ML-Based Follower Jamming Rejection in Slow FH/MFSK Systems with an Antenna Array

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***Abstract*—The jamming robustness of frequency hopping (FH) systems with M-ary frequency shift keying (MFSK) modulation may be potentially neutralized by a follower partial-band jammer. In this paper, a maximum likelihood (ML)-based algorithm that uses a two-element array is proposed for joint follower jamming rejection and symbol detection in slow FH/MFSK systems over quasi-static flat Rayleigh fading channels. The algorithm is derived by treating both the received jamming components and the unknown data symbols as deterministic quantities to be jointly estimated in an integrated ML operation. In addition, an approximate expression for the symbol error rate (SER) of the proposed scheme is derived when BFSK signaling is employed in a jamming dominant scenario. Analytical and simulated results show that the proposed approach is able to remove jamming and outperform the conventional and sample matrix inversion (SMI)-based beam-formers in the presence of a follower partial-band jammer.**

***Index Terms*—Partial-band jammer, frequency hopping (FH), antenna array, quasi-static flat fading, maximum likelihood (ML), spatial-correlation, interference cancellation.**

I. INTRODUCTION

THE use of frequency-hopping spread-spectrum (FHSS) techniques for highly secure data transmission has been employed intensively in civilian and military wireless communications. However, under severely jammed propagation channels, the received jamming signal, whose power is comparable with or much greater than the signal power, will very likely induce an unacceptable degradation on FH detection performance [1]. For such applications, the use of anti-jamming is crucial to alleviate these detrimental effects so as to maintain a reliable communication channel in the presence of intentional interferers. Specifically, the performance of FHSS systems can be severely degraded in the presence of an intermittent jammer, such as a pulsed noise or a partial band jammer [1], that is present for only a fraction of the time. The detrimental effect caused by intermittent jamming may be compensated by appropriate channel coding. Unfortunately, even with channel coding, the performance of FHSS systems may still be significantly degraded in the presence of a follower partial-band jammer that has the capability to determine the frequency slot of the spread-spectrum bandwidth currently being used during some initial observation interval, and then injects the jamming signal in that frequency slot [2]. Fast hopping may be used to protect against such interference by prohibiting a follower jammer from having sufficient time to determine the desired signal's frequency slot and transmit an interfering signal. However, there is a penalty incurred in subdividing a signal into several FH elements. This is due to the fact that the energy from these separate elements has to be combined noncoherently. In addition, in FH systems, the transmitters and receivers contain clocks that must be synchronized. That is, the transmitters and receivers must hop at the same rate at the same time. The faster the hopping rate, the higher the jamming resistance, and the more accurate the clocks must be. This means that a highly accurate clock is required to allow a very fast hop rate for the purpose of defeating a follower jammer. It has been shown in [3] that under certain environments, the required accuracies can be achieved only with atomic clocks. As a result, some systems may still have limitations that do not allow for fast hopping [4].

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Investigations on FHSS systems in the presence of partial-band jamming have been carried out in [5]-[11], while studies on follower jamming mitigation have been documented in [4],[12],[13]. Specifically, in [4], a countermeasure to a partial-band Gaussian noise follower jammer was proposed for FHSS communications. The proposed scheme makes use of randomized decisions by the transmitter and the receiver to lure the jammer so that system performance can be improved. Of course, this implies that both the transmitter and receiver have to require a higher level of synchronization. In [12], the spatial dimension provided by an antenna array was exploited to achieve a better rejection of the follower jammer based on the classical sample matrix inversion (SMI) algorithm. However, this algorithm requires identical antenna gains for all receive antenna elements at the direction of arrival (DOA) of the jammer and does not work properly over flat fading channels. Similarly, while a variety of broadband source tracking algorithms [14]-[16] are available, they may not function properly under a flat fading scenario.

In this paper, we formulate a signal model that takes into consideration the effect of a follower jammer explicitly, and then propose a maximum likelihood (ML)-based joint interference cancellation and symbol detection scheme for slow FH/MFSK systems over quasi-static flat fading channels. The scheme is based on a two-element array where, at each element, N samples are extracted from the received signals within each transmitted symbol interval. By exploiting the unknown spatial correlation of the jamming components between the two antenna elements, a closed-form expression for the ML estimates of the jamming components is derived, leading to interference rejection and symbol detection being carried out in a unified ML framework.

Note that in present wireless communication systems such as GSM and Bluetooth based systems as well as other potential future ones using FH techniques, there is always the threat of Denial-of-Service (DoS) attack by intentional interferers [17], [18]. Specifically, the former is very vulnerable to jamming attack [17]. Under severely jamming scenarios where the jamming power is much greater than the signal power and the channel suffers from quasi-static flat fading, the proposed ML-based interference rejection structure and algorithm would provide a basis for the formulation of an appropriate solution to maintain a reliable communication channel.

The rest of this paper is organized as follows. Section II describes the system model. The derivation of the proposed interference rejection scheme is presented in Section III. The performance of the scheme is analyzed in section IV, where an approximate expression for SER is derived. Simulation results and discussion are given in Section V. Section VI concludes this paper.

II. System Model

Consider a MFSK modulated slow FH system. To suppress the detrimental effects of a follower partial band jammer, we explore the use of a simple two-element receiving array, where the received signal from each element is down converted and sampled at N times the symbol rate. The samples collected from the two antenna elements over one symbol duration will be used to estimate the desired information symbol by using a ML-based detection scheme, which will be described in more details in Section III.

Without loss of generality, consider the detection of the symbol in a hop over the interval $0 < t < T_s$, where T_s is the symbol duration. The complex envelop of the transmitted signal can be expressed by

$$s(t) = e^{(j2\pi(f_i+d_0f_d)t)}, \tag{1}$$

where f_i is the hopping frequency, $d_0 \in [0, 1, \ldots, M-1]$ represents the information symbol, and f_d stands for the frequency spacing between two adjacent MFSK tones. Note that, unlike conventional MFSK systems, the proposed scheme does not require the MFSK tones to be orthogonal.

As described in [19], a follower jammer first measures the hopping frequency and the spectrum of the desired hop and then injects the available transmitting power discriminately to the currently used frequency slot. Without perfect knowledge of the desired signal but knowing the hopping frequency of the desired signal, such a jammer will most likely transmit a signal that is different, perhaps noise like, from the desired signal and that will cover the entire band of the latter. The complex envelop of a follower partial-band jamming signal can thus be represented as

$$J(t) = n_j(t)\, e^{(j2\pi(f_i+B_j/2)t)}, \tag{2}$$

where $n_J(t)$ is a baseband equivalent band-limited signal with bandwidth B_J and can be modeled as a zero mean band-limited Gaussian random process. The exponential term in (2) indicates that this baseband signal is up converted to cover the frequency slot currently occupied by the desired signal in all the hops.

Assuming that the desired signal and the follower jamming signal experience a quasi-static flat Rayleigh fading channel, the received signal at the p-th antenna element will be given by

$$r_p(t) = \alpha_p s(t) + \beta_p J(t) + w_p(t),\ p = 1, 2, \tag{3}$$

where $w_p(t)$ is the complex white Gaussian receiver noise, and the complex coefficients α_p and β_p account for the overall effects of phase shifts, fading and antenna response for the desired signal and the jamming signal at the pth antenna element, respectively. Under a quasi-static flat fading channel, these fading coefficients can be assumed to be constant over one hop duration, equivalently a coherent interval.

Note that unlike the signal models in [7],[8],[12] which are derived for multiple partial-band and follower jamming signals coming from different directions, the signal model used in this paper is more applicable for a single follower jammer with known timing in a slow flat fading scenario.

At the pth antenna element, the received signal is sampled at N times the symbol rate. Using Equations (1), (2) and (3), the n-th sample is

$$r_{p,n} = \alpha_p \exp\left(f\omega_n(d_0)\right) + \beta_p J_n + w_{p,n}, \tag{4}$$

where

$$\begin{aligned} r_{p,n} &= r_p\left(\left(\frac{1}{2N}+\frac{n}{N}\right)T_s\right), \\ \omega_n(d_0) &= 2\pi\left(f_i+d_0f_d\right)\left(\frac{1}{2N}+\frac{n}{N}\right)T_s, \\ J_n &= J\left(\left(\frac{1}{2N}+\frac{n}{N}\right)T_s\right), \end{aligned} \tag{5}$$

and

$$w_{p,n} = w_p\left(\left(\frac{1}{2N}+\frac{n}{N}\right)T_s\right), \quad \text{for } n = 0, 1, \ldots, N-1.$$

Based on (4), the signal-to-jamming power ratio (SJR) and signal-to-noise power ratio (SNR) are SJR=P_S/P_J and SNR=P_S/P_N , respectively, with

$$P_S = E\left(\left|\alpha_p \exp\left[j\omega_n(d_0)\right]\right|^2\right) = E\left(\left|a_p\right|^2\right)$$

and

$$P_J = E\left(\left|\beta_p\right|^2\right) E\left(\left|J_n\right|^2\right)$$

and

$$P_N = E\left(\left|\omega_{p,n}\right|^2\right)$$

For convenience, Equation (4) can be written in vector form for the N samples from the two antenna elements as follows:

$$\mathbf{r}_1 = \alpha_a \mathbf{s}(d_0) + \mathbf{v} + \mathbf{w}_1, \tag{6}$$

and

$$\mathbf{r}_2 = \alpha_a \mathbf{s}(d_0) + \eta\mathbf{v} + \mathbf{w}_2 \tag{7}$$

where

$$\begin{aligned} \mathbf{r}_p &= \left[r_{p0}, r_{p,1}, r_{p,N-1}\right]^T, p = 1, 2 \\ \mathbf{s}(d_0) &= \begin{bmatrix} \exp\left(j\omega_0(d_0)\right), \\ \exp\left(j\omega_1(d_0)\right), \ldots, \exp\left(j\omega_{N-1}(d_0)\right) \end{bmatrix}^T \end{aligned} \tag{8}$$

$$\mathbf{v} = \beta_1 \left[J_0 J_1 J_{N-1} \right]^T, \eta = \beta_2/\beta_1,$$
$$\mathbf{w}_p = \left[w_{p,0} w_{p,1} w_{p,N-1} \right]^T \text{ with } p = 1, 2.$$

As the hopping frequency and spectrum of the desired signal need to be found, a follower jammer will not transmit any jamming signal during the initial measurement phase, and will be activated only after some delay following the beginning of each frequency hop [2], [12]. As a result, it would be reasonable to assume that the desired signal's channel gains, $\alpha_p (p = 1, 2)$, have been estimated and known to the receiver prior to the onset of the follower jamming signal. This is because ML-based channel estimation, described in Appendix B, can be easily performed blindly within a very short interval at the beginning of a hop. In the presence of the desired signal's channel knowledge, the main problem in jamming rejection and symbol detection is thus to estimate the data symbol d0 from received signal vectors $\mathbf{r}_p (p = 1, 2)$ in the presence of unknown jamming components η and $\mathbf{v}$ as well as independent receiver noise $\mathbf{w}_p (p = 1, 2)$.

As described in Appendix C, using the available channel estimates of the desired signal $\widehat{\alpha}_p, p = 1, 2$, a simple beamforming structure with weighting vector $\mathbf{g} = [\widehat{\alpha}_2 - \widehat{\alpha}_1]^T$ can be employed to place a null toward the desired signal. Deploying the technique in [12], the onset of the jamming signal can be detected by determining the time when a significant increase in the output signal power has occurred. Based on the detected jammed or unjammed status of the system, the appropriate algorithm can be employed for subsequent symbol detection. In particular, the unjammed symbols are detected by using the conventional ML technique, while the jammed symbols can be detected by the proposed approach which will be described in details in Section III.

III. ML-Based Interference Suppression and Detection Scheme

In the section, we formulate a ML-based interference rejection and detection scheme that can effectively suppress the jamming signal. Noting that the jamming components from the two antenna elements are spatially correlated through some unknown coefficients η, the vector of jamming components $\mathbf{v}$ and η will be treated as deterministic quantities to be estimated by the ML technique. This approach is different from the conventional one, where the jamming components are simply regarded as receiver noise.

Since MFSK modulation is employed, the desired symbol d_0 is given by only one of the alphabet $\{0, 1, \ldots, M-1\}$. A joint ML estimation of d_0, η and $\mathbf{v}$ can thus be expressed as

$$\widehat{\eta}, \widehat{\mathbf{v}}, \widehat{d}_0 = \arg\min_{\eta, \mathbf{v}, d} \left\{ \begin{array}{c} \|\mathbf{r}_1 - \alpha_1 \mathbf{s}(d) - \mathbf{v}\|^2 \\ + \|\mathbf{r}_2 - \alpha_2 \mathbf{s}(d) - \eta \mathbf{v}\|^2 \end{array} \right\}, \quad (9)$$

where $d \in \{0, 1, \ldots, M-1\}$ is the candidate symbol to be searched in the ML cost function.

For convenience, let us define

$$\mathbf{z}_p(d) = \mathbf{r}_p - \alpha_p \mathbf{s}(d), \text{ for } p = 1, 2, \quad (10)$$

so that the cost function in (9) becomes

$$\Gamma(d) = \|\mathbf{z}_1(d) - \mathbf{v}\|^2 + \|\mathbf{z}_2(d) - \eta \mathbf{v}\|^2. \quad (11)$$

Differentiating the cost function $\boldsymbol{\Gamma}(d)$ with respect to $\mathbf{v}$ and η, respectively, and setting the results to zero, we obtain

$$\mathbf{v} = \frac{\mathbf{z}_1(d) + \eta^* \mathbf{z}_2(d)}{1 + |\eta|^2}, \quad (12)$$

and

$$\eta = \frac{\mathbf{v}^H \mathbf{z}_2(d)}{\|\mathbf{v}\|^2}. \quad (13)$$

Substituting (12) into (11) yields

$$\Gamma(d) = \frac{\|\mathbf{z}_2(d) - \eta \mathbf{z}_1(d)\|^2}{1 + |\eta|^2}, \quad (14)$$

and by substituting (12) into (13), we get

$$a(d)\eta^2 + b(d)\eta - a^*(d) = 0, \quad (15)$$

where

$$a(d) = \mathbf{z}_2^H(d)\mathbf{z}_1(d), \quad (16)$$

and

$$b(d) = \|\mathbf{z}_1(d)\|^2 - \|\mathbf{z}_2(d)\|^2. \quad (17)$$

As a result, the closed-form expressions for the ML estimates of η which are the solutions to (15) can be determined by

$$\eta_1(d) = \frac{-b(d) - \sqrt{b^2(d) + 4|a(d)|^2}}{2a(d)}, \quad (18)$$

and

$$\eta_2(d) = \frac{-b(d) + \sqrt{b^2(d) + 4|a(d)|^2}}{2a(d)}. \quad (19)$$

In accordance with (9), (14), (18) and (19), an ML estimate of the transmitted symbol d_0 is therefore

$$\widehat{d}_0 = \arg\min_d \{\Gamma_1(d), \Gamma_2(d); d = 0, 1, \ldots, M-1\}, \quad (20)$$

where

$$\Gamma_2(d) = \frac{\|\mathbf{z}_2(d) - \eta \mathbf{z}_1(d)\|^2}{1 + |\eta_i(d)|^2}, \text{ for } i = 1, 2. \quad (21)$$

Equations (18) and (19) indicate that there are two possible estimates of η for a fixed value of d. Consequently, in accordance with (20), it seems that we have to calculate the two cost functions $\Gamma_1(d)$ and $\Gamma_2(d)$ corresponding to a fixed d for the purpose of estimating the desired symbol. Fortunately, as shown in the Appendix A, $\Gamma_2(d)$ is always smaller than $\Gamma_1(d)$ for a fixed value of d. Therefore, it is sufficient to just compute the cost function $\Gamma_2(d)$ corresponding to $\eta_2(d)$ in (19). As a result, the decision rule of (20) can be simplified to be given by

$$\widehat{d}_0 = \arg\min_d \{\Gamma_2(d); d = 0, 1, \ldots, M-1\}. \quad (22)$$

The detailed procedure for implementing the proposed ML-based interference rejection and detection algorithm can be summarized as follows:

1. initialize the candidate symbol $d = 0$;
2. calculate both $\mathbf{z}_1(d)$ and $\mathbf{z}_2(d)$ based on (5), (8), (10) as well as knowledge of α_1 and α_2;
3. compute both $a(d)$ and $b(d)$ using (16) and (17);
4. calculate $\eta_2(d)$ using (19);

TABLE I
COMPUTATIONAL COMPLEXITY OF THE PROPOSED ALGORITHM

Step	Number of real addition	Number of real multiplication
2	$8NM$	$8NM$
3	$8NM$-$3M$	$8NM$
5	$6NM$+M	$6NM$+$3M$

5. compute $\Gamma_2(d)$ based on (21);
6. if $d = M-1$, go to Step 7; otherwise $d = d+1$ and return to Step 2;
7. obtain the ML estimate of the transmitted symbol $\widehat{d_0}$ based on (22).

The computational burden of the proposed algorithm is mainly due to Steps 2, 3 and 5, since only these three steps involve vector operations. The numbers of real addition and real multiplication used in these steps are shown in Table I. It is easy to see that the computational complexity of the proposed algorithm is $O(20NM)$ in terms of the number of real additions and multiplications needed.

Note that the proposed algorithm and structure is based on the use of two receive antennas to remove unknown but spatially correlated jamming. With a single antenna, it will not be possible to remove the jamming, which is in the same frequency band as the signal. The use of more than two antennas will lead to better performance if there is only a single jammer. However, the cost may be significantly larger in terms of the space needed and the additional receiving electronics, especially in a mobile application where space and power supply is restricted.

IV. PERFORMANCE ANALYSIS

In the section, we will derive an approximate expression for the symbol error rate (SER) of the proposed ML-based jamming rejection and symbol detection scheme. For the sake of simplicity, we consider only BFSK signaling over a jamming dominant channel, noting that the case for M-ary signaling can be similarly analyzed.

Taking the two possible BFSK symbols to be equiprobable, using the decision rule of (22), and assuming, without loss of generality, that the transmitted symbol value is $d_0 = 0$, the SER can be easily shown to be

$$P_e = \Pr\{f(0) > f(1)\}, \tag{23}$$

where the two conditional cost functions $f(0)$ and $f(1)$ are given by

$$f(m) = \Gamma_2(d=m)_{d_0=0}\, m = 0, 1. \tag{24}$$

Similarly, the resulting input signal vectors now become

$$\mathbf{r}_1 = \alpha_1 \mathbf{s}(0) + \mathbf{v} + \mathbf{w}_1, \tag{25}$$

and

$$\mathbf{r}_2 = \alpha_2 \mathbf{s}(0) + \eta\mathbf{v} + \mathbf{w}_2. \tag{26}$$

Using (10), (21), (24), (25) and (26), the conditional cost function $f(0)$ can be determined by

$$f(0) = \frac{\left\|\eta\mathbf{v} + \mathbf{w}_2 - \eta_2^+(0)(\mathbf{v}+\mathbf{w}_1)\right\|^2}{1 + \left|\eta_2^+(0)\right|^2}, \tag{27}$$

where (see Equation (28) on next page). After some manipulation and simplification, we have

$$f(0) = \frac{\|\eta\mathbf{v} + \mathbf{w}_2\|^2 + \|\mathbf{v} + \mathbf{w}_1\|^2 - h_0}{2}, \tag{29}$$

where $h_0 = \sqrt{\begin{array}{c}\left(\|\eta\mathbf{v} + \mathbf{w}_2\|^2 - \|\mathbf{v}+\mathbf{w}_1\|^2\right)^2 \\ +4\left|[\eta\mathbf{v}+\mathbf{w}_2]^H[\mathbf{v}+\mathbf{w}_1]\right|^2\end{array}}$.

Under a severely jammed channel, where the power of the jamming signal is much greater than that of receiver noise $\mathbf{w}_p(p=1,2)$, the high order terms with respect to receiver noise $\mathbf{w}_p(p=1,2)$ can be omitted in a power series expansion of h_0. As a result, h_0 can be approximated by using just the zeroth and first order terms with respect to $\mathbf{w}_1$ and $\mathbf{w}_2$. The conditional cost function $f(0)$ can therefore be approximated by

$$f(0) = \frac{\|\eta\mathbf{v}+\mathbf{w}_2\|^2 + \|\mathbf{v}+\mathbf{w}_1\|^2 - \overbrace{\|\mathbf{v}\|^2\left(1+|\eta|^2\right)}^{\text{zeroth order term}}}{2} \overbrace{-\mathrm{Re}\left\{\mathbf{w}_2^H\mathbf{v}\eta\right\} - \mathrm{Re}\left\{\mathbf{w}_1^H\mathbf{v}\right\}}^{\text{first order term}} \tag{30}$$

Similarly, substituting (10), (21), (25) and (26) into (24) yields the conditional cost function as

$$f(1) = \frac{\|\mathbf{s}_2 + \eta\mathbf{v}_2 + \mathbf{w}_2\|^2 + \|\mathbf{s}_1 + \mathbf{v} + \mathbf{w}_1\|^2 - h_1}{2}, \tag{31}$$

where

$$h_1 = \left(\begin{array}{c}\left(\|\mathbf{s}_2+\eta\mathbf{v}+\mathbf{w}_2\|^2 - \|\mathbf{s}_1+\mathbf{v}+\mathbf{w}_1\|^2\right)^2 \\ +4\left|[\mathbf{s}_2+\eta\mathbf{v}+\mathbf{w}_2]^H[\mathbf{s}_1+\mathbf{v}+\mathbf{w}_1]\right|^2\end{array}\right)^{1/2}$$

and

$$\mathbf{s}_p = \alpha_p[\mathbf{s}(0) - s(1)] \text{ with } p = 1,2.$$

Using a power series expansion of h_1 and carrying out the same analysis as for h_0, it can be shown that $f(1)$ can be approximated by Equation (32) shown at top of next page where

$$\begin{aligned} q_0 &= \left(\|\mathbf{s}_2+\eta\mathbf{v}\|^2 - \|\mathbf{s}_1+\mathbf{v}\|^2\right)^2 \\ &\quad +4\left|[\mathbf{s}_2+\eta\mathbf{v}]^H[\mathbf{s}_1+\mathbf{v}]\right|^2 \\ \mathbf{q}_1 &= 4(\mathbf{s}_1+\mathbf{v})\left(\|\mathbf{s}_1+\mathbf{v}\|^2 - \|\mathbf{s}_2+\eta\mathbf{v}\|^2\right) \\ &\quad +8(\mathbf{s}_1+\eta\mathbf{v})[\mathbf{s}_2+\eta\mathbf{v}]^H(\mathbf{s}_1+\mathbf{v}), \end{aligned}$$

and

$$\begin{aligned} \mathbf{q}_2 &= 4(\mathbf{s}_1+\eta\mathbf{v})\left(\|\mathbf{s}_2+\eta\mathbf{v}\|^2 - \|\mathbf{s}_2+\eta\mathbf{v}\|^2\right) \\ &\quad +8(\mathbf{s}_1+\mathbf{v})[\mathbf{s}_1+\mathbf{v}]^H(\mathbf{s}_2+\eta\mathbf{v}). \end{aligned}$$

$$\begin{aligned} \eta_2^+(0) = \eta_2(d=0)\big|_{d_0=0} &= \frac{\|\eta\mathbf{v}+\mathbf{w}_2\|^2 - \|\mathbf{v}+\mathbf{w}_1\|^2 + W_{12}}{2\left[\eta\mathbf{v}+\mathbf{w}_2^H\right]\left[\mathbf{v}+\mathbf{w}_1\right]}, \\ W_{12} &= \sqrt{\left(\|\mathbf{v}+\mathbf{w}_1\|^2 - \|\eta\mathbf{v}+\mathbf{w}_2\|^2\right)^2 + 4\left|\left[\eta\mathbf{v}+\mathbf{w}_2\right]^H\left[\mathbf{v}+\mathbf{w}_1\right]\right|^2} \end{aligned} \tag{28}$$

$$\begin{aligned} f(1) &= \frac{\|\mathbf{s}_1+\mathbf{v}+\mathbf{w}_1\|^2 + \|\mathbf{s}_2+\eta\mathbf{v}+\mathbf{w}_2\|^2 - \overbrace{\sqrt{q_0}}^{\text{zeroth order term}}}{2}, \\ & \overbrace{-\,\mathrm{Re}\left\{\mathbf{w}_2^H\frac{\mathbf{q}_1}{4\sqrt{q_0}}\right\} - \mathrm{Re}\left\{\mathbf{w}_2^H\frac{\mathbf{q}_2}{4\sqrt{q_0}}\right\}}^{\text{first order term}} \end{aligned} \tag{32}$$

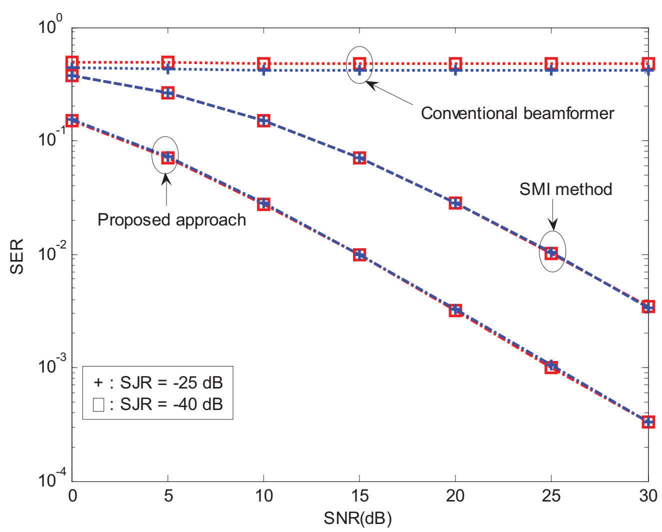

Fig. 1. Performance of the proposed approach under various SJRs with BFSK modulation and $N = 4$.

By substituting (30) and (32) into (23), the SER is thus determined approximately by

$$P_e = \Pr\{\Delta > 0\}, \tag{33}$$

where

$$\begin{aligned} \Delta = & -\|\mathbf{s}_2+\eta\mathbf{v}\|^2 - \|\mathbf{s}_1+\mathbf{v}\|^2 + \sqrt{q_0} \\ & +\mathrm{Re}\left\{\mathbf{w}_2^H\left(\frac{\mathbf{q}_2}{2\sqrt{q_0}} - 2\mathbf{v}\eta - 2\mathbf{s}_2\right)\right\} \\ & +\mathrm{Re}\left\{\mathbf{w}_1^H\left(\frac{\mathbf{q}_1}{2\sqrt{q_0}} - 2\mathbf{v} - 2\mathbf{s}_1\right)\right\}. \end{aligned}$$

Note that the quantity Δ includes the linear combination of the real and imaginary parts of the independent Gaussian receiver noise samples $w_{p,n}$. As a result, Δ is also Gaussian distributed and its mean μ_Δ and variance σ_Δ^2 can therefore be computed by

$$\mu_\Delta = -\|\mathbf{s}_2+\mathbf{v}\eta\|^2 - \|\mathbf{s}_1+\mathbf{v}\|^2 + \sqrt{q_0}, \tag{34}$$

and

$$\sigma_\Delta^2 = \sigma^2\left(\left\|\frac{\mathbf{q}_2}{2\sqrt{q_0}} - 2\mathbf{v}\eta - 2\mathbf{s}_2\right\|^2 + \left\|\frac{\mathbf{q}_1}{2\sqrt{q_0}} - 2\mathbf{v} - 2\mathbf{s}_1\right\|^2\right), \tag{35}$$

where σ^2 is the variance of the real and imaginary parts of the zero-mean white Gaussian receiver noise samples $w_{p,n}$.

In accordance with (33), (34) and (35), the SER can be computed approximately by

$$P_e = Q\left(-\frac{\mu_\Delta}{\sqrt{\sigma_\Delta^2}}\right), \tag{36}$$

where $Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^{+\infty}\exp\left(-\frac{t^2}{2}\right)dt$.

V. Simulations and Discussion

Numerical simulations have been conducted to validate the performance of the proposed interference suppression scheme for a slow FH system. In this system, each hop has 4 MFSK symbols, the symbol rate is 200000 symbols per second, and the hop rate is 50000 hops per second. The ratio of the unjammed interval to the hop duration, R_U, is given by 0.025 for all except the last result (Fig. 5).

Fig. 1 shows the SER of the proposed scheme versus the signal-to-noise ratio (SNR) when the signal- to-jamming ratio (SJR) is -25dB and -40dB. BFSK modulation is used and the number of samples per symbol is $N = 4$. For comparison, the results of using the conventional beamformer [20] and the SMI-based beamformer are also plotted. As can be seen, the performance of the proposed scheme differs only slightly for the various SJRs used, which is highly desirable in military communications. Also, unlike the conventional beamformer, no error floor exists for the proposed scheme. This is because the latter regards the jamming components as deterministic quantities to be estimated while the conventional beamformer simply treats the jamming components as receiver noise.

Furthermore, the proposed scheme is able to offer a better performance than the other methods since it is a ML-based approach. However, in the unlikely event that $\alpha_p = \beta_p$, as when both signal and jammer are from the same direction or there is no distinction between the signal and the jammer in terms of channel gains, all the algorithms will fail.

In fact, since there is no distinction between the signal and the jammer in terms oftransmission characteristics and the jamming signal is unknown, it will not be possible for any statistical signal processing algorithm to reject the jamming signal. Similarly, when two jammers are present and both are

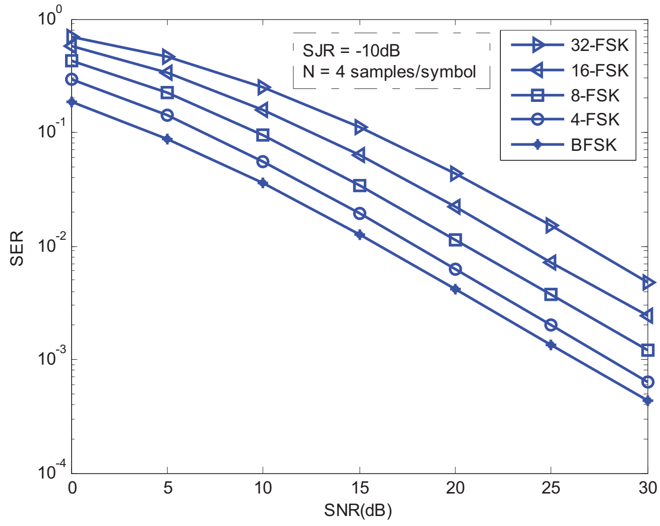

Fig. 2. Performance of the proposed scheme under various modulation levels and $N = 4$ samples/symbol.

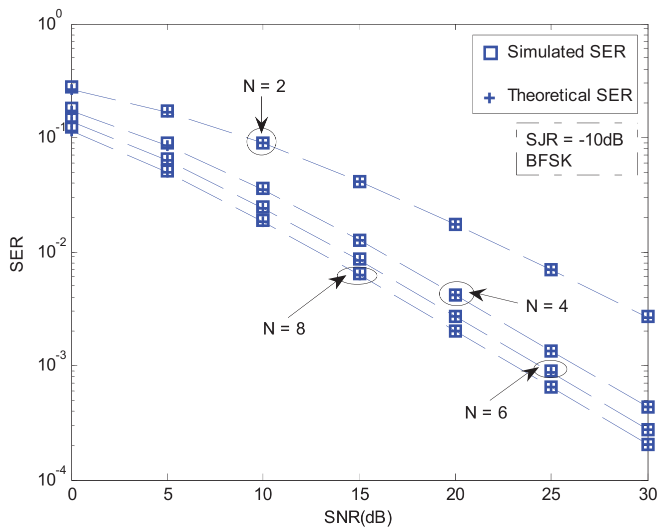

Fig. 3. Performance of the proposed scheme under various numbers of samples per symbol and the tightness of the theoretical and simulated SER values for BFSK signaling.

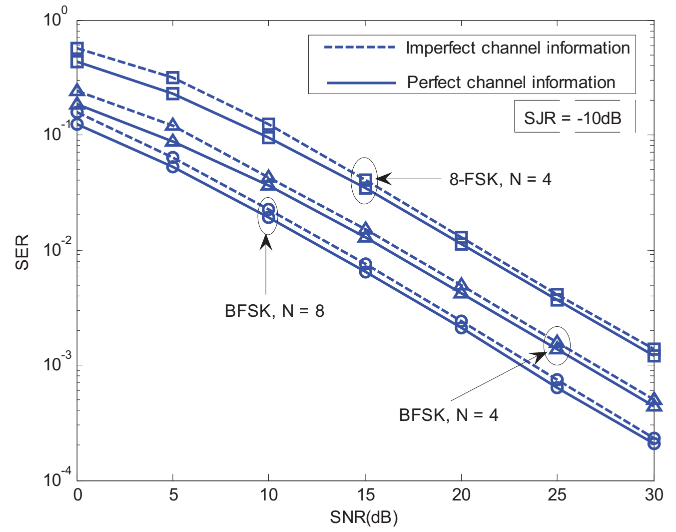

Fig. 4. Performance of the proposed scheme when the desired signal's channel gains are blindly estimated by using the ML technique in Appendix B within the unjammed interval of a hop.

unknown, it will not be possible for the proposed scheme, the SMI method and other similar techniques to work properly. This is because the array is a two-element one and the presence of two jammers will give rise to an under-determined system where the number of unknown parameters is more than number of the degrees of freedom that the system has.

Fig. 2 illustrates the performance of the proposed detection scheme under various modulation levels. The SJR is -10 dB and the number of samples per symbol is $N = 4$. As observed, the performance of the proposed scheme degrades as the modulation level increases.

Fig. 3 investigates the performance of the proposed scheme as the number of samples per symbol is changed. BFSK modulation is used and SJR is -10 dB. It can be seen that the proposed scheme has a better performance as the number of samples per symbol is increased. The validity of the performance analysis for the proposed scheme is also demonstrated in Fig. 3 from noting that the SER values from simulation are remarkably close to the corresponding analytical curves.

The results from Figs. 1, 2 and 3 have been obtained by assuming perfect channel estimation. To investigate the effect of imperfect channel estimation, Fig. 4 shows the performance of the proposed scheme with imperfect knowledge of the desired signal's channel gains, blindly estimated by using the ML technique (as described in Appendix B) within the unjammed interval of a hop. Obviously, at SJR=-10dB and using just 4 received samples in a very short unjammed interval of a hop to estimate the channel gains, the resulting SER performance in the case of imperfect channel estimation is very close to that in the case of perfect channel estimation.

Fig. 5 investigates the timing of the jamming signal on the system performance. The values of R_U used for the three sets of results are 0.025, 0.25 and 0.5, and the results are obtained as follows.

The dotted curves are obtained from using 10 samples of the received signals at the beginning of each hop in the ML approach (in Appendix B) to estimate the desired signal's channel response. Then, a simple beamforming structure is employed to place a null toward the desired signal. Using the technique in [12], the onset of jamming can then be detected by determining the time when a significant increase in the signal power at the beamformer's output has occurred.

Based on the detected jammed or unjammed status of the system, detection of the jammed symbols are carried out by the proposed approach, while that for the unjammed symbols are performed by using the conventional ML technique. The curves in Fig. 5 denote the overall SER results, including the SER performance in both the jammed and unjammed portions of each hop. As described, the dotted curves in Fig. 5 are obtained with imperfect channel estimates.

On the other hand, the solid curves are based on using the exact channel response of the desired signal. The minor performance degradation betweenthe two sets of curves again indicates that the new algorithm does not require very accurate channel information.

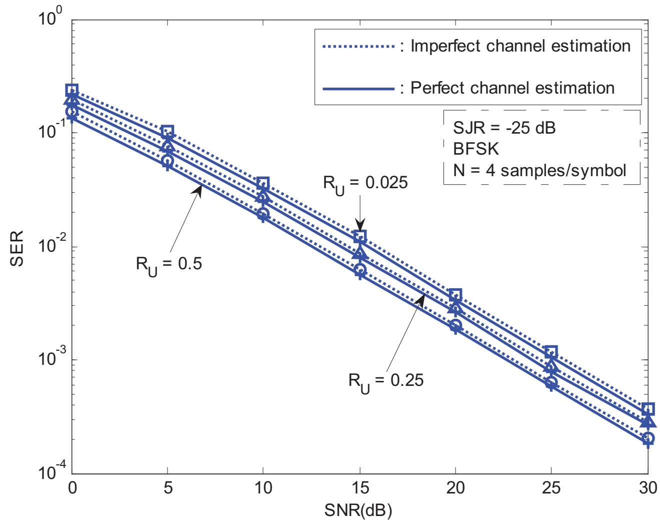

Fig. 5. Performance of the proposed scheme under various unjammed intervals in a hop.

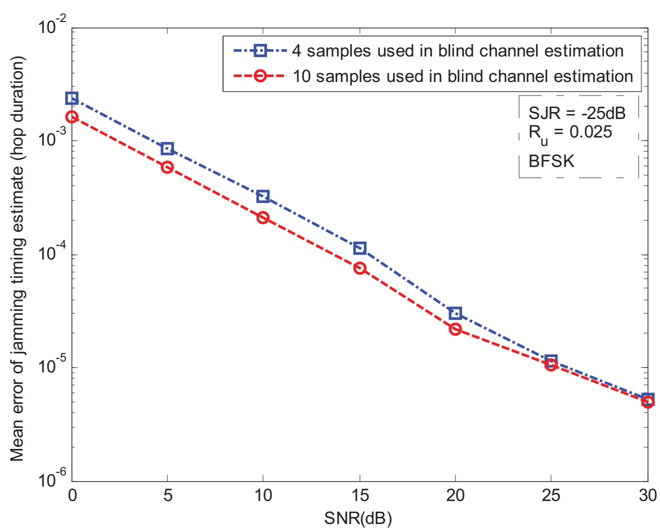

Fig. 6. Estimation of jamming timing.

The effect of the timing of the jamming signal can be studied in more detail by comparing the three sets of results in Fig. 5, each for a different value of R_U. Note that the lower the value of R_U, the more jammed the hop will be. As can be seen, while an increase in the jamming duration will worsen the SER performance, the use of the new algorithm has the effect that such deterioration becomes rather insignificant.

Finally, Fig. 6 examines the issue of jamming timing estimation. Specifically, the result is obtained from using the blind ML channel estimation algorithm given in Appendix B to estimate the channel gains of the desired signal, followed by implementing the beamformer in Appendix C to reject the desired signal based on these estimated gains, and then using the algorithm in [12] to detect the onset of jamming. The two curves in the figure show how the mean jamming timing estimate error, normalized with respect to the hop duration, changes as a function of SNR when 4 and 10 samples are used in the blind ML channel estimation procedure. As can be seen, using 10 received samples will give a more accurate timing estimation. However, this difference is rather insignificant, especially when the SNR is large. Also, even with a small number of samples, accurate timing estimate can be quite readily performed.

It should also be noted that other mitigation techniques, such as channel coding and interleaving, could also be used for the anti-jamming purpose. In fact, channel coding and interleaving is effective to intermittent jamming, such as a pulsed noise or a partial band jammer. However, even with channel coding and interleaving, the performance of FHSS systems will still deteriorate significantly in the presence of a follower jammer which is on most of the time. On the other hand, the proposed algorithm is able to suppress such a jammer. On the issue of complexity, the proposed algorithm operates only at the receiver and, as discussed in Section III, the implementation complexity is low. Comparatively, channel coding and interleaving techniques need to be used at both the transmitter and receiver, while interleaving will increase delay. Nevertheless, to further enhance performance, an appropriate channel coding and interleaving scheme may be used on top of the proposed algorithm.

VI. Conclusion

In this paper, a novel maximum likelihood (ML)-based joint interference cancellation and symbol detection scheme is proposed for slow FH/MFSK systems in the presence of a follower partial-band jammer over quasi-static flat Rayleigh fading channels. Based on the unknown spatial correlation of the jamming components from two antenna elements, a ML cost function is formulated to jointly perform symbol detection and interference rejection in an integrated ML operation. Based on a derived closed-form expression for the ML estimates of the received jamming components, the proposed scheme possesses a low computational complexity. It is robust against imperfect channel estimates, and has a much better SER performance than the conventional beamformer and the SMI method in the presence of a follower partial-band jammer.

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Appendix A
Proof of Inequality $\Gamma_2 d < \Gamma_1 (d)$

Substituting (16) and (17) into (18) and (19) yields

$$\eta_1 (d) = \frac{\|\mathbf{z}_2 (d)\|^2 - \|\mathbf{z}_1 (d)\|^2 - Z_{12}}{2\mathbf{z}_2^H (d)\, \mathbf{z}_1 (d)}, \tag{37}$$

and

$$\eta_2 (d) = \frac{\|\mathbf{z}_2 (d)\|^2 - \|\mathbf{z}_1 (d)\|^2 + Z_{12}}{2\mathbf{z}_2^H (d)\, \mathbf{z}_1 (d)}, \tag{38}$$

where $Z_{12} = \sqrt{\left(\|\mathbf{z}_2 (d)\|^2 - \|\mathbf{z}_1 (d)\|^2\right) + 4\left|\mathbf{z}_2^H (d)\, \mathbf{z}_1 (d)\right|^2}$.

Substituting (37) and (38) into the numerator of (21), respectively, we deduce

$$\Gamma_1 (d) \|\mathbf{z}_1 (d)\|^2 + \frac{\sqrt{\begin{array}{c}\left(\|\mathbf{z}_2 (d)\|^2 - \|\mathbf{z}_1 (d)\|^2\right)^2 \\ +4\left|\mathbf{z}_2^H (d)\, \mathbf{z}_1 (d)\right|^2\end{array}}}{1 + |\eta_2 (d)|^2}, \tag{39}$$

and

$$\Gamma_2(d)\,\|\mathbf{z}_1(d)\|^2 + \frac{\sqrt{\begin{array}{c}\left(\|\mathbf{z}_2(d)\|^2 - \|\mathbf{z}_1(d)\|^2\right)^2 \\ +4\left|\mathbf{z}_2^H(d)\,\mathbf{z}_1(d)\right|^2\end{array}}}{1+|\eta_2(d)|^2}. \quad (40)$$

As can be observed from (39) and (40), it is obvious that $\Gamma_2(d)$ is always smaller than $\Gamma_1(d)$.

Appendix B
Blind ML Estimation of the Desired Signal's Channel Gains

In the unjammed portion of the hop, a joint ML estimation of d_0, α_1 and α_2 can be obtained from

$$\widehat{\alpha}_1, \widehat{\alpha}_2, \widehat{d}_0 = \arg\min_{\alpha_1,\alpha_2,d_0} \left\{ \begin{array}{c} \|\mathbf{r}_1 - \alpha_1\mathbf{s}(d_0)\|^2 \\ +\|\mathbf{r}_2 - \alpha_2\mathbf{s}(d_0)\|^2 \end{array} \right\}. \quad (41)$$

Differentiating $\|\mathbf{r}_1 - \alpha_1\mathbf{s}(d_0)\|^2 + \|\mathbf{r}_2 - \alpha_2\mathbf{s}(d_0)\|^2$ with respect to α_1 and α_2, respectively, and setting the results to zero, we have

$$\alpha_p = \frac{\mathbf{s}^H(d_0)\,\mathbf{r}_p}{\|\mathbf{s}(d_0)\|^2}, p = 1, 2. \quad (42)$$

Substituting (42) into (41) then yield

$$\widehat{d}_0 = \arg\min_{d_0} \left\{ \begin{array}{c} \left\|\mathbf{r}_1 - \frac{\mathbf{s}^H(d_0)\mathbf{r}_1\ \mathbf{s}(d_0)}{\|\mathbf{s}(d_0)\|^2}\right\|^2 \\ +\left\|\mathbf{r}_2 - \frac{\mathbf{s}^H(d_0)\mathbf{r}_{12}\ \mathbf{s}(d_0)}{\|\mathbf{s}(d_0)\|^2}\right\|^2 \end{array} \right\}. \quad (43)$$

where $d_0 = 0, 1, \ldots, M-1$.

Based on the estimate of transmitted symbol $\widehat{d}_0$, the blind ML estimates of α_1 and α_2 are thus

$$\widehat{\alpha}_p = \frac{\mathbf{s}^H\left(\widehat{d}_0\right)\mathbf{r}_p}{\|\mathbf{s}(d_0)\|^2}, p = 1, 2. \quad (44)$$

Appendix C
Beamforming Structure for Nulling the Desired Signal

Based on the estimates of the desired signal's channel gains $\widehat{\alpha}_p, p = 1, 2$, a simple beamforming structure with a weighting vector of $\mathbf{g} = [\widehat{\alpha}_2 - \widehat{\alpha}_1]^T$ can be employed to null a signal with these gains. Specifically, the output from this beamforming is

$$y_n = \mathbf{g}^T\mathbf{r}_n, \quad (45)$$

where $\mathbf{r}_n = [r_{1,n} r_{2,n}]^T$ has forms given by (4). Thus, if the estimated channel gains $\widehat{\alpha}_p, p = 1, 2$ are indeed closed to the actual channel gains $\alpha_p, p = 1, 2$, the desired signal will be closed to being perfectly or completely rejected.

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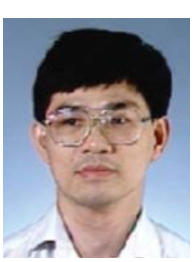

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