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Optimum Combining for Coherent FFH/DS Spread Spectrum Receivers in the Presence of Multi-Tone Jammer

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ABSTRACT Fast Frequency Hopping/Direct Sequence (FFH/DS) hybrid spread spectrum is an effective technique of the communication network for the robot swarms operating in complex electromagnetic environment. It combines the advantages of both Frequency Hopping Spread Spectrum (FHSS) and Direct Sequence Spread Spectrum (DSSS), and also uses the diversity-combining in time and frequency domain to effectively reject unwanted jamming. Multi-Tone Interference (MTI) is one of the most deleterious jamming waveforms for frequency hopping systems, and consequently a severe threat for FFH/DS systems. In this article, we explore the effect of MTI on coherent FFH/DS, and we have shown that, if spectral sensing mechanism for MTI frequency and power monitoring is available, the MTI in FFH/DS reception is a set of complex sinusoid components with deterministic amplitudes and random phases, instead of the conventionally assumed Gaussian random variables. In light of this new finding, a novel combining method is proposed. It utilizes the spectral sensing information to adaptively compute the combining weights, following Maximum Signal to Interference plus Noise (MSINR) law. Its combining performance is analyzed and compared with other existing methods, and verified by simulations. It is shown that the proposed adaptive MSINR combining has satisfactory performance under both medium and strong MTI scenarios.

INDEX TERMS Diversity reception, jamming, spread spectrum communication.

I. INTRODUCTION

In recent years, swarm robotics has been taking off in areas such as surveillance, emergency rescue, and military applications. Robot swarms can offer a variety of obvious advantages over human operators, since they are capable of eliminating the risk of injury, and also assimilating high-volume data at a speed far exceeding human ability [1]. Robot swarms also aim at achieving challenging tasks or significantly improving mission performance compared with a single robot, which demands consensus and cooperation among robots [2].

Many multi-robot coordination problems require intra-swarm shared situational awareness [3], which is most likely achieved by wireless communication techniques. Due to the Size, Weight, And Power (SWAP) constraints [4] in swarm robot design, the wireless communication operates at low transmitting power over a remote area, so it is extremely vulnerable to jamming at the physical layer. Given the potential operational environment, wireless communication must be reliable and robust against unintentional disturbance and hostile electronic countermeasure. So a wireless communication network with physical layer anti-jamming ability is crucial for robot swarm coordination.

Spread spectrum is a effective technique against hostile jamming attack and unintentional interference. Frequency Hopping/Direct Sequence (FH/DS) hybrid spread spectrum, by combining the merits of two types of major spread spectrum techniques, plays an important role in both commercial and military applications for its strong resistance against intentional jamming. Based on the relative proportion of frequency hopping duration to symbol duration, FH/DS systems are divided into two types: slow and fast. In a Slow Frequency

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Hopping/Direct Sequence (SFH/DS) system, the hopping duration, in which the hopping carrier dwells on a certain frequency bin, is larger than the symbol duration. A Fast Frequency Hopping/Direct Sequence (FFH/DS) system, by contrast, has shorter frequency hopping duration than symbol duration.

MultiTone Interference (MTI) is one of the most commonly encountered interference waveforms for wireless communication, and also the most deleterious one for frequency hopping systems. By placing strong tones on operating frequency band, a MTI jammer tries to disturb as many frequency bins of the hopping pattern as possible. For a SFH/DS receiver,if one of the jamming tones falls into the transmitting frequency bin, a sequence of symbols are corrupted. For a FFH/DS receiver, by dividing the symbol power into plural frequency bins, it gains an advantage of diversity to disperse and randomize the jamming. In order to fully utilize the diversity gain, the signal components on multiple frequency bins have to be combined by the receiver, and the combining method affects the performance of reception.

A handful of FFH/DS combining methods for rejecting jamming have been brought out in literature. The most commonly used combining method is Equal Gain Combining (EGC) by summing up all the frequency bins equally, as suggested in [11] and [14]. In [12], authors proposed a Selective Combining (SC) scheme based on perfect side information, in which the receiver detects and discards all the jammed frequency bins, and then combines all jamming-free bins equally. Authors of [13] investigated a Digital Adaptive Gain Control (DAGC) combining, also known as noise normalization combining, in which the received power on all the frequency bins are normalized by a DAGC circuit.

In literature, it is usually assumed that the tones of MTI are whiten into band-limited Gaussian noises with a uniform Power Spectral Density (PSD) [7] by the FFH/DS receiver, which means that, when a jamming tone that hits the hopping frequency bin, the despread procedure spreads its spectrum with a local Pseudo-Random (PN) sequence, resulting it behaving like a band-limited white noise with a bandwidth roughly equal to the chip rate.

The whitening assumption suggests that the local waveforms in despread are totally random, resulting the responses of MTI unpredictable. However, for a receiver, the locally generated PN sequences are actually deterministic. With nowadays spectrum sensing and signal detection techniques, the receiver can easily monitor the frequency band and estimate the power distribution of the MTI. Therefore, the responses of MTI are predictable.

Based on this perception, in this article, we propose a novel combining method for FFH/DS systems in the presence of MTI. This method adaptively adjusts its combining weights to maximize the output Signal to Interference plus Noise Ratio (SINR). We attempt to approximate the performance of proposed combining method and verify its performance through Monte-Carlo simulations. And the influence of incorrect estimation of MTI is also discussed.

The remainder of this paper is organized as follows. The system model, MTI model, and the combining method is introduced in Section II, Section III, and Section IV. The performance of proposed method is analyzed and compared with other methods in Section V. The effect of estimation errors are analyzed in Section VI. Eventually the Monte-Carlo simulations are presented in Section VII.

II. SYSTEM MODEL

Assume s is a phase shifted symbol with constant power as $|s|^2$. It is modulated by K frequency bins/time slots with $\{f_k|k = 0, 1, \ldots, K-1\}$ as the frequency hopping pattern and $\{c_k(t)|k = 0, 1, \ldots, K-1\}$ as the power normalized PN sequence waveforms. The symbol duration T_s and frequency hopping duration T_h are proportional by the pattern size K, as $T_s = KT_h$. The transmitted signal $x(t)$, as indicated in Figure 1, is given by

$$x(t) = s\sum_{k=0}^{K-1} cos(2\pi(f_c + f_k)t)c_k(t - kT_h), \tag{1}$$

in which f_c represents the central frequency of system. To maintain the carrier phase continuous, all the frequencies in hopping pattern $\{f_k\}$ are chosen such that $T_h(f_c + f_k)$ is always integral.

Here we consider the robot swarm operating aerially within a range from several kilometers to tens of kilometers. Thus the transmitted signal goes through a direct line-of-sight channel under presence of a MTI jammer, before arriving the receiver. We make a practical assumption that the total power of the MTI jammer is finite, hence the best strategy of the jammer is to disturb as many frequency bins as possible [5]. In the worst case, every sinusoid components of the MTI jamming is located right on a frequency bin of the hopping pattern, also known as single-tone per band strategy. The received signal under that circumstance is given by

$$\begin{aligned} r(t) = \sqrt{2A_c}s\sum_{k=0}^{K-1} cos(2\pi(f_c + f_k)(t - \tau))c_k(t - kT_h - \tau) \\ + \sum_{k=0}^{K-1} \sqrt{2J_k}cos(2\pi(f_c + f_k + f_{j,k})t + \theta_{j,k}) + n_r(t). \end{aligned} \tag{2}$$

In the expression above, A_c denotes the FFH/DS signal power arriving receiver frond end. τ is the transmitting delay of the channel. J_k indicates the existence and power of the jamming tone of kth frequency bin in the hopping pattern, which has an in-band frequency $f_{j,k}$ and a random phase $\theta_{j,k}$. $n_r(t)$ represents the additive Gaussian white noise with its single-sided PSD as n_0 Watt per Hz.

Figure 1 illustrates the structure of a coherent FFH/DS receiver. Here we assume that synchronization has been achieved. The signal $r(t)$ is firstly demodulated from system central frequency f_c, and then coherently dehopped by the local hopping carrier $exp(-i2\pi f_k t + i\hat{\theta}_k)$, in which i is the

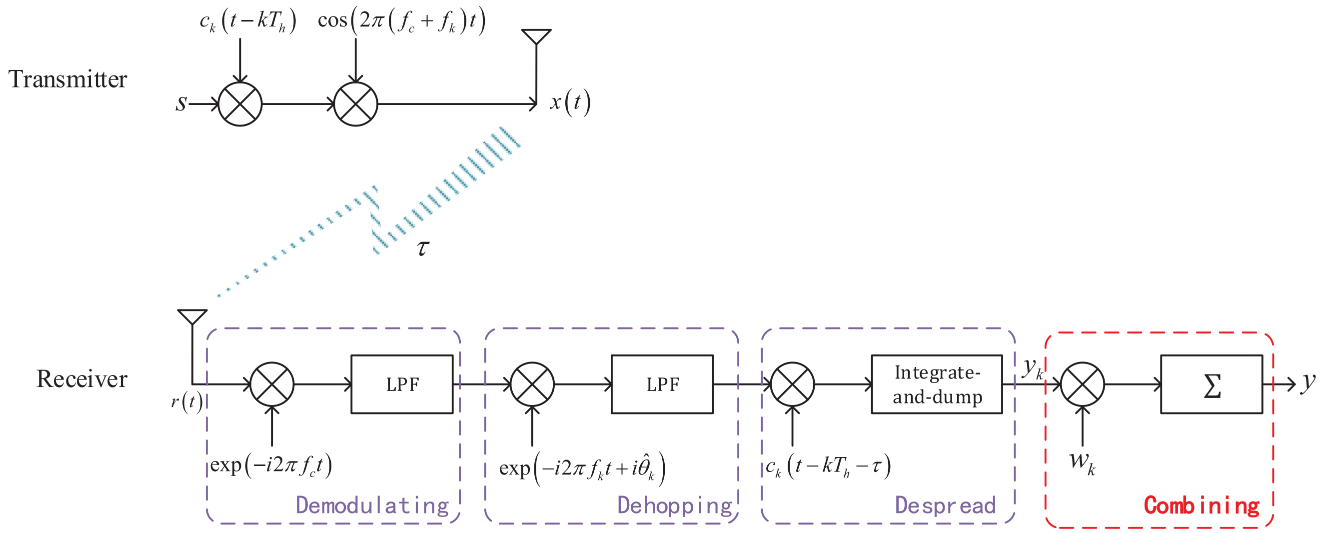

FIGURE 1. A flowchart of the FFH/DS transmitter and receiver.

imaginary unit as $i = \sqrt{-1}$, and $\hat{\theta}_k$ denotes the dehopping phase, as

$$\hat{\theta}_k = 2\pi(f_c + f_k)\tau. \tag{3}$$

The coherent dehopping of a fast frequency hopping system had once been regarded as infeasible, due to the deficiency of high quality frequency synthesizers. Nowadays, extremely fast hopping speed DDS synthesizer is rapidly becoming an alternative to the traditional frequency-agile analog-based PLL synthesizers, and this technological improvement makes the coherent FFH/DS achievable [6].

After being dehopped, the signal is then despread by the synchronized PN sequence waveforms $c_k(t - kT_h - \tau)$. Since the PN sequence waveforms are assumed power normalized, they obey

$$\int_0^{T_h} |c_k(t)|^2 dt = 1. \tag{4}$$

The despread results of K bins are then sampled and combined. Since the reception is coherent, to achieve constructive combining, all the combining weights $\{w_k|k=0, 1, \ldots, K-1\}$ are non-negative real numbers. For the ease of analysis, the sum of combining weights is normalized, given as

$$\sum_{k=0}^{K-1} w_k = 1. \tag{5}$$

Therefore, the result of combining is

$$y = \sqrt{A_c}s + \sum_{k=0}^{K-1} w_k N_k + \sum_{k=0}^{K-1} w_k M_k, \tag{6}$$

in which N_k and M_k are the noise and MTI components of kth frequency bin. Assuming the Low Pass Filters (LPF) as ideal with 0dB attenuation in its passband, based on the waveform normalization (4), N_k follows a complex Gaussian distribution with mean of zero and variance of n_0. The characteristics of MTI components $\{M_k\}$ will be explained in next section.

III. FFH/DS RECEPTION UNDER MTI

In (2), J_k denotes the power of jamming tone falling on kth frequency bin. J_k is either zero (if there is actually no jamming tone on kth frequency bin) or a positive real number that indicates the tone power.

After demodulating and dehopping, the jamming tone becomes $\sqrt{J_k} exp(i2\pi f_{j,k} t + i\theta_{j,k} + i\hat{\theta}_k)$, and then it is despread by the locally generated PN waveform. As illustrated in Figure 1, the despread operation includes two parts: multiplier by $c_k(t)$ and a integrate-and-dump circuit with reset time T_h. So, after despread, MTI components becomes

$$\begin{aligned} M_k &= \int_0^{T_h} c_k(t)\sqrt{J_k} exp(i2\pi f_{j,k} t + i\theta_{j,k} + i\hat{\theta}_k)dt \\ &= \sqrt{J_k} exp(i\theta'_{j,k}) \int_0^{T_h} c_k(t) exp(i2\pi f_{j,k} t)dt, \end{aligned} \tag{7}$$

in which

$$\theta'_{j,k} = 2\pi f_{j,k}(kT_h + \tau) + \theta_{j,k} + \hat{\theta}_k. \tag{8}$$

Despite its property of pseudo randomness, for the authorized receiver, the PN waveform $c_k(t)$ is still a time-limited deterministic signal. It has a frequency representation by Fourier transform function, given as

$$C_k(f) = \int_{-\infty}^{+\infty} c_k(t) exp(-i2\pi f t)dt. \tag{9}$$

We substitute (9) into (7), and then obtain

$$M_k = \sqrt{J_k |C_k(f_{j,k})|^2} exp(i\theta'_{M,k}), \tag{10}$$

in which

$$\theta_{M,k} = 2\pi f_{j,k}(kT_h + \tau) + \theta_{j,k} + \hat{\theta}_k + angle[C_k(-f_{j,k})]. \tag{11}$$

The denotation $angle[\cdot]$ in (11) represents the phase angle of a complex number. Due to randomness of the transmitting

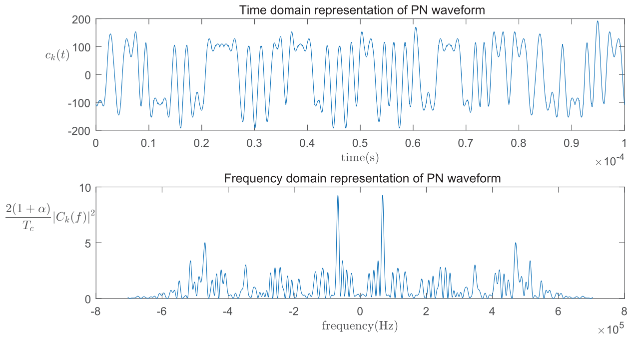

FIGURE 2. The representations of a pseudorandom number sequence.

delay τ, $\theta_{M,k}$ becomes a random variable distributing uniformly from 0 to 2π.

The LPF of dehopping has a passband width corresponding to the bandwidth of PN waveform to maximally reject all kinds of out-band disturbance. Clearly the in-band frequency $f_{j,k}$ of the jamming tone is within the bandwidth of $C_k(f)$, so it goes through the 0dB attenuation passband, resulting the LPF output amplitude as $\sqrt{J_k |C_k(f_{j,k})|^2}$. If the jamming tone power J_k and frequency f_{jk} is estimated correctly, the jamming output amplitude $\sqrt{J_k |C_k(f_{j,k})|^2}$ is knowable.

So when MTI goes through the FFH/DS receiver, it ends up with a set of complex sinusoid variables $\{M_k\}$, whose phase angles obey a uniform random distribution, and amplitudes depend on the power and frequency of the jamming tones. In following we will discuss the properties of MTI output powers $|M_k|^2$.

Conventionally, in the FH/DS hybrid spread spectrum systems, the processing gain of PN waveform in one time slot is assumed to be large enough that the jamming tones are spread into a fairly flat Power Spectral Density (PSD) [7]. PSD indicates the expected value for the spectral power distribution of a random process, yet the PN waveform $c_k(t)$ is actually a shaped waveform of a deterministic sequence with deterministic frequency representation $C_k(f)$. Figure 2 demonstrates the time and frequency representations of a pseudo randomly generated binary sequence with length 128.

The expression of PN waveform $c_k(t)$ is

$$c_k(t) = \sum_{l=0}^{L-1} b_k[l]\, \delta(t - lT_c) * h(t), \tag{12}$$

in which $b_k[l]$ is the PN sequence of kth frequency bin, with length L, chip duration T_c, and chip shaping filter impulse response $h(t)$. $h(t)$ has its frequency domain description $H(f)$ that is bandlimited and determinative to the bandwidth of PN waveform.

Without loss of generality, we assume the PN sequences $b_k[l]$ are binary with values -1 and $+1$. Corresponding to the time domain expression (12), the Fourier transform function is given by

$$\begin{aligned} C_k(f) &= \left(\sum_{l=0}^{L-1} b_k[l] \exp(-i2\pi l T_c f) \right) H(f) \\ &= B_k(f) H(f), \end{aligned} \tag{13}$$

where $B_k(f)$ denotes Discrete Fourier Transform (DFT) of PN sequence $b_k[l]$, as

$$B_k(f) = \sum_{l=0}^{L-1} b_k[l] \exp(-i2\pi l T_c f). \tag{14}$$

Here we do not intent to discuss the specific design and structure of $b_k[l]$. Instead we just believe that the PN sequence is random enough to meet curtain requirements, for example, the DFT test.

DFT test is a powerful evaluation for randomness of pseudo random sequences [8]. This evaluation claims that if a binary sequence is ideally random, the real and imaginary parts of its DFT function are random, independent and both following zero-mean normal distribution. We assume $b_k[l]$ satisfies this description and the truncation effect of the finite time window is negligible. The in-band frequencies of MTI tones are uniformly distributed within the chip shaping filter passband. Therefore the amplitude square $|B_k(f)|^2$ follows chi-squared distribution with 2 degrees of freedom:

$$\frac{2}{L} |B_k(f)|^2 \sim \chi^2(2). \tag{15}$$

The power normalized pulse shaping filter $h(t)$ is assumed to have ideally flat magnitude response within its bandwidth, as given by

$$|H(f)|^2 = \begin{cases} \dfrac{T_c}{L(1+\alpha)}, & \text{if } |f| < \dfrac{1+\alpha}{2T_c}, \\ 0, & \text{if } |f| \geq \dfrac{1+\alpha}{2T_c}, \end{cases} \tag{16}$$

where α is the roll-off factor of the filter.

Based on (13), (14), (15), and (16), if the frequency f is a uniformly distributed random variable, the amplitude square of the PN waveform DFT function follows a chi-square distribution as

$$\frac{2(1+\alpha)}{T_c} |C_k(f)|^2 \sim \chi^2(2). \tag{17}$$

So the MTI output powers $|M_k|^2$ obey a prior probability distribution as

$$|M_k|^2 \sim \frac{T_c J_k}{2(1+\alpha)} \chi^2(2), \tag{18}$$

whereas both the real and imaginary parts of M_k have prior probabilities obeying normal distribution with variance of $\frac{T_c J_k}{2(1+\alpha)}$.

IV. THE MAXIMUM SIGNAL TO INTERFERENCE PLUS NOISE RATIO (MSINR) COMBINING

In (6), the combining result y contains two parts: the signal symbol $\sqrt{A_c}s$, and the weighted summation of disturbances, including noise components $\{w_k N_k\}$ and jamming components $\{w_k M_k\}$.

As pointed out by Rice in [9] and verified by Helstrom in [10], the summation of numerous sine waves phased at random has approximately a Gaussian distribution. Accordingly the summation of weighted jamming components approximately follows a complex Gaussian distribution as

$$\sum_{k=0}^{K-1} w_k M_k \sim \mathcal{CN}\left(0, \sum_{k=0}^{K-1} J_k w_k^2 |C_k(f_{j,k})|^2\right). \tag{19}$$

The summation of noise components also follows a complex Gaussian distribution as

$$\sum_{k=0}^{K-1} w_k N_k \sim \mathcal{CN}\left(0, \sum_{k=0}^{K-1} w_k^2 n_0\right). \tag{20}$$

All the N_k and M_k are zero-mean and independent among each other. Because the reception is coherent and uninvolved of any interaction between desired signal, noise and jamming components, their independence remains.

So the received signal y approximately follows a complex normal distribution as

$$y \sim \mathcal{CN}\left(\sqrt{A_c}s, \sum_{k=0}^{K-1} {w_k}^2 \mu_k\right), \tag{21}$$

where μ_k denotes the disturbance power of k's frequency bin as

$$\mu_k = n_0 + J_k |C_k(f_{j,k})|^2. \tag{22}$$

As assumed, there is some spectral sensing mechanism to monitor the received signal spectrum. The jamming tone powers $\{J_k\}$ and frequencies $\{f_{j,k}\}$, along with the environment noise spectral density n_0, are known. So the disturbance powers $\{\mu_k\}$ of all hopping frequency bins can be calculated, and the Signal to Interference plus Noise Ratio (SINR) of combining result y is given by

$$\rho = \frac{A_c|s|^2}{\sum_{k=0}^{K-1} {w_k}^2 \mu_k}. \tag{23}$$

The symbol s is phase shift modulated, so the combined symbol power $A_c|s|^2$ in (23) is fixed. Maximizing the SINR is thus equivalent to minimizing the weighted disturbance summation $\sum_{k=0}^{K-1} {w_k}^2 \mu_k$, under the constraint of combining weight normalization as in (5).

According to the Cauchy-Schwarz inequality, there is

$$\sum_{k=0}^{K-1} \left|\frac{1}{\sqrt{\mu_k}}\right|^2 \sum_{k=0}^{K-1} |w_k \sqrt{\mu_k}|^2 \geq \left|\sum_{k=0}^{K-1} w_k\right|^2. \tag{24}$$

Based on (5) and (24), a lower limit appears as

$$\sum_{k=0}^{K-1} {w_k}^2 \mu_k \geq \frac{1}{\sum_{k=0}^{K-1} {\mu_k}^{-1}}, \tag{25}$$

in which the equality holds if and only if

$$\mu_0 w_0 = \mu_1 w_1 = \mu_2 w_2 = \ldots = \mu_{K-1} w_{K-1}. \tag{26}$$

So, according to (26) and (5), to achieve Maximum Signal-to-Interference-plus-Noise Ratio (MSINR), the combining weights must obey

$$w_k = \frac{{\mu_k}^{-1}}{\sum_{k=0}^{K-1} {\mu_k}^{-1}}. \tag{27}$$

And the MSINR is

$$\rho_{max} = A_c\, |s|^2 \sum_{k=0}^{K-1} \mu_k^{-1}. \tag{28}$$

V. THE DIVERSITY GAIN

In this section, we will firstly present the output Signal to Noise Ratio (SNR) of proposed method, and then briefly review existing combining methods of FFH/DS receiver, including equal gain combining, selective combining, and digital automatic gain control combining. Their performances of noise resistance and jamming resistance are given out and compared with the proposed MSINR method.

According to (6), (20), and (27), the SNR of proposed method is

$$SNR_m = \frac{A_c|s|^2}{n_0} \frac{\left(\sum_{k=0}^{K-1} {\mu_k}^{-1}\right)^2}{\sum_{k=0}^{K-1} {\mu_k}^{-2}}. \tag{29}$$

In the sense of noise resistance, its diversity gain is not optimal, but we will show that the diversity gain loss is well rewarded with outstanding jamming resistance for both medium and strong MTI.

A. EQUAL GAIN COMBINING

Consider the situation in which jammer is absent, and the weights in (27) becomes

$$w_k = \frac{n_0^{-1}}{\sum_{k=0}^{K-1} n_0^{-1}} = \frac{1}{K}, \tag{30}$$

Because the power of signal and the spectral density of noise are assumed identical in all the frequency bins, the combining scheme in (30) resembles to the Equal Gain Combining (EGC), as all the branches being weighted equally [14].

The SNR of EGC is given by

$$SNR_e = \frac{KA_c\,|s|^2}{n_0}. \tag{31}$$

In such EGC method, owning to frequency diversity and coherent combining, clearly the combining output SNR increases by K times to the single frequency bin output SNR. It is also equivalent to the output SNR of a constant carrier DSSS system with the same DS process gain.

However, EGC performance suffers from the presence of MTI jammer greatly, as its SINR given as

$$\rho_e = \frac{A_c|s|^2}{\sum_{k=0}^{K-1} {w_k}^2 \mu_k} = \frac{KA_c|s|^2}{n_0 + \frac{1}{K}\sum_{k=0}^{K-1} J_k |C_k(f_{j,k})|^2}. \tag{32}$$

There is an additional term in the denominator of (32) that indicates the average power of jammer tones output, and it determines the SINR degradation. As jamming powers getting strong, the EGC performance degrades greatly.

B. SELECTIVE COMBINING

If the MTI power in (27) is much stronger than the thermal noise, the combining weights of jammed frequency bins are much less than the uncorrupted ones, leaving the jammed frequency bins virtually extinguished, similarly to the Selective Combining (SC), which is also known as equal gain combining with perfect side information [11]. In such SC scheme, receiver detects and discards the frequency bins hit by jamming tones, and then adds the signals from hit-free frequency bins equally.

Assuming there are Q bins corrupted by MTI, the combining weights of SC are given as

$$w_k = \begin{cases} \frac{1}{K-Q}, & \text{if the frequency binis not hit by MTI,} \\ 0, & \text{if the frequency bin is hit by MTI.} \end{cases} \tag{33}$$

And the output SNR of SC is

$$SNR_s = \frac{(K-Q)\, A_c\, |s|^2}{n_0}. \tag{34}$$

Clearly the SC is totally immune to jammer, and its SINR stays the same with SNR, without any degradation, as

$$\rho_s = \frac{(K-Q)\, A_c\, |s|^2}{n_0}. \tag{35}$$

But its performance is inherently crippled by the general extinguishment. the receiver suffers from a degradation of $-10\log_{10}((K-Q)/K)$ in decibel, even if the jamming power is weak.

C. DIGITAL AUTOMATIC GAIN CONTROL

As aforementioned, the EGC and SC have shortcomings to accommodate jamming power changing. In [13], authors introduced the Digital Automatic Gain Control (DAGC) diversity combining that can adapt different jamming powers.

TABLE 1. FFH/DS system parameters.

Parameter	Value
Frequency bins number K	32
Jamming tones number Q	28
DS processing gain L	128
FFH hopping duration T_h	10^{-4}s

In this method, a DAGC circuit is added after dehopping, resulting the frequency bins respectively weighted by the reciprocal of their in-band powers.

The weights of DAGC are given by

$$w_k = \frac{\left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-1}}{\sum_{k=0}^{K-1} \left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-1}}, \tag{36}$$

where the denotation L is processing gain of PN code $c_k(t)$, as

$$L = \frac{(1+\alpha)\, T_h}{T_c}. \tag{37}$$

The SNR and SINR of DAGC method are thus given by

$$SNR_d = \frac{A_c|s|^2}{n_0} \frac{\left(\sum_{k=0}^{K-1} \left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-1}\right)^2}{\sum_{k=0}^{K-1} \left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-2}} \tag{38}$$

and

$$\rho_d = \frac{A_c|s|^2 \left(\sum_{k=0}^{K-1} \left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-1}\right)^2}{\sum_{k=0}^{K-1} \left(A_c\, |s|^2 + Ln_0 + T_h J_k\right)^{-2} \left(n_0 + J_k |C_k(f_{j,k})|^2\right)} \tag{39}$$

D. PERFORMANCE COMPARISON

To clarify the performance difference between various combining methods, we assign the FFH/DS system parameters as in Table 1.

The input Jamming to Signal Ratio (JSR) is defined as the power ratio between the signal component and MTI component in (2), given as:

$$JSR = \frac{T_h}{A_c|s|^2} \sum_{k=0}^{K-1} J_k. \tag{40}$$

With jammer of medium power, for example as JSR = 15dB, the output SNR and SINR performances of aforementioned methods is illustrated in Figure 3. In general, the jamming tones with medium power have little contribution in the output disturbance, resulting the respective SINR and SNR in every method approximately identical, except SC method. EGC, DAGC, and proposed MSINR has similar performance,

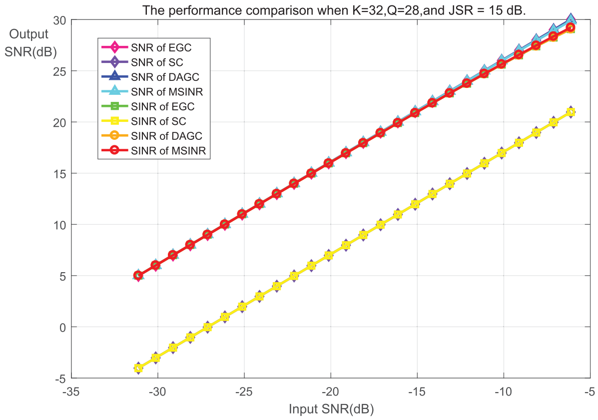

FIGURE 3. **The performance comparison under medium MTI.**

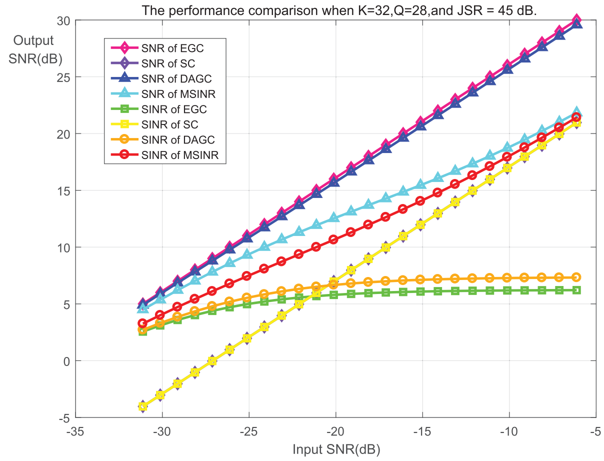

FIGURE 4. **The performance comparison under strong MTI.**

where as SC method suffers from a large gap around 9dB due to the blunt extinguishment.

On the other hand, Figure 4 demonstrates the situation with strong jammer of $JSR = 45B$. In this situation, the jamming tones have dominated influence of output disturbance. Although EGC still has the optimal SNR curve, it has worst output SINR curve. EGC and DAGC also work poorly under strong jamming and have disappointingly low output SINR. The proposed method has the best SINR performance and the second best SNR performance next only to EGC.

The proposed MSINR method has highest output SINR in both medium and strong jamming situations. The advantage is more significant with the presence of strong jamming and severe thermal noise, which makes it a promising solution for FFH/DS system that is required to be reliable and robust against unintentional disturbance and hostile electronic countermeasure.

VI. THE LOSS OWING TO INACCURATE MTI ESTIMATION

The proposed MSINR combining is established on correct estimation of spectral sensing. The environment noise exists chronically with constant power, so it is easier to monitor. On the other hand, the MTI jammer may appear for only a short time and its parameters may change rapidly, in which case, there are possibly some mistakes in its estimation. In this section, we discuss the influence of both frequency and power estimation errors of MTI, and give out the combining gain loss owing to the inaccurate estimations.

A. FREQUENCY ESTIMATION ERROR

When the frequencies information is inaccurate, there is a gap between estimated jamming tone frequency $\hat{f}_{j,k}$ and the true value $f_{j,k}$, and the estimated power of M_k deviates by a ratio ϵ as

$$\epsilon = \frac{J_k|C_k(f_{j,k})|^2}{J_k|C_k(\hat{f}_{j,k})|^2} = \frac{|C_k(f_{j,k})|^2}{|C_k(\hat{f}_{j,k})|^2}. \tag{41}$$

The randomness of the PN codes results $|C_k(\hat{f}_{j,k})|^2$ independent with $|C_k(f_{j,k})|^2$. Both of them obey the same prior probability of chi-square distribution that given by (17), resulting the deviation ratio a variable of centralized F-distribution:

$$\epsilon \sim F(2, 2), \tag{42}$$

and the probability density function of ϵ is denoted by $f_F(\epsilon)$.

Because the numerator and denominator parts of (41) are reciprocal, without loss of generality, we consider that the estimation and combining weights maintain the same as in (27), and instead the output jamming power $|M_k|^2$ shifted by ϵ.

Assuming that there is one jamming tone being inaccurately estimated and it falls on the first frequency bin, the combined SINR becomes

$$\begin{aligned}
\rho &= \frac{A_c|s|^2}{\sum_{k=0}^{K-0} w_k{}^2\mu_k - w_0{}^2(1-\epsilon)J_0|C_0(f_{j,0})|^2} \\
&= \frac{A_c|s|^2}{\left(\sum_{k=0}^{K-0} \mu_k{}^{-1}\right)^{-1} + w_0{}^2(\epsilon-1)J_0|C_0(f_{j,0})|^2} \\
&= \frac{A_c|s|^2 \sum_{k=0}^{K-0} \mu_k{}^{-1}}{1 + w_0{}^2(\epsilon-1)J_0|C_0(f_{j,0})|^2 \sum_{k=0}^{K-0} \mu_k{}^{-1}} \\
&= \frac{\rho_{max}}{1 + w_0{}^2(\epsilon-1)J_0|C_0(f_{j,0})|^2 \sum_{k=0}^{K-0} \mu_k{}^{-1}}.
\end{aligned} \tag{43}$$

The ratio of the maximum SINR and the SINR under frequency estimation error is

$$\frac{\rho_{max}}{\rho} = 1 + w_0{}^2(\epsilon-1)J_0|C_0(f_{j,0})|^2 \sum_{k=0}^{K-0} \mu_k{}^{-1}. \tag{44}$$

The ratio has expectation as

$$\begin{aligned}
&\mathrm{E}\left[\frac{\rho_{max}}{\rho}\right] \\
&= \int_{-\infty}^{+\infty}\left(1 + w_0{}^2(\epsilon - 1)J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}\right)f_F(\epsilon)d\epsilon \\
&= 1 + w_0{}^2 J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}\int_{-\infty}^{+\infty}(\epsilon - 1)f_F(\epsilon)d\epsilon \\
&= 1 + (\mathrm{E}[\epsilon] - 1)w_0{}^2 J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}, \qquad (45)
\end{aligned}$$

where $\mathrm{E}[\epsilon]$ indicates the mean of distribution $F(2, 2)$. Unfortunately the mean does not exist. With the aid of numerical calculation, we give out an approximation as

$$\begin{aligned}
&\mathrm{E}[\epsilon] = 2.092, \\
&\mathrm{E}\left[\frac{\rho_{max}}{\rho}\right] = 1 + 1.092 w_0{}^2 J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}. \qquad (46)
\end{aligned}$$

And the average degradation owing to frequency estimation error is given in decibels by

$$d_f = -\log_{10}\left(1 + 1.092 w_0{}^2 J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}\right). \qquad (47)$$

B. POWER ESTIMATION ERROR

When the power information is inaccurate, the estimated jamming tone power $\hat{J}_k$ differs from the true value J_k. Similarly to the situation of frequency estimate error, there will also be a deviation of the estimated power, given as

$$\frac{\hat{J}_k|C_k(f_{j,k})|^2}{J_k|C_k(f_{j,k})|^2} = \frac{\hat{J}_k}{J_k}. \qquad (48)$$

For a power measurement, it may suffer from both system errors and random errors. The system errors, such as parameter bias and operator bias, are essentially fixed, but varying from measurement to measurement. Without any specification of measurement, the system errors are treated as random variables, and merged into random errors. The normal distribution, for its inherent connection with the laws of physics, is often applicable to equipment parameters emerging from manufacturing processes. For a measurement, unless information to the contrary is available, the normal distribution should be applied as the default distribution [15].

So we assume the MTI power estimation results following the normal distribution, as:

$$\frac{\hat{J}_k}{J_k} \sim \mathcal{N}\left(1, \sigma_{pe}^2\right), \qquad (49)$$

in which σ_{pe}^2 denotes the unitized variance of power estimation. It must be emphasized that in the previous analysis of frequency estimation error, we view the estimation and combining weights maintain the same as in (27), and instead the output jamming power $|M_k|^2$ shifted by ϵ. Here we continue to use this analytical approach. So the shifted ratio ϵ is given as

$$\epsilon = \frac{J_k}{\hat{J}_k}, \qquad (50)$$

and it follows a reciprocal normal distribution with mean 1 and variance σ_{pe}^2. The mean of reciprocal normal distribution does not exist. With the aid of numerical calculation, we give out some of approximated means in Table 2.

TABLE 2. Approximated means of reciprocal normal distribution.

σ_{pe}^2	0.25	0.16	0.09	0.04	0.01
$\mathrm{E}[\epsilon]$	1.8655	1.4740	1.1591	1.0463	1.0103

Substituting the approximated means in Table 2 into (45), the average SINR degradation can be obtained. For example, if the power measurement error has unitized variance 0.25, the average SINR degradation is

$$d_f = -\log_{10}\left(1 + 0.8655 w_0{}^2 J_0|C_0(f_{j,0})|^2\sum_{k=0}^{K-0}\mu_k{}^{-1}\right). \qquad (51)$$

VII. MONTE-CARLO SIMULATIONS

In this section, Monte-Carlo simulations are conducted, to demonstrate the Bit Error Ratio (BER) performance of proposed MSINR combining method under different situations, and also further verify its advantage over other existing methods.

A. DEGRADATION OF RECEIVER SENSITIVITY

As previously analyzed, MTI acts as an additional disturbance to the receiver. It contributes to noise floor increase. Consequently the receiver sensitivity degrades. Sensitivity degradation of proposed MSINR combining method under presence of MTI is verified by the Monte-Carlo BER simulations. The simulation FFH/DS system parameters are listed in following Table 3.

TABLE 3. Simulation parameters.

Parameter	Value
Modulation scheme	Coherent BPSK
Spread spectrum scheme	FFH/DS
DS processing gain L	128
Frequency bins number K	32
FFH hopping duration T_h	10^{-4}s
Symbol duration	3.2×10^{-3}s

The proposed MSINR combining method is implemented in a coherent BPSK-FFH/DS receiver with 32 frequency bins/time slots per symbol. Q out of 32 frequency bins are corrupted by MTI jammer with single-tone per band strategy. Figure 5 and Figure 6 compare the BER performance under medium and strong jamming for $JSR = 15dB$ and $JSR = 45dB$ respectively. The jamming tone number Q varies from 2 to 28. The results under absence of jammer is provided as a benchmark. It can be observed that, under medium jamming as

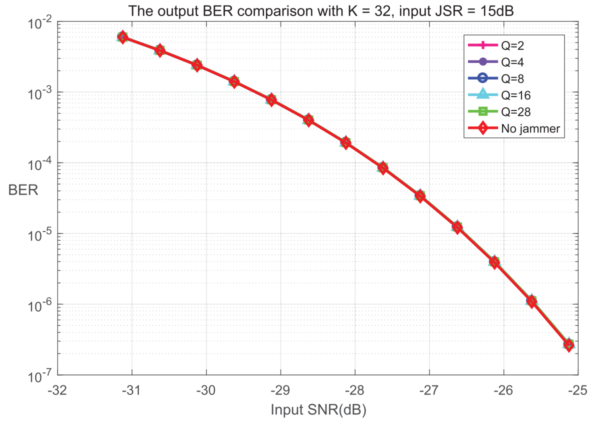

FIGURE 5. BER comparison under medium MTI jammer.

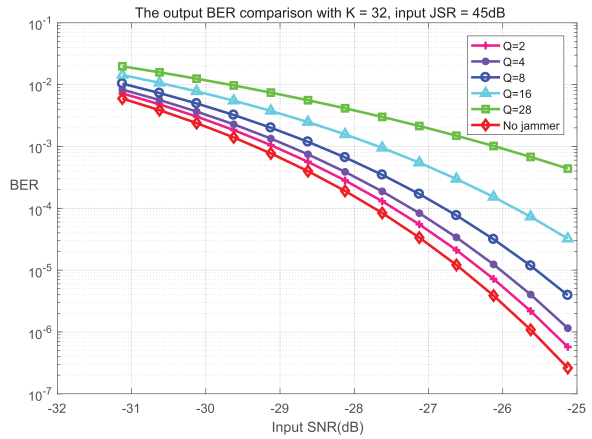

FIGURE 6. BER comparison under strong MTI jammer.

$JSR = 15dB$, the receiver is hardly influenced, with all the curves basically superpose with the benchmark. On the other hand, under strong jammer as $JSR = 45dB$, the BER degradation gets worse when the more bins get jammed. It is consistent with the concept that, the most effective strategy of jammer is to disturb as many frequency bins as possible. So the simulation results of $Q = 28$ give the worst performance, and show a degradation about $3.5dB$ at the receiver sensitivity for $BER = 10^{-3}$.

B. PERFORMANCE COMPARISON

In Figure 8 and Figure 7, the simulation BER results for the proposed MSINR method are compared with other methods introduced in Section V, including EGC, SC, and DAGC. The system parameters are consistent with Table 3, and 28 out of 32 frequency bins are corrupted by relatively medium and strong jamming with $JSR = 15dB$ and $JSR = 45dB$ respectively. The results under absence of jammer is provided as a benchmark. It can be observed in Figure 7 that EGC and DAGC methods are similar to MSINR method for immune the medium jamming, whereas the SC has inherent $9dB$ degradation for its general extinguishment strategy. Under relatively strong jamming, as shown in Figure 8, the proposed MSINR combining method is significantly better than the other methods, and shows the lest degradation of receiver sensitivity as about $3.5dB$ at $BER = 10^{-3}$, whereas other

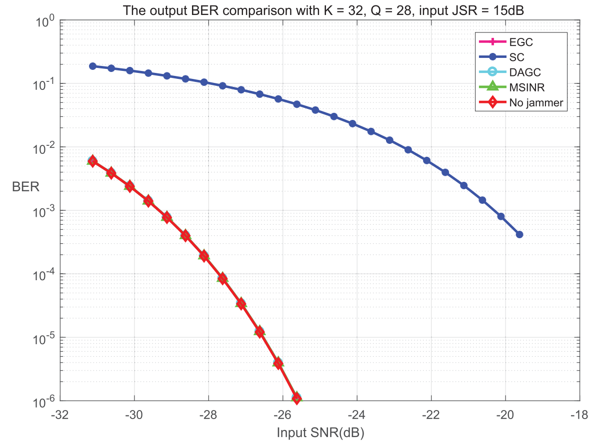

FIGURE 7. The output BER comparison with medium MTI jammer.

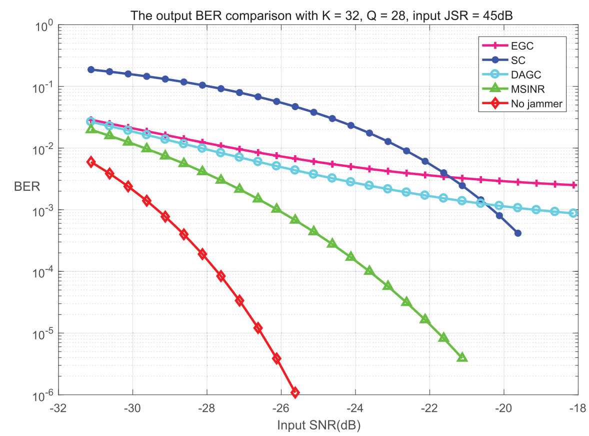

FIGURE 8. The output BER comparison with strong MTI jammer.

methods all suffer from degradation no less than $9dB$. And it is noteworthy that the BER curve of SC method crossover with EGC and DAGC curves, making it the second satisfying only to the proposed MSINR method under relatively high SNR and JSR condition.

VIII. CONCLUSION

In this article, we proposed a novel combining method of FFH/DS system based on Maximum Signal-to-Noise-plus-Interference Ratio (MSINR) law. A closed-form expression of the output SINR is presented. And the combining gain and the average degradation due to inaccurate estimation are analyzed. The innovation of this combining method is utilizing the knowledge of spectral sensing and the deterministic characteristics of the PN waveform to achieve maximal output SINR under MTI. The proposed method is compared with existing methods in literature, and according to results of analysis and simulation, the proposed method has outstanding performance on output SINR and BER under both medium and strong jamming condition. Its advantages are significant under weak signal and strong jammer condition, making it suitable for robot swarm wireless network operating in harsh environment. This article also provides a new viewpoint to the MTI in FFH/DS reception, which can be helpful for further study in areas such as hopping pattern designing and PN code designing.

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