes it may be desirable to erase $a_{l_{ML}}$. How and when this should be done depends on the specific code and the decoding scheme. Here we assume that an (n,k) Reed-Solomon code with the simplest bounded-distance decoding is used, which can decode correctly if and only if $2n_r + n_e \le n-k$. We use Reed-Solomon code and the simple decoding scheme so that our results may be compared to previous results [7] [10] more easily. Deriving and utilizing the side information to assist decoding for more powerful decoding scheme as proposed in [11] [12] and other codes is currently being investigated.

When an (n,k) Reed-Solomon code and bounded-distance decoding is used, the decoding error probability is given by

$$P_E = \sum_{i=0}^{n} \sum_{j=j_0(i)}^{n-i} \frac{n!\,p_r^i p_e^j}{i!\,j!\,(n-i-j)!} \left(1-p_r-p_e\right)^{n-i-j} \tag{10}$$

where p_r and p_e are the probability of a symbol in the codeword being in error and erased, respectively, and $j_0(i) = \max(n-k+1-2i, 0)$.

Baum and Pursely showed in [10] that the Chernoff bound on P_E in (10) is given by

$$P_E \le \min_{\beta>1} \beta^{k-n} \sum_{i=0}^{n} \binom{n}{i} \left(\beta^2-1\right)^i \left(p_r + \frac{1}{\beta+1}p_e\right)^i \tag{11}$$

They further proposed to minimize $p_r + \frac{1}{\beta+1}p_e$ for some fixed value of β ($\beta > 1$). By formulating the problem in a Bayesian decision framework, the authors showed that the optimal decision rule, in terms of minimizing $p_r + \theta p_e$ ($\theta < \frac{1}{2}$), is to insert an erasure if

$$\frac{f\left(\mathbf{Y}_t|a_{l_{ML}}\right)}{\sum_{k=0}^{n-1} f\left(\mathbf{Y}_t|a_k\right)} \le 1-\theta \tag{12}$$

and outputs $a_{l_{ML}}$ otherwise.

C. Likelihood ratio threshold test (LRTT)

The ratio threshold test [9] is a well-known technique for erasure insertion in FH systems with FSK modulation, where the demodulator inserts an erasure whenever the ratio of the second largest to the largest envelope detector's output exceeds a threshold. The LRTT technique proposed here is based on the same idea. Specifically, if we denote $a_{l'_{ML}}$ to be the symbol such that $f\left(\mathbf{Y}_t|a_{l'_{ML}}\right) > f\left(\mathbf{Y}_t|a_l\right)$ for all $l \ne l_{ML}$ or l'_{ML}. The demodulator outputs an erasure if

$$\frac{f\left(\mathbf{Y}_t|a_{l'_{ML}}\right)}{f\left(\mathbf{Y}_t|a_{l_{ML}}\right)} \ge \tau \tag{13}$$

for fixed $\tau \in (0,1)$. When (13) is not satisfied, the demodulator outputs $a_{l_{ML}}$. The LRTT rule is based on the rationale that when the noise plus interference is insignificant, it is expected that all $f\left(\mathbf{Y}_t|a_l\right)$ would be negligible except for $l = l_{tx}$ and the ratio in (13) is small, where l_{tx} is the transmitted symbol. On the other hand, when this ratio is significant, or when $f\left(\mathbf{Y}_t|a_{l'_{ML}}\right)$ is comparable to $f\left(\mathbf{Y}_t|a_{l_{ML}}\right)$, it may imply that the noise plus interference is strong and cast doubts on if $l_{ML} = l_{tx}$. Therefore, this ratio can be considered a measure of the unreliability of the ML estimate l_{ML}. When this measure exceeds the prescribed threshold τ, the demodulator inserts an erasure rather than to output l_{ML}.

The LRTT rule can also be viewed as an approximation to the Bayesian rule. If we neglect $f\left(\mathbf{Y}_t|a_l\right)$ ($l \ne l_{ML}, l'_{ML}$) in the sum of the denominator of (12), the two equations (12) and (13) become equivalent with $\tau = \theta/(1-\theta)$.

D. Implementation

One of the main advantages of the DSTM proposed in [4] is that it allows efficient implementation for ML demodulation in AWGN [13]. This properties can also be used here to simplify the implementation of the LRTT rule. We first notice that the exponents of both terms in (6) has the same function form and can be computed as follows. If we denote $g_{t,l}\left(\sigma_N^2\right) = \mathrm{tr}\,\mathbf{Y}_t^H \mathbf{C}_{y|l,N}^{-1}\mathbf{Y}_t$ and $g_{t,l}\left(\sigma_P^2\right) = \mathrm{tr}\,\mathbf{Y}_t^H \mathbf{C}_{y|l,P}^{-1}\mathbf{Y}_t$, (6) becomes

$$f\left(\mathbf{Y}_t|a_l\right) = \frac{1-\rho}{\pi^{4N}\xi_N^{2N}} e^{-g_{t,l}(\sigma_N^2)} + \frac{\rho}{\pi^{4N}\xi_P^{2N}} e^{-g_{t,l}(\sigma_P^2)} \quad (14)$$

After substituting (8) and $\mathbf{Y}_t$ into $g_{t,l}\left(\sigma_N^2\right)$ and simplifying, it can be shown that

$$g_{t,l}(x) = \frac{2(1+x^2)\left\|\mathbf{Y}_t\right\|_F^2 + R_h(1)(\varphi_{t,l}+\theta_t)}{2[(1+x^2)^2 - R_h^2(1)]} \quad (15)$$

where

$$\begin{aligned}
\overline{R}_{t,1} &\triangleq \sum\nolimits_{n=1}^{N}[r_{2t-2,n}^* r_{2t,n} + r_{2t+1,n}^* r_{2t-2,n} \\
&\quad - r_{2t,n} r_{2t-1,n}^* + r_{2t+1,n}^* r_{2t-1,n}] && (16)\\
\overline{R}_{t,2} &\triangleq \sum\nolimits_{n=1}^{N}[r_{2t-2,n}^* r_{2t,n} - r_{2t+1,n}^* r_{2t-2,n} \\
&\quad + r_{2t,n} r_{2t-1,n}^* + r_{2t+1,n}^* r_{2t-1,n}] && (17)\\
\varphi_{t,l} &\triangleq \left|\overline{R}_{t,1} - s_{l,1}\right|^2 + \left|\overline{R}_{t,2} - s_{l,2}\right|^2 && (18)\\
\theta_t &= -\left|\overline{R}_{t,1}\right|^2 + \left|\overline{R}_{t,2}\right|^2 - 2 && (19)
\end{aligned}$$

We see from (15) that $g_{t,l}(x)$ depends on l only through $\varphi_{t,l}$. Therefore, $\{g_{t,l}(\sigma_N^2)\}_{l=0}^{n-1}$ and $\{g_{t,l}(\sigma_P^2)\}_{l=0}^{n-1}$ have exactly the same ordering as $\{\varphi_{t,l}\}_{l=0}^{n-1}$. Consequently, the first two largest values of $\{f(\mathbf{Y}_t|a_l)\}_{l=0}^{n-1}$ corresponds to the first two minimal values of $\{\varphi_{t,l}\}_{l=0}^{n-1}$, which can be found efficiently. This avoids computing $f(\mathbf{Y}_t|a_l)$ for all $l = 0, \ldots, n-1$, and reduces the complexity significantly for large constellations.

IV. Numerical Results

In this section, we illustrate the gain of the Bayesian and LRTT erasure insertion relative to non-erasure insertion (ML demodulation) in terms of decoding error probability. We simulated p_r and p_e and then computed P_E by (10). This is more accurate than simulating P_E directly because the latter is much smaller than p_r or p_e. We use a $(16, 4)$ Reed-Solomon code and assume $f_d T_s = 0.05$ in the following results. The parameters $f_d T_s$, σ_N^2, σ_P^2, ρ are assumed to be known in the erasure insertion demodulation. The estimation of these parameters can be facilitated by the expectation-maximization (EM) algorithm, which will be presented in a subsequent paper and not discussed here. Fig.3 and Fig.4 show the P_E versus $1/\sigma_I^2$ curves at $1/\sigma_N^2$ equals 14dB and 22dB, respectively, while $\rho = 0.2$. In each figure, the curves for the Bayesian scheme and the LRTT scheme are plotted. The performance of ML demodulation with error-only bounded-distance decoding and the receiver of "perfect" side information are also drawn on the same figure for comparison. The receiver with perfect side information refers to a genie-aided receiver that erases

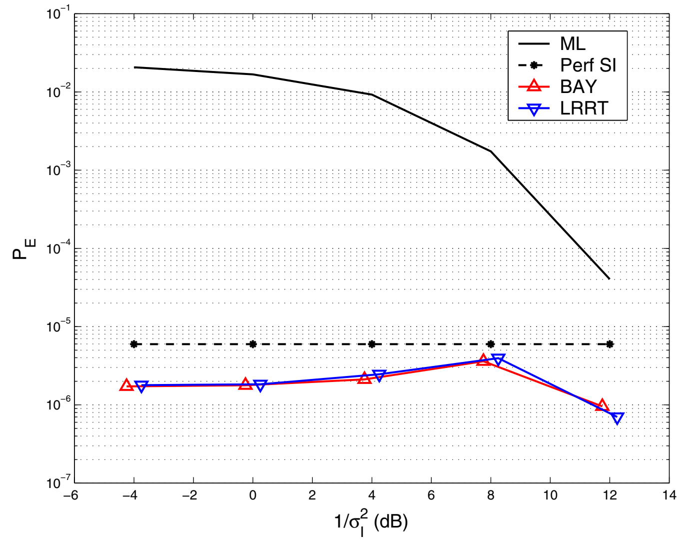

Fig. 3. P_E versus $\frac{1}{\sigma_I^2}$(dB) ($f_d T_s = 0.05$, (16,4) RS code, $\frac{1}{\sigma_N^2} = 14$dB, $\rho = 0.2$)

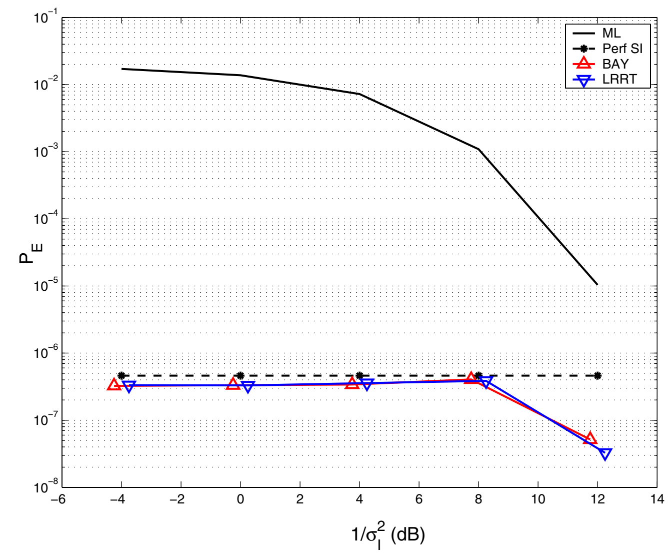

Fig. 4. P_E vs $\frac{1}{\sigma_I^2}$(dB) ($f_d T_s = 0.05$, (16,4) RS code, $\frac{1}{\sigma_N^2} = 22$dB, $\rho = 0.2$)

all received symbols contaminated by PBI and otherwise employs ML demodulation without erasure insertion. All erasure insertion schemes employ the optimal threshold by searching a number of candidate values.

We can see from Fig.3 and Fig.4 that the erasure insertion schemes provide significantly lower P_E. The genie-aided "perfect" side-information receiver erases symbols independent of the PBI's power and consequently achieves a decoding error probability invariant to the magnitude of $1/\sigma_I^2$. Both the Bayesian and LRTT receiver outperform the perfect side-information receiver because they have the flexibility of inserting erasure depending on the specific realization of the noise plus interference. The LRTT scheme has almost identical

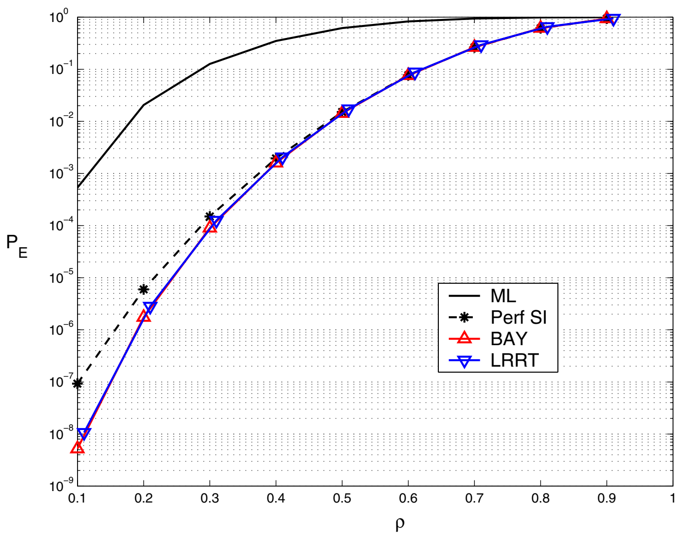

Fig. 5. P_E vs ρ ($f_dT_s = 0.05$, (16,4) RS code, $\frac{1}{\sigma_N^2}=14$dB, $\frac{1}{\sigma_I^2}=-4$dB)

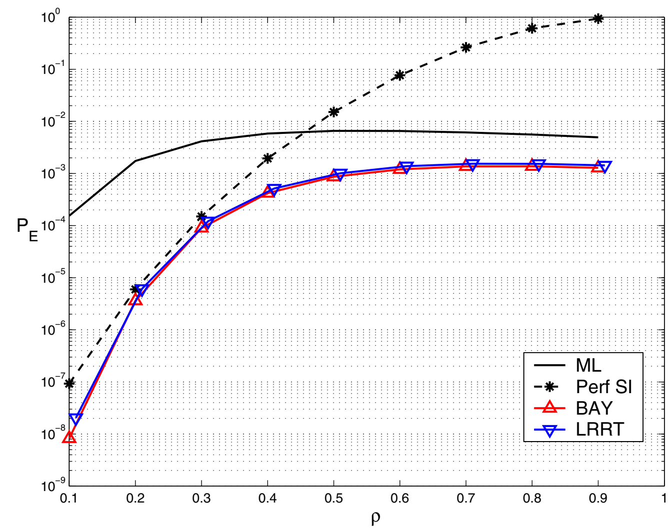

Fig. 6. P_E vs ρ ($f_dT_s = 0.05$, (16,4) RS code, $\frac{1}{\sigma_N^2}=14$dB, $\frac{1}{\sigma_I^2}=8$dB)

performance compared to the Bayesian scheme in both Fig.3 and Fig.4. It is interesting to notice from Fig.3 that P_E of both Bayesian and LRTT erasure insertion may not be monotonically decreasing as $1/\sigma_I^2$ increases. The same non-monotonicity is also observed in [10] for a non-orthogonal modulation other than the orthogonal FSK modulation.

Fig.5 and Fig.6 show the curves of P_E versus ρ at $1/\sigma_I^2$ equals -4dB and 8dB respectively, while $1/\sigma_N^2 = 14$dB. Again a large improvement is realized with erasure insertion schemes compared to the non-erasure insertion scheme especially when ρ is small. As ρ increases, the PBI becomes more like thermal noise and this difference decreases. When the PBI's average power is high ($1/\sigma_I^2$=-4dB), the erasure insertion schemes slightly outperform the receiver with perfect side information. When the PBI's average power is low, the receiver with perfect side information performs poorly at large ρ due to unnecessarily inserting erasures, while the two erasure insertion schemes still have good performance. Again, we observe close performance between the Bayesian and LRTT scheme. This makes the LRTT scheme very attractive due to its low complexity.

V. Conclusion

We studied two erasure insertion schemes which improve the decoding error probability significantly for a coded, differentially modulated MIMO slow frequency-hopping system in presence of partial-band interference/jamming. The first scheme is an extension of [10] to differential space-time modulation and minimizes a Chernoff-bound on the decoding error probability. The second scheme is based on the ratio of the first two largest likelihood function value and can be implemented at lower complexity compared to the Bayesian erasure insertion when the differential modulation in [4] is used. Both schemes achieve close performance in terms of decoding error probability, which is significantly lower than non-erasure insertion decoding, especially when the PBI is severe. The decoding performance of erasure insertion presented in this paper can be viewed as a lower bound because we assume optimal threshold and perfect knowledge of $R_h(1)$, σ_N^2, σ_P^2, ρ. Estimation of these parameters and the effect of estimation error in addition to mismatched threshold on the performance of erasure insertion are being investigated and will be discussed in future publications.

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