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Performance Evaluation of Coded Hybrid Spread Spectrum System under Frequency Selective Fading Channel

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ABSTRACT
Spread Spectrum technology can be considered as one of the most important communication technologies. Its basic idea is depending on enlarging the modulated signal Bandwidth throughout a second modulation. For the hybrid spread spectrum, most of the researches and applications focused on the Direct Sequence/Frequency Hopping Spread Spectrum, on the other hand a few researches deal the Direct Sequence/Frequency Hopping/Time Hopping hybrid spread spectrum systems. In this study, the performance of a coded Direct Sequence/Fast Frequency Hopping/Time Hopping system is analyzed and evaluated under the additive white Gaussian noise and frequency selective fading channel. The results show that the presented system provides a great performance enhancement due to its multi-spread spectrum techniques.

Key words: Hybrid spread spectrum, direct sequence spread spectrum, frequency hopping, time hopping

INTRODUCTION
Since the Federal Communication Commission (FCC) has been established, its main basis in managing the spectrum allocations was depending on the request-by-request basis. By increasing the applications, it has realized that it had no more spectra to allocate (Billa *et al.*, 2012), which gave rise to spread spectrum techniques as one of the available alternative solutions. The spread spectrum can be defined as the transmission way in which the signal occupies a bandwidth much wider than its original band width. This wider bandwidth is obtained by using certain spreading code completely independent on the message signal and has much higher bit rate than the original signal. This spreading code is synchronized in the receiver to retrieve the original signal throughout the despreading process (Nawkhare *et al.*, 2013). Generally, the spread spectrum techniques can be classified according to the spreading code injection way to the information. The main types of spread spectrum are the Direct Sequence Spread Spectrum (DSSS), chirp Spread Spectrum (CSS), Frequency Hopping Spread Spectrum (FHSS), Time Hopping Spread Spectrum (THSS) and hybrid spread spectrum (Rappaport, 2009).

In the DSSS technique, the original signal bandwidth enlarged throughout injecting high rate spreading code. This is performed by a direct multiplication between the original waveform and the spreading code. DSSS has a lot of applications in many fields especially in the underwater acoustic communications (Guasong *et al.*, 2010; Chen *et al.*, 2012; Guosong and Hefeng, 2012). The FHSS has a different idea; the spreading process is performed by switching the carrier frequency (hopping) periodically from one to another according to a specific spreading code. Generally, the hopping carrier frequencies are spaced apart with the same data modulation bandwidth (Chen, 2007). Since, the FH has an efficient effect against the fading channels, in addition to its ability to counter narrowband interference, a lot of applications used this technique especially in the military field and for the security purposes (Fu and Chen, 2013). Chirp modulation technique has the concept of increasing the frequency ('up-chirp') or decreasing the frequency ('down-chirp') with the time. THSS has the same idea of the FHSS, meanwhile the hopping process performed in the time domain according to a specific code. Nowadays, DS/FH hybrid spread spectrum technique is discussed throughout many study and has been used widely in communication system as a reference, for example in Long *et al.* (2012), the author discusses the increase of the spectrum efficiency by using the scheme combining conventional M_1-ary orthogonal spread spectrum with M_2-ary phase shift keying, it is found that this method provide more anti-jamming abilities than DS/FH system in the same spectrum efficiency. Also, some recent researches focused on and discussed the security advantages of the DS/FH hybrid spread spectrum relative to the pure DSSS and FHSS (Martin *et al.*, 2013). On the other hand, the DS/FH/TH hybrid spread spectrum technique didn't take the same chance as it rarely discussed and applied due to its complexity.

In this study the performance of a coded DS/FFH/TH hybrid spread spectrum system is analyzed and evaluated under AWGN in addition to frequency selective fading channel. The system constructed from conventional coding technique with Viterbi decoder and three kinds of spread spectrum technique, which are the Direct Sequence Spread Spectrum (DSSS), fast Frequency Hopping Spread Spectrum (FFHSS), in addition to Time Hopping Spread Spectrum (THSS).

SYSTEM MODEL BLOCK DIAGRAM

The whole construction of the hybrid spread spectrum system can be shown in Fig. 1, 2 which represents the transmitter and the receiver of the system, respectively.

Transmitter: The transmitter system is constructed from the baseband data unit followed by encoder (error correcting code) unit with rate equal to 2. The output of this encoder is spreaded

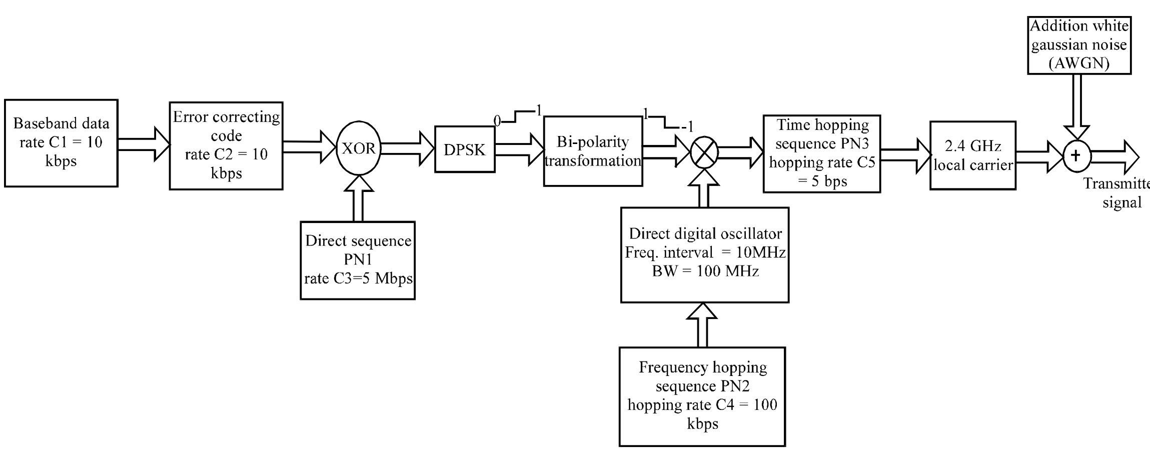

Fig. 1: Block diagram of the hybrid spread spectrum transmitter

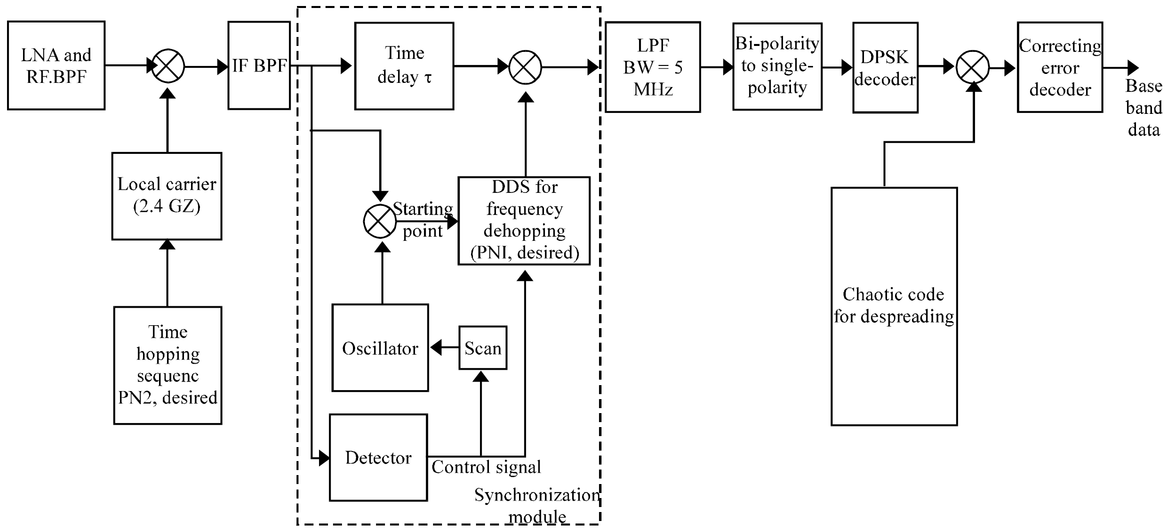

Fig. 2: Block diagram of the hybrid spread spectrum receiver

throughout the spreader unit, which generates a long code PN1 with spreading gain 250. The spreaded signal is modulated using Differential Phase Shift Keying (DPSK) module to avoid the ambiguity of the phase. The modulated signal then will have a bi-polarity transformation and then the hopping frequency is injected through the direct digital oscillator, after that the frequency hopped signal will be transmitted in hopped time windows by using the time hopping unit to the noisy channel (Mansour and Fu, 2013).

Receiver: The receiver system is constructed mainly from a low noise amplifier in addition to radio frequency band pass filter unit. The output of this unit is entering to a time hopping gate represented by the local carrier frequency synthizer which is controlled by the time hopping sequence PN3. The following module is the synchronization module, which constructed from the time delay, oscillator, detector and scan unit. The output of this module becomes the input of the frequency dehopping module. After that a low pass filter is used and the signal is converted from a bi-polarity signal to a single-polarity signal. A DPSK demodulation is performed by the DPSK decoder unit, which prepare the signal to the dispreading process by using the code sequence PN1. Finally the base band data is retrieved by using the error correction code decoder unit.

The synchronization process between the received signal and the locally generated hopping frequencies is performed by the Synchronization Module, to achieve the frequency dehopping process. The main idea is depending on transmitting an independent signal other than the hopping frequencies and with lower frequency value. This frequency is acting as a guide frequency as shown in Fig. 3, in which a version of the received signal becomes an input to the detector. The function of the detector unit is to detect the received guide frequency using the FFT technique and its output is a control signal to control the scan unit and the Direct Digital Synthesizer unit (DDS).

The scanning process is performing by correlating locally generated frequencies with the received signal. As soon as the correlation becomes maximum, the scan unit stops scanning and it locks the oscillator on the detected desired frequency. Also, there is a controlled signal sent from the detector unit to the Direct Digital Synthesizer unit (DDS) to make the frequency image selection to precede the frequency dehopping process. It is very important to identify the frequency starting

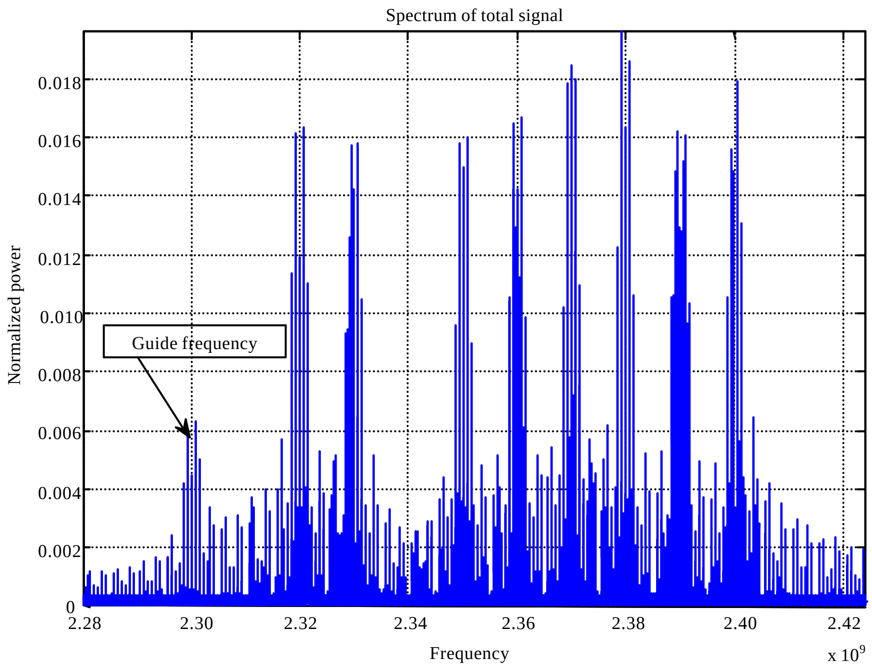

Fig. 3: Received signal spectrum including the guide frequency

point, which ensures the synchronization process. Since, these processes consume a time, this time must be taken into consideration. The time delay unit calculates this processing time and delay the receiving signal to coincident with the locally hopping frequencies generated from the DDS.

MATHEMATICAL SYSTEM MODEL

For simplicity, the analysis of the transmitter system is briefly presented, since it discussed in detail in our previous study (Mansour and Fu, 2013). To identify the mathematical model of the given system, let the baseband data be represented as:

$$d(t)=\sum_{i=0}^{M-1} x(t-imT_s) \tag{1}$$

where, M is the No. of bits generated during the assigned time frame, x (t) is the rectangular pulse with duration mT_s, m is the No. of samples in one bit duration and T_s is the sampling time which equals to the inverse of sampling frequency F_s. In the case of Additive White Gaussian Noise, the total received signal at the receiver front panel can be expressed as:

$$R(t)=\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1} z_k(t-T_{ss}[i_c N+j])\mathrm{Re}\left[\exp\left(J2\pi[f_c-f_p]\frac{m_c}{G_h}l_h T_s\right)\right]W_k(l_t T_s)+n(t) \tag{2}$$

where, Mc is the No. of coded bits generated during the same assigned time frame such that M_c = 2M, m_c, is the No. of samples in one coded bit duration with m_c = m/2. z(t) is also a rectangular pulse with the same duration nT_s, N is spreading code length. f_c is the carrier frequency taken to be 2.4 GHz. f_p is the hopping frequency taking the values of {0, f_h, $2f_h$, $3f_h$,... ,$7f_h$} according to the controlled code sequence PN2, where f_h is the frequency interval which sets to be 10 MHz. G_h. is the hopping spreading gain, which is the ratio between the hopping rate and the coded bit rate. W (l_t T_s) is the time window in which the signal will be transmitted and n (t) represents the AWGN component.

According to the receiver block diagram, the first block facing the received signal is the low noise amplifier, which its output can be expressed as:

$$R_a(t)=a\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1}z_k(t-T_{ss}[i_cN+j])\mathrm{Re}\left[\exp\left(J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right]W_k(l_tT_s)+an(t) \tag{3}$$

where, is the amplification factor applied by the low noise amplifier. Since, the time hopping process is the last process in the transmitter, it will be the first process in the receiver,the receiver will enable the time window to recover the time hopping process, so:

$$R_{TH}(t)=a\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1}z_k(t-T_{ss}[i_cN+j])\mathrm{Re}\left[\exp\left(J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right]\_\,an(t)W_k(l_tT_s) \tag{4}$$

Which mean that the received signal is received in the time duration specified by the W_k (l_tT_s) function, which defining which PN3 will be used according to the unified frequency as mentioned before. The following step is to make the dehopping process, assuming a complete synchronization, the De hopped signal can be expressed as:

$$R_{FH}(t)=a\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1}z_k(t-T_{ss}[i_cN+j])+an(t)W_k(l_tT_s)\mathrm{Re}\left[\exp\left(J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right] \tag{5}$$

This equation means that the original signal is recovered to its bi-polar spreaded form, while the noise spectrum component is frequency hopped on the frequency hopping pattern specified by the code PN2. The despreading process is performed after that to recover the original coded signal as:

$$R_{DS}(t)=a\sum_{i_c=0}^{M_c-1}z_k(t-i_cm_cT_s)+a\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1}n(t-i_cNnT_s-jnT_s)W_k(l_tT_s)\mathrm{Re}\left[\exp\left(-J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right] \tag{6}$$

This equation shows that the original coded data is reconstructed by the despreading process, while the noise portion is spreaded throughout the long code and according to the code PN1. This procedure is the main reason of the system performance enhancement, in which the DS spreading gain provides the advantage of the performance. The final stage is the error correction and decoding process, which recover the original data that expressed as:

$$R_{DS}(t)=a\sum_{i_c=0}^{M_c-1}x(t-i_c m_c T_s)+a\sum_{i_c=0}^{M_c-1}\sum_{j=0}^{N-1}n(t-2iNnT_s-jnT_s)W_k(l_tT_s)\mathrm{Re}\left[\exp\left(-J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right] \quad (7)$$

This represents the final form of the recovered signal, in which the original data is reconstructed with the noisy part represents the AWGN affected by the decoding, despreading, frequency dehopping and time dehopping processes.

In the case of frequency fading channel, the total received signal at the receiver front panel can be expressed as:

$$R_{DS}(t)=a\sum_{i_c=0}^{M_c-1}\sum_{q=0}^{Q-1}\sum_{i_c=0}^{M_c-1}z_{k,q}(t-T_{ss}[i_cN+j]-\tau_q)\mathrm{Re}\left[\exp\left(J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right]W_k(l_tT_s) \quad (8)$$

where, Q is the No. of the multipath and τ is the multipath time delay. Following the same sequences mentioned before, the final form in the frequency selective fading channel will be:

$$\begin{aligned}R_{FSF}(t)=&\ a\sum_{i_c=0}^{M_c-1}x(t-imT_s)+a\sum_{q=0}^{M_c-1}\sum_{i_c=0}^{Q-2}\sum^{N-1}V_{k,q}(t-T_{ss}[i_cN+j]-\tau_q)\mathrm{Re}\left[\exp\left(J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right]\\ &W_k(l_tT_s)+a\sum_{i=0}^{M-1}\sum_{j=0}^{N-1}n(t-2iNnT_s-jnT_s)W_k(l_tT_s)\mathrm{Re}\left[\exp\left(-J2\pi f(k)\frac{m_c}{G_h}l_hT_s\right)\right]\end{aligned} \quad (9)$$

In this equation, the first term represents the received signal from the main path, while the second term represents the rest of the remaining paths and finally the last term represents the noise component. In the second term, $V_{k,q}$ are the delayed signals versions received from the other multipaths and despreaded by the assigned spreading codType equation here. It is clear that the effect of this part will be attenuated due to the cross correlation between the main spreading code and its shifted versions in the delayed signal versions, in addition to the frequency hopping effect.

ERROR PROBABILITY

Let P_s denotes the channel-symbol error probability, which is the probability of error in a demodulated code symbol. It is assumed that the channel-symbol errors are statistically independent and identically distributed. Let P_w denotes the word error probability, which is the probability that a received word is not decoded correctly due to both undetected errors and decoding failures. There are distinct ways in which i errors may occur among v symbols. Since, a received sequence may have more than t errors but no information-symbol errors:

$$P_w\le\sum_{i=t+1}^{v}\binom{v}{i}P_s^i(1-P_s)^{v-i} \quad (10)$$

Let P_{ud} denote the probability of an undetected error and P_{df} denote the probability of a decoding failure. For a bounded-distance decoder:

$$P_{ud} + P_{df} \sum_{i=t+1}^{v} \binom{v}{i} P_s^i (1-P_s)^{v-i} \tag{11}$$

Thus, it is easy to calculate P_{df} once P_{ud} is determined. If channel-symbol errors in a received word are statistically independent and occur with the same probability P_s, then the probability of an error in a specific set of positions that results in a specific set of i erroneous symbols is:

$$P_e(i) + \left(\frac{P_s}{q-1}\right)^i (1-P_s)^{v-i} \tag{12}$$

In which each symbol is selected from an q alphabet of symbols, for an undetected error to occur at the output of a bounded-distance decoder, the No. of erroneous symbols must exceed t and the received word must lie within an incorrect decoding sphere of radius t. Let N (l, i) is the No. of sequences of Hamming weight i that lie within a decoding sphere of radiusassociated with a particular code word of weight l, then:

$$P_{ud} = \sum_{i=t+1}^{v} P_s(i) \sum_{l=\max(i-td_m)}^{\min(i+t,v)} A_l N(l,i) \tag{13}$$

Where A_l is the No. of code words with weight l. Consider sequences of weight i that are at distance s from a particular code word of weight l, where $|l-i| \le s \le t$. By counting these sequences and then summing over the allowed values of s, we can determine N (l, i). The counting is done by considering changes in the components of this code word that can produce one of these sequences. Let j denote the No. of nonzero code word symbols that are changed to zeros, the No. of codeword zeros that are changed to any of the (q-1) nonzero symbols in the alphabet and β the No. of nonzero code word symbols that are changed to any of the other (q-2) nonzero symbols.

For a sequence at distance s to result, it is necessary that $0 \le j \le s$. No. of sequences that can be obtained by changing any j of the nonzero symbols to zeros is:

$$\binom{l}{i}$$

Where:

$$\binom{b}{a} = 0$$

if a>b. For a specified value of j, it is necessary that α = j+1-l to ensure a sequence of weight i. The number of sequences that result from changing any α of the v-1 zeros to nonzero symbols is:

$$\binom{v-1}{a} (q-1)^a$$

For a specified value of j and hence α, it is necessary that β = s-j-α = s+l-I-2j to ensure a sequence at distance. No. of sequences that result from changing β of the l-j remaining nonzero components is:

$$\binom{l-j}{\beta}(q-2)^{\beta}$$

where, $0^x = 0$ if $x \neq 0$ and $0^0 = 1$ Summing over the allowed values of s and j, it is obtained:

$$N(l,i)=\sum_{s=|t-i|}^{t}\sum_{j=0}^{s}\binom{l}{j}\binom{v-1}{j+i-1}\binom{l-j}{s+l-i-sj}\times(q-1)^{j+i-l}(q-2)^{s+l-isj} \tag{14}$$

Equation 11 and 12 allow the exact calculation of P_{ud}. When q = 2, the only term in the inner summation of (20) that is nonzero has the index j=(s+l-i)/2 provided that this index is an integer and 0≤(s+l-i)/2≤s. Using this result, we find that for binary codes:

$$N(l,i)=\sum_{s=|t-i|}^{t}\begin{pmatrix} v-1 \\ \frac{s+i-1}{2}\end{pmatrix}\begin{pmatrix} l \\ \frac{s+l-i}{2}\end{pmatrix}, q=2 \tag{15}$$

where, for any nonnegative integer m. Thus, N(l, l) = 1 and N(l, i) = 0 for |l-i|≥t+1. Let P_{is} (j) denote the probability of an error in information symbol j at the decoder output. In general, it cannot be assumed that P_{is} (j) is independent of j. The information-symbol error probability, which is defined as the unconditional error probability without regard to the symbol position, is:

$$P_{is}=\frac{1}{k}\sum_{j=1}^{k}P_{is}(j) \tag{16}$$

The random variables Z_j, j = 1, 2, ..., k are defined so that $Z_j = 1$ if information symbol j is in error and $Z_j = 0$ if it is correct. The expected number of information-symbol errors is:

$$E[I]=E\left[\sum_{j=1}^{k}Z_j\right]=\sum_{j=1}^{k}E[Z_j]=\sum_{j=1}^{k}P_{is}(j) \tag{17}$$

where, E denotes the expected value. The information-symbol error rate isdefined as E[I]/k. Equation 22 and 23 imply that:

$$P_{is}=\frac{E[I]}{k} \tag{18}$$

For the Viterbi decoder, Let a(l, i) denote the number of paths of the trellis diagram that diverging at a node from the correct path, each having Hamming weight l and incorrect information symbols over the unmerged segment of the path before it merges with the correct path. Let df denote the minimum free distance, which is the minimum distance between any two code words. Let $E[N_e(j)]$ denote the expected value of the number of errors introduced at node j. Therefore, if there are N branches in a complete path:

$$P_{is}=\frac{1}{kN}\sum_{j=1}^{N}E[N_e(j)] \tag{19}$$

Let $B_j(l, i)$ denote the event that the path with the largest metric diverges at node j and has Hamming weight l and i incorrect information bits over its unmerged segment. Then:

$$E[N_e(j)]=\sum_{i=1}^{I_j}\sum_{l=df}^{D_j}E[N_e(j)/B_j(l,i)]P[B_j(l,i)] \tag{20}$$

where, $E[N_e(j)/B_j(l,i)]$ is the conditional expectation of $N_e(j)$ given the event $B_j(l, i)$, $P[B_j(l, i)]$ is the probability of this event and I_j and D_j are the maximum values of i and l respectively, that are consistent with the position of node j in the trellis. When $B_j(l, i)$ occurs, i bit errors are introduced into the decoded bits; thus:

$$E[N_e(j)/B_j(l,i)]=i \tag{21}$$

Since, the decoder may already have departed from the correct path before node the union bound gives:

$$P[B_j(l,i)] \le a(l,i)P_2(l) \tag{22}$$

where, $P_2(l)$ is the probability that the correct path segment has a smaller metric than an unmerged path segment that differs in l code symbols. Substituting Eq. 18 to 20 into Eq. 17 and extending the two summations to ∞, obtained:

$$P_b \le \frac{1}{k}\sum_{i=1}^{\infty}\sum_{l=d_f}^{\infty} ia(l,i)P_2(l) \tag{23}$$

The information-weight spectrum or distribution is defined as:

$$B(l)=\sum_{i=1}^{\infty} ia(l,i), l \ge d_f \tag{24}$$

In terms of this distribution, Eq. 23 becomes:

$$P_b \le \frac{1}{k}\sum_{i=1}^{\infty} B(l)P_2(l) \tag{25}$$

When the demodulator makes hard decisions and a correct path segment is compared with incorrect one, correct decoding results if the number of symbol errors in the demodulator output is less than half the No. of symbols in which the two segments differ. Assuming the independence of symbol errors, it follows that for hard-decision decoding:

$$P_2(l)=\begin{cases}\displaystyle\sum_{i=(l+1)/2}^{l}\binom{l}{i}P_s^i(1-P_s)^{l-i}, & \text{l is odd}\\ \displaystyle\sum_{i=(l/2+1}^{l}\binom{l}{i}P_s^i(1-P_s)^{l-i}+1/2\binom{l}{i}[P_s(1-P_s)]^{l/2}, & \text{l is even}\end{cases} \tag{26}$$

Table 1: Parameter values of rate 1/2 conventional codes

			B (d_f+i) for i = 0, 1, 2, ..., 6						
K	d_f	Generators	0	1	2	3	4	5	6
3	5	5, 7	1	4	12	32	80	192	448
4	6	15, 17	2	7	18	49	130	333	836
5	7	23, 35	4	12	20	72	225	500	1324
6	8	53, 75	2	36	32	62	332	701	2342
7	10	133, 171	36	0	211	0	1404	0	11,633
8	10	247, 371	2	22	30	148	340	1008	2642
9	12	561, 763	33	0	281	0	2179	0	15,035
10	12	1131, 1537	2	21	100	186	474	1419	3542

Table 2: Parameter values of rate 1/3 conventional codes

			B (d_f+i) for i = 0, 1, 2, ..., 6						
K	d_f	Generators	0	1	2	3	4	5	6
3	8	5, 7, 7	3	0	15	0	58	0	201
4	10	13, 15, 17	6	0	6	0	58	0	118
5	12	25, 33, 37	12	0	12	0	56	0	320
6	13	47, 53, 75	1	8	26	20	19	62	86
7	15	117, 127, 155	7	8	22	44	22	94	219
8	16	225, 331, 367	1	0	24	0	113	0	287
9	18	575, 673, 727	2	10	50	37	92	92	274
10	20	1167, 1375, 1545	6	16	72	68	170	162	340

Table 3: Parameter values of rate 1/4 conventional codes

			B (d_f+i) for i = 0, 1, 2, ..., 6						
K	d_f	Generators	0	1	2	3	4	5	6
3	10	5, 5, 7, 7	1	0	4	0	12	0	32
4	13	13, 13, 15, 17	4	2	0	10	3	16	34
5	16	25, 27, 33, 37	8	0	7	0	17	0	60
6	18	45, 53, 67, 77	5	0	19	0	14	0	70
7	20	117, 127, 155, 171	3	0	17	0	32	0	66
8	22	257, 311, 337, 355	2	4	4	24	22	33	44
9	24	533, 575, 647, 711	1	0	15	0	56	0	69
10	27	1173, 1325, 1467, 1751	7	10	0	28	54	58	54

For binary codes of rates 1/2, 1/3 and 1/4, codes with these favorable distance properties have been determined. For these codes and constraint lengths up to 12, Table 1, 2 and 3 list the corresponding values of d_f and B(d_f+i), i= 0, 1, 2, ... , 7. Also, listed in octal form are the generator sequences that determine which shift-register stages feed the modulo-two adders associated with each code bit. For example, the best K = 3, rate-1/2 code Table 1 has generator sequences 5 and 7.

RESULTS

Figure 4 shows the applying of the BER theoretical equation for the BPSK and the DBPSK with the practical simulation for the raw data without modulation and the practical simulation of the DBPSK modulation. It can be cleared that the practical DBPSK presents about 2.5 dB gain at BER 10^{-4}.

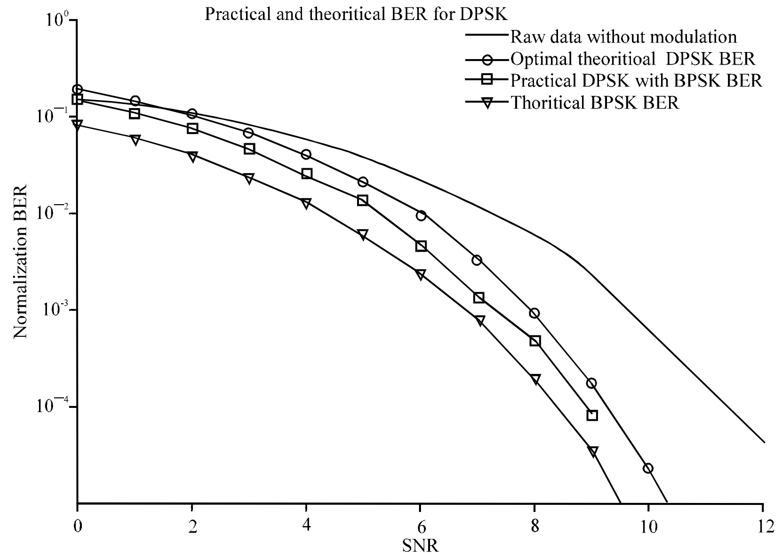

Fig. 4: Practical and theoretical BER for DPSK

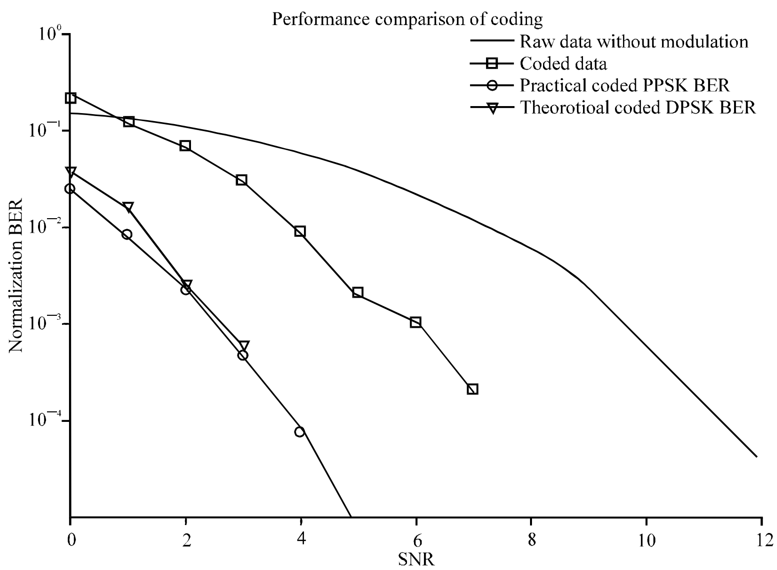

Fig. 5: Practical and theoretical BER for coded and coded DBPSK

Figure 5 discuss the performance comparison of the theoretical and practical BER for the error correction coding. The figure clears that the absolute coding provides about 4 dB gain at BER equal to 10^{-4}, while the combination between the ECC and DBPSK modulation may increase the enhancement to be 4 dB.

Figure 6 shows the performance comparison of the theoretical and practical BER for the whole presented coded hybrid spread spectrum system. The figure shows that the system presents BER equal to 10^{-4} at SNR -23 dB, while the original data without any modulation shown in Fig. 7 presents BER equal to 10^{-4} at approximately SNR 11 dB. This means that there is about 34 dB gain from the presented system, actually this gain is a combination of many various techniques gains.

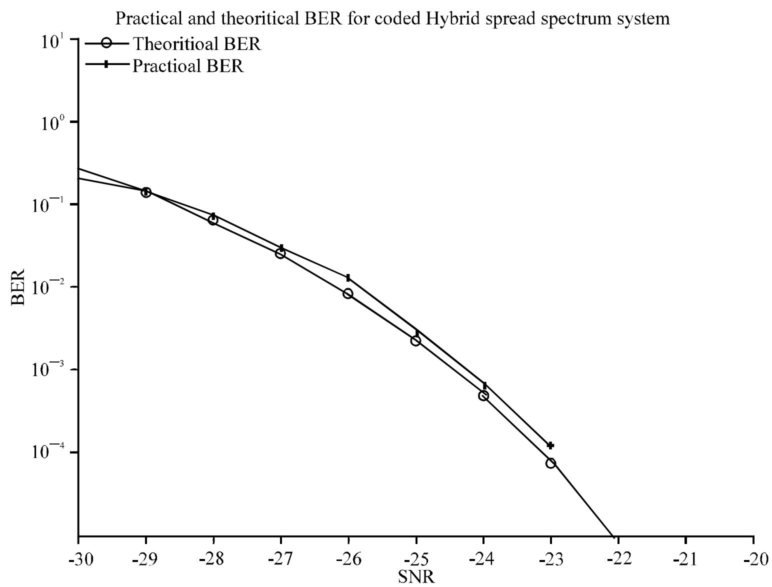

Fig. 6: Practical and theoretical BER for coded hybrid spread spectrum system

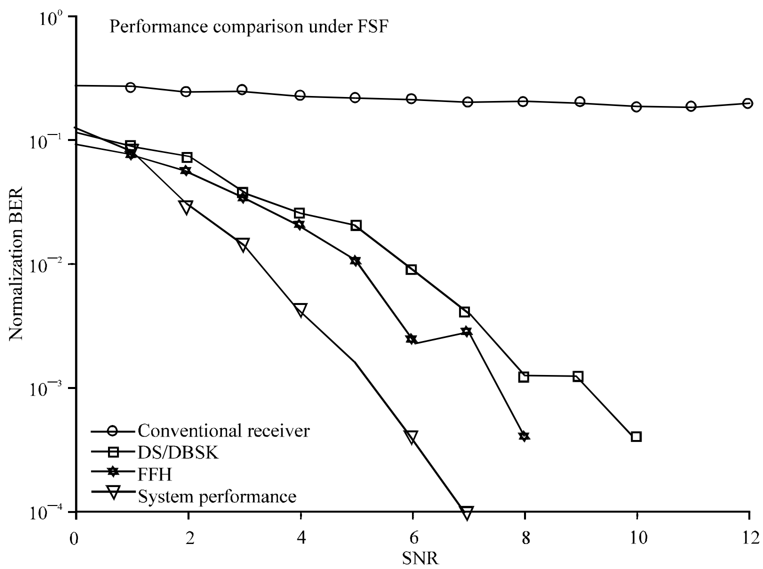

Fig. 7: Performance of the coded hybrid spread spectrum system under frequency selective fading channel

These techniques are mainly the error correction coding, the DBPSK, the DS spread spectrum, the frequency hopping and the time hopping. The error correction coding without any modulation presents about 4 dB gain at BER equal to 10^{-4}, while the coding with DPSK gives about 7 dB gain as shown in Fig. 7 and the DBPSK presents about 2.5 dB. Since the spreading code length equal to 250 and the ECC rate is 2, then the total spreading code will be 500, which gives processing gain about 27 dB. Finally, it is well known that the FH has a very small effect in case of AWGN, since its main effect appears under the fading channel. The figure shows a great coincident between theoretical and practical BER along the presented SNR, which indicates that both the theoretical analysis and the simulation is correct.

In Fig. 7, the performance of the different stages of the proposed system is discussed under the frequency selective fading channel. The figure shows that the performance of the conventional receiver over the frequency selective fading channel is completely failed, since the BER exhibit an error floor of about 10^{-4} over all the SNR range. Meanwhile the performance of the FFH is better than that of the DS by about 1.5 dB at SNR 10^{-3}, while the performance of the presented system provides about 2 dB over the FFH and 3.5 dB over the DS. It is clear that the coding, DS and FFH combination in the presented system enhance the performance and attenuate the effect of the frequency selective fading.

CONCLUSION

In this study, a detail study for a coded hybrid spread spectrum receiver system is presented. The study started with an introduction for the overall system, a detailed explanation for the transmitter and receiver systems is presented. The mathematical model for all the system iscreated for the AWGN and FSF channels. The theoretical BER is derived and compared with the practical system simulation under the AWGN channel. The simulation results show that there are a great coincidence between the theoretical and practical results and also that the presented system have a great performance enhancement especially in the FSF channel due to its multi spreading techniques.

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