Joint iterative channel estimation and decoding under impulsive interference condition

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***Abstract*—Even though Low-Density-Parity-Check (LDPC) code which has the decoding performance close to the Shannon Limit and it is designed as a powerful forward-error-correction (FEC) code in the Additive White Gaussian Noise (AWGN) channel, simulation results show that the performance of LDPC decoder is degraded when exposed to the impulsive noise. According to such a impulsive noise impact, joint iterative channel estimation and decoding technique is proposed in this paper so as to decrease the effect of impulsive interference while less complicated in processing. The proposed methods decreases the complexity by implementing the simple way of channel estimation and applying joint iterative technique between channel estimation and LDPC decoding under two kind of impulsive noise; pulsed radio frequency interference(RFI) and symmetric alpha-stable ($S\alpha S$). In the optimal decoder, channel parameter estimation can be as accurate as possible. Because computed in every time of iterative decoder, channel parameters have been always optimized resulting in the enhancement of LDPC decoder performance.**

***Keyword*—LDPC decoding, pulsed RFI, symmetric alpha-stable, Joint iterative, Channel estimation.**

I. INTRODUCTION

Reliability is an important variable in communication systems. Undesirable signal should be eliminated. Not only planetary communications systems, but also satellite communications systems are commonly interrupted by impulsive noise; pulsed radio frequency interference (RFI) and symmetric alpha- stable noise ($S\alpha S$).

Several researches which investigate the impact of pulsed RFI on LDPC decoder performance found that regardless of the level of signal to noise ratio (SNR), the Bit-Error-Rate (BER) is always large [1]. Pulsed RFI frequently has the duty cycles in the order of 5% or smaller through code length. In addition, pulsed RFI randomly turns up in the form of block and the noise variance varies with time. Currently, several applications, such as ISS ACS Transponders, 4th Gen Transponders, Integrated Receivers and satellite communication receiver, are involved with pulsed RFI, resulting in wide awareness of the impact of pulsed RFI.

$S\alpha S$ distribution frequently used for simulating the impulsive interference. $S\alpha S$ noise is constructed based on two important parameters; characteristic exponent and dispersion parameter, in which different value of the two parameters give different heavy tail noise level. $S\alpha S$ noise has impact on several communication system such as Orthogonal Frequency Division Multiplexing (OFDM) [2].

LDPC codes, invented by Gallager, achieve near capacity performance in a wide class of channels [1]. Generally, various simulations and applications are based on AWGN channel. However, our simulation is unable to consider being only AWGN channel because of pulsed RFI and $S\alpha S$ noise appearing. In impulsive channel, an initial Log Likelihood Ratio (LLR) has been developed in many ways [3] [4] [5] [6]. This work constructs the initial LLR based on probability density function (PDF). Recently, joint iterative technique have been extensively implement in several applications [1] [7] [8] and attractive for mitigating the effect of heavy noise. The proposed method presented the joint iterative between channel estimation and LDPC decoding technique which has been developed to re-processing parameter estimation for constructing the PDF during each iterative time of LDPC decoder processes.

On the one hand, pulsed RFI has close form of PDF and the SNR has been investigated by several researchers [4] to obtain the noise variance and LLR in consequently. Although all of them work very well when the SNR value is large, their performance suffers at negative SNR. The iterative SNR estimation [9] is rest on an iterative solution for the maximum likelihood estimation of the amplitude from which the SNR is competent to compute. Its results show that it exhibits a lower bias and normalized mean squared error than other techniques and that the useful range extends to negative SNR.

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However, the iterative SNR estimation is time-consuming because of demanding a lot of complicated processes and taking a large of calculating time depending on the number of iteration. On the other hand, there is no close form of PDF for $S\alpha S$ distribution, thus SNR cannot implemented to obtain the noise variance. Geometric signal-to-noise ratio (GSNR) is implemented. Some of research works [10] [11] [12] applied Particle Filter (PF) to estimate the important parameters in $S\alpha S$ distribution and get the appreciated Log Likelihood Ratio (LLR) in consequently. Since the number of particle which deploy in PF process, complexity in processing is concern problem. In this work, we implement commonly use of the second order moment (M_2) method [4] to estimate the noise variance under pulsed RFI channel and use of Logarithmic Moment (LM) [13] which has no complicated mathematic formula to estimate the channel parameter under $S\alpha S$ channel.

The rest of this paper is organized as follows. Section II presents the system model of the two channel type; the pulsed RFI, $S\alpha S$, and the influence on LDPC decoder performance. Section III, which reviews channel estimation techniques, consists of M_2 SNR estimation and GSNR estimation. In section IV, the joint iterative channel estimation and decoding approaches are presented for the both channels. The simulation results of the proposed techniques demonstrate in section V. The conclusion is offered in the last section.

II. SYSTEM MODEL

A. Pulsed RFI Model

The system of pulsed RFI channel [1] illustrates as Fig. 1 and (1) where x_k is transmitted sequence and y_k is received sequence. The transmitted signal (x_k) is interrupted by pulsed RFI ($n_{rfi,k}$) and AWGN ($n_{AWGN,k}$) which both of them have zero mean, but possess difference in noise variance. Distribution of AWGN and pulsed RFI are capable of representing as $N(0,\sigma_1^2 = \sigma^2 + \vartriangle\sigma)$ and $N(0,\sigma_2^2)$ respectively. In practical, because most of the channel is time varying SNR mismatch over an AWGN channel, the noise variance is explained as $\sigma_1^2 = \sigma^2 + \vartriangle\sigma$ when σ^2 signify as the noise variance of AWGN and $\vartriangle\sigma$ is symbolic of the random walk signal of AWGN.

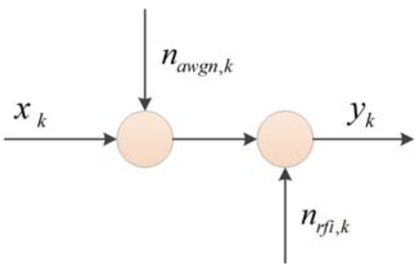

Fig. 1 System model for pulsed RFI experiment.

$$y_k = x_k + n_{awgn,k} + g(t) * n_{rfi,k} \qquad (1)$$

Where

$n_{awgn,k} \sim N(0,\sigma_1^2 = \sigma^2 + \vartriangle\sigma)$

$n_{rfi,k} \sim N(0,\sigma_2^2)$

Fig. 2 shows pulsed RFI in single source. Generally, pulsed RFI randomly turns up throughout the length of x_k and has duty cycles 3-5% of the code length. In addition, pulsed RFI has the noise variance changing with time. Pulse train function ($g(t)$) is assumed to be periodic gating function with pulse repetition rate of $1/\mathrm{T}$ and duty cycle τ/T and is used for generating pulsed RFI.

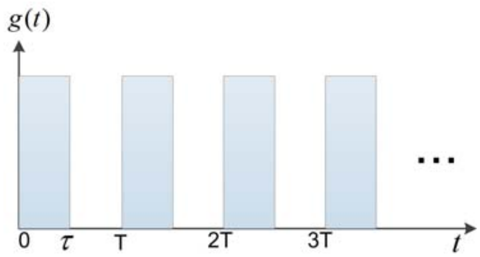

Fig. 2 Gating function for pulsed RFI model.

Fig. 3 demonstrates the influence of pulsed RFI on LDPC decoder performance. It is found that whatever SNR is, BER under pulsed RFI environment is always seriously.

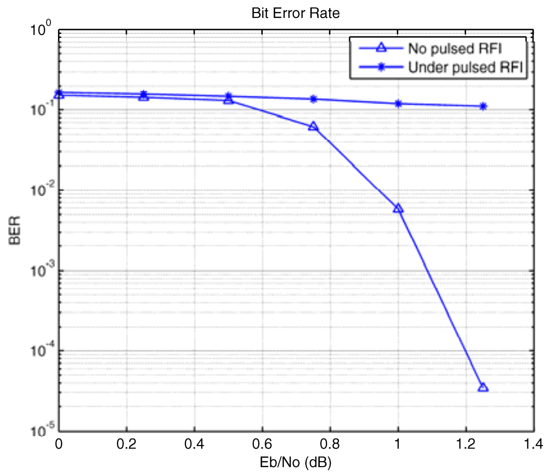

Fig. 3 Pulsed RFI impact on LDPC decoder performance simulated with the AR4JA (8192, 4096) code.

B. Symmetric Alpha-Stable noise Model

The $S\alpha S$ channel describes by a transmitted sequence x_k is interrupted by $S\alpha S$ noise which depict in Fig. 4 and (2), resulting in y_k sequence.

$$y_k = x_k + n_{S\alpha S,k} \qquad (2)$$

When we mention to $S\alpha S$ channel, one properties of $S\alpha S$ distribution [14] [15] is that there is no closed form expression of the PDF (except in case of Cauchy and Gaussian distribution), the most convenient way to explain them is using their characteristic function, which define as:

$$\varphi(\omega) = \exp(j\delta\omega - \gamma |\omega|^{\alpha}) \qquad (3)$$

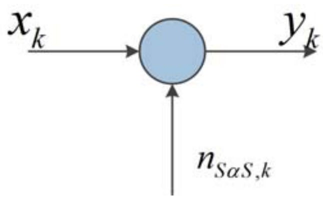

Fig. 4 System model for $S\alpha S$ experiment.

Where

- α is the characteristic exponent. It is vary between $0<\alpha\le 2$, and it describes the tail of the distribution. A small positive value of α points out the severe impulsiveness, resulting in heavy-tail of the distribution. The distribution is close to Gaussian behavior, if α near to 2. A value of α = 1 means Cauchy distribution.
- γ is the dispersion parameter or scale ($\gamma>0$). It indicates the spread of the density around the location parameter. It is similar to the variance of Gaussian density. In Gaussian density case, this value is equal to half of the variance of Gaussian density.
- δ is the location parameter ($-\infty<\delta<\infty$). It corresponds to the mean for $1<\alpha\le 2$, and corresponds to the median for $0<\alpha\le 1$.

Fig. 5 illustrates the $S\alpha S$ sequence at difference α values. They show different impulsiveness with $\alpha=1$, $\alpha=1.2$, $\alpha=1.7$, and $\alpha=2$, respectively. An amplitude of impulse for small α cases show very high interrupted level whereas the value of α which is near to 2 illustrates lower amplitude of impulse which no large of interrupting.

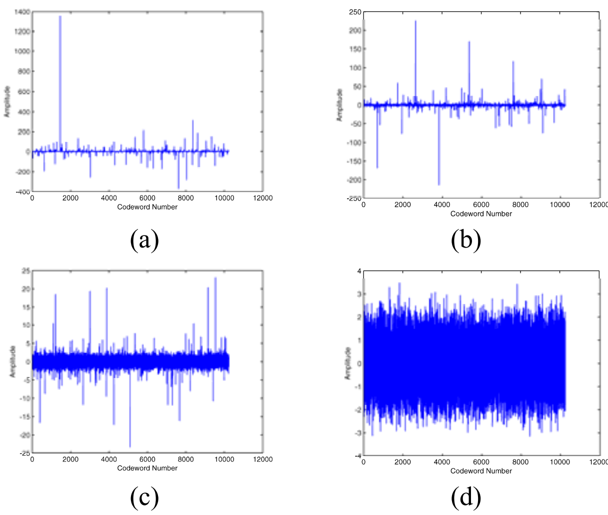

Fig. 5 The received signal interrupting by $S\alpha S$ Noise with (a) $\alpha=1$; (b) $\alpha=1.2$; (c) $\alpha=1.7$;(d) $\alpha=2$.

When we mention to the alpha-stable distribution, it is a four-parameter family of distributions and is generally denoted by $S(\alpha,\beta,\gamma,\delta)$. In the $S\alpha S$ distribution, β value which explains the skewness of the distribution is equal to 0. We can conclude that the family of alpha-stable distribution is a rich class, and includes the following distributions:

- The Gaussian distribution $N(\mu,\sigma^2)$ is provided by $S(2,0,\sigma/\sqrt{2},\mu)$.
- The Cauchy distribution is denoted by $S(1,0,\gamma,\delta)$.

The PDF of Gaussian and Cauchy distribution are given by:

$$f_{gaussian}(x)=\frac{1}{\sqrt{4\pi\gamma}}\exp\left\{-\frac{(x-\delta)^2}{4\gamma}\right\} \tag{4}$$

$$f_{Cauchy}(x)=\frac{\gamma}{\pi\left[\gamma^2+(x-\delta)^2\right]} \tag{5}$$

Even though when $0<\alpha<2,\alpha\neq 1,2$, no closed- form expressions exist for the PDF, but we can be computed the PDF, $f(x)$ by taking the inverse Fourier transform of the characteristic function, resulting in:

$$f_{S\alpha S}(x)=\frac{1}{2\pi}\int_{-\infty}^{\infty}\exp(-\gamma|\omega|^{\alpha})e^{j\omega(x-\delta)}d\omega \tag{6}$$

Fig.6 shows the impact of SαS noise on LDPC decoder performance when LLR calculate from normal distribution. Transmitted signal is more interrupted when the value of α go to small value.

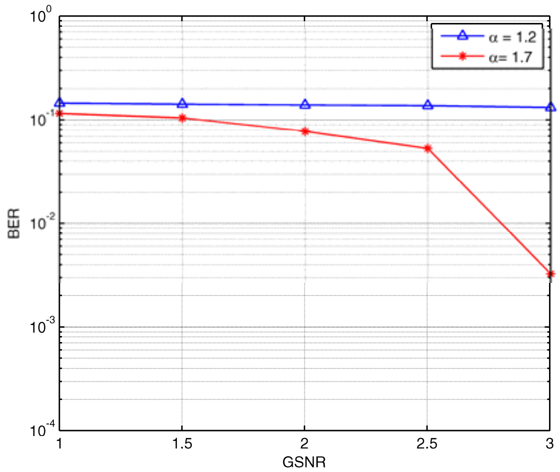

Fig. 6 $S\alpha S$ noise impact on LDPC decoder performance simulated with the AR4JA (8192, 4096) code.

III. CHANNEL PARAMETERS ESTIMATION

A. Pulsed RFI Model

Various digital communication applications, such as power control, bit error estimation, and turbo decoding, involve the knowledge of the SNR. For optimal performance, SNR estimation must be as accurate as possible. Several techniques have been proposed for SNR estimation. M_2 [4] is popular method in several applications because it provides a simple processing. SNR is capable of computing from the ratio of the signal mean squared (A) and the noise variance (σ^2) which both of them are expressed as:

$$A=\frac{1}{N}\sum_{k=1}^{N}|y_k| \tag{7}$$

$$\sigma^2 = \frac{1}{N}\sum_{k=1}^{N}(y_k - A)^2 \tag{8}$$

B. Symmetric Alpha-Stable noise Model

As mentioned above, a difficult task arises in using of $S\alpha S$ model because there is no the closed form of PDF. They do not exist the second-order moment, or the variance. Therefore, another estimated method for noise power is necessary. For instance, in this experiment we use GSNR [15]:

$$GSNR = \frac{1}{2C_g}\left(\frac{A}{S_0}\right)^2 \tag{9}$$

Where
A is the signal amplitude.
S_0 symbolic to the geometric noise power.

$$S_0 = \frac{(C_g\gamma)^{1/\alpha}}{C_g} \tag{10}$$

$C_g = 1.78$ is the experiential of the Euler constant. The constant $2C_g$ in (9) ensures for the SNR in case of $\alpha = 2$ corresponding to the Gaussian distribution with variance $\sigma^2 = 2\gamma$. In digital communication system, we are interested in characteristic of BER in term of E_b / N_0 (the ratio of signal energy per bit to the noise spectral density). This means that the parity bits do not represent transmitted information and their energy must be removed over information bits. Therefore, we can define E_b / N_0 for $S\alpha S$ channel as:

$$\frac{E_b}{N_0} = \frac{GSNR}{2rm} \tag{11}$$

Where r is code rate and m is the number of bits carried per M-array symbol ($m = 1$ for binary code).

IV. JOINT ITERATIVE CHANNEL ESTIMATION AND DECODING

A. Pulsed RFI Model

The system model which mentioned in the section II points out that both AWGN and pulsed RFI are capable of explaining as normal distribution. Based on the PDF of the distribution, we are able to form the initial LLR of simulated channel. If the assumed noise PDF has distribution as:

$$f(n) = \frac{1-\varepsilon}{\sqrt{2\pi\sigma_1^2}}e^{-\frac{n^2}{2\sigma_1^2}} + \frac{\varepsilon}{\sqrt{2\pi\sigma_2^2}}e^{-\frac{n^2}{2\sigma_2^2}} \tag{12}$$

Where $\sigma_2^2 > \sigma_1^2$

Contaminated Gaussian Log-Likelihood-Ratio (CGLLR) [1] has been constructed. CGLLR provides initial information more accurate than implementing Gaussian LLR. CGLLR is re-written as the follow:

$$CGLLR(y_k) = \log\left(\frac{\frac{1-\varepsilon}{\sqrt{2\pi\sigma_1^2}}e^{-\frac{(y_k-1)^2}{2\sigma_1^2}} + \frac{\varepsilon}{\sqrt{2\pi\sigma_2^2}}e^{-\frac{(y_k-1)^2}{2\sigma_2^2}}}{\frac{1-\varepsilon}{\sqrt{2\pi\sigma_1^2}}e^{-\frac{(y_k+1)^2}{2\sigma_1^2}} + \frac{\varepsilon}{\sqrt{2\pi\sigma_2^2}}e^{-\frac{(y_k+1)^2}{2\sigma_2^2}}}\right) \tag{13}$$

Where

- $\sigma_1^2 = \sigma^2 + \Delta\sigma$ is the noise variance of Gaussian part of the PDF.
- σ_2^2 is the noise variance of the contaminating heavy-tailed PDF (pulsed RFI noise). Chosen ranges are $1.0 < \sigma_2^2 \le 4.0$.
- ε is the percentage of sample from the heavy-tailed PDF. It can range from $0 < \varepsilon \le 0.5$ with the value greater than 0.3 occurring with very low probability.

To implement CGLLR, two parameters must be computed. Firstly, the percentage of sample from the heavy-tailed PDF (ε) or pulsed RFI part is calculated by higher order moments. Fig.7 illustrates detection of pulsed RFI technique by M_2 method. Let N is code length and G is number of group.

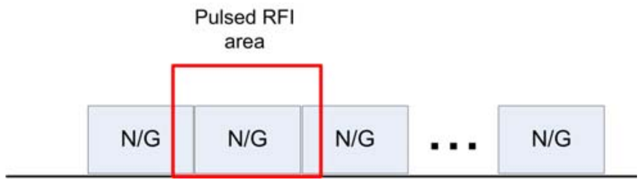

Fig. 7 Detection pulsed RFI by the second order moment.

N is divided into G groups; thus, each group is composed of N/G samples. Because M_2 value of pulsed RFI area is higher than any other region, we take the advantage from this point to capture the area that pulsed RFI appearing. M_2 of every group is calculated and the group which has the largest M_2 value is capable of considering to be pulsed RFI region. Therefore, number of sample of pulsed RFI is approximately equal to N/G. To ensure that all pulsed RFI samples are detected, a few samples both left and right side of that group are included. In other words the pulsed RFI region is approximately equal to the area of group which it has the highest M_2 value including its side-band. Further precision may divide N into several groups and each group contains smaller samples. Applying the same technique; find M_2 of every group, consequently, pulsed RFI area may contain more than one group which have the highest M_2 value respectively. In practice, M_2 are estimated by their respective time averages for both real and complex channels as:

$$M_2 \approx \frac{1}{N}\sum_{k=1}^{N}|y_k|^2 \tag{14}$$

The second factor that needs to compute for implementing CGLLR is the noise variance of Gaussian part of the PDF ($\sigma_1^2 = \sigma^2 + \Delta\sigma$) because the channel status information is unknown. In various applications, M_2 is used because it is the simple way to obtain σ_1^2 and does not take time to calculate. However, the channel in this work is not same as other works since pulsed RFI also appears in the channel. In this environment, M_2 is unable to provide very accurate value. The proposed method to overcome the pulsed RFI can dealing with this problem [1]. The proposed approach is constructed by the principle of joint iterative channel estimation and decoding. It bases on the fundamental that each iterative time of decoding process, the noise variance (σ_1^2) is re-computed by M_2 method to modified information and obtain more accurate noise variance. Schematic of the proposed technique implementing under pulsed RFI shows as Fig. 8 and Algorithm 1.

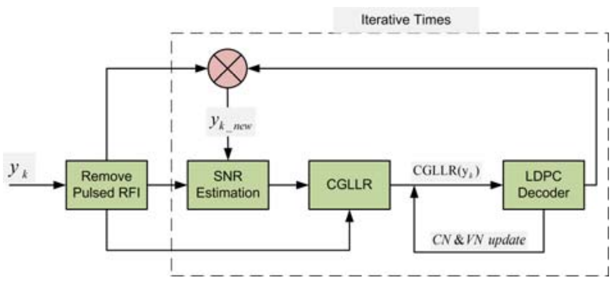

Fig. 8 Schematic block diagram of the proposed technique.

The received signal (y_k) is erased pulsed RFI area which detecting by M_2 method. The received signal after deleting pulsed RFI region denote as y_{k_noRFI}. Consequent step is the noise variance (σ_1^2) evaluation. On the one hand, if it is the first time of LDPC decoder iterations, y_{k_noRFI} will be used to perform the noise variance calculation. On the other hand, the noise variance is computed from y_{k_new} for remaining iteration. $CGLLR(y_k)$ is determined based on the estimated noise variance. We carry out the LDPC decoder only once iterative time. The results consist of hard-decision output, values of check node (CN) , and variable node (VN) updating. Hard-decision outputs are feed back to multiply with the pre- eliminated pulsed RFI signal (y_k) resulting in the new received signal (y_{k_new}). In order to keep LDPC decoder processes, values of check nodes (CN) and variable nodes (VN) updating are also feedback so as to use in the initial state of the next LDPC iteration. In the check stopping criteria step, if the number of iterative time is not equal to the maximum LDPC iterations or decoding error is not equal to zero, go back to step 2.) and increases the number of iterations by one until reach the stopping criterion.

B. *Symmetric Alpha-Stable noise Model*

Joint iteration technique between $S\alpha S$ channel parameter estimation and LDPC decoder allows LLR value updating on fly in every LDPC iterative times, which lead to acquire a high decoder performance. Additionally, because it is not so much mathematics calculation steps in estimated parameter values, the whole processes of the proposed method could has an enhanced results while no computation loss. This section consists of two parts. Firstly, LLR computation is mentioned based on the LM method. Secondly, the joining process of LDPC decoder and estimation of $S\alpha S$ channel parameters is described.

Algorithm 1 Joint iterative between channel estimation and decoding under pulsed RFI condition.

1: **procedure** pulsedRFI (y_k)
2: **Eliminate pulsed RFI**
3: **if** decoding iterative =1 **then**
$\sigma_1^2 = \mathrm{var}(y_{k\_noRFI})$
4: **else**
$\sigma_1^2 = \mathrm{var}(y_{k\_new})$
5: **end if**
6: **LLR computation**
7: **Once iterative time LDPC decoder**
8: **Output feedback**
$y_{k\_new} = y_k \cdot hard - decision$
Check Node (CN) update vale
Variable Node (VN) update value
9: **Check stopping criteria**
10: **if** iterative time = max iteration || Error=0 **then**
Go to 13
11: **else**
Back to 3
12: **end if**
13: **Output codeword**
14: **end procedure**

In (3) there are two key parameters for calculate the PDF of $S\alpha S$ distribution: characteristic exponent (α) and dispersion parameter (γ). The location parameter (δ) can be discarded because it is a position and no influence on the simulation analyses. In this experiment, those two parameters are computed from LM method [13]. Let $X \sim S(\alpha, 0, \gamma, 0)$ is sequence of $S\alpha S$ signal then:

$$L_1 = \mathrm{E}\left[\log|X|\right] = \psi_0(1-\frac{1}{\alpha}) + \frac{1}{\alpha}\log|\frac{\gamma}{\cos\theta}| \tag{15}$$

$$L_2 = \mathrm{E}\left[(\log|X| - \mathrm{E}\left[\log|X|\right])^2\right] = \psi_1(\frac{1}{2}+\frac{1}{\alpha^2}) - \frac{\theta^2}{\alpha^2} \tag{16}$$

Where
L_1 is logarithmic moment at order 1
L_2 is logarithmic moment at order 2

$$\psi_0 = -0.57721566..., \qquad \psi_1 = \frac{\pi^2}{6} \tag{17}$$

We can solve for α by setting $\theta = 0$ and then calculate for γ value. The LLR calculation for $S\alpha S$ channel is related to the PDF of the $S\alpha S$ distribution which express as:

$$LLR(y_k \mid x_k) = \ln\left(\frac{P(y_k \mid x_k = +1)}{P(y_k \mid x_k = -1)}\right) \tag{18}$$

In $S\alpha S$ channel, the joint iterative between channel estimation and LDPC decoder perform as Fig. 9 and Algorithm 2. The processes begin with an initial LLR value.

For the first iteration of LDPC, LLR is calculated by setting α equal to 2 (Gaussian LLR). According to (15) - (16), we can obtain the γ value. From these two parameter values, we can compute LLR. The initial LLR value are input to LDPC decoder resulting in three output parameters.

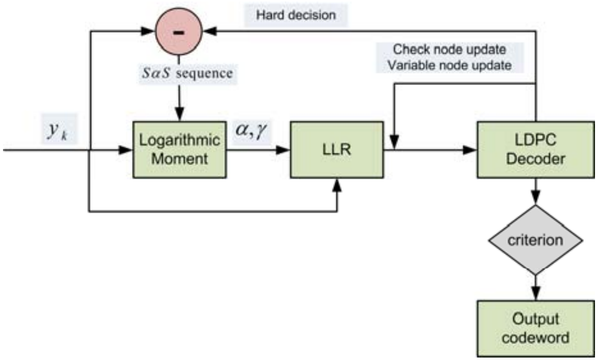

Fig. 9 Schematic block diagram of the proposed technique for joint iterative between channel estimation and decoding under $S\alpha S$ channel.

The first two outputs are variable node and check node values of LDPC, which will be used for the next iteration of the decoder. The third output is hard-decision, it will be feedback to subtract from received signal (y_k) to estimate α and γ values by LM method before calculating the LLR value for the next loop of LDPC decoder. The output from subtracting hard-decision with received signal (y_k) is $S\alpha S$ noise sequence, which can be used to estimate α and γ values as (15)-(16). Because two important parameters are estimated in every iterative time, the LLR value can update to more accurate value which has influence on the decoder performance. The loops are continuing until reach the criterion and get output codeword in the final step.

V. Simulation Results and Discussion

The LDPC code studied here is the AR4JA (8192, 4096) code recommended by the Consultative Committee for Space Data Systems (CCSDS) and it has been chosen for some deep space applications. In this section, BER of the joint iterative channel estimation and decoding or the proposed approaches are presented for both under pulsed RFI and $S\alpha S$ channel.

A. Pulsed RFI Model

In the simulation, we set the noise variance of the contaminating heavy-tailed (pulsed RFI) PDF (σ_2^2) by 4.0 and set the random walk signal of AWGN ($\Delta\sigma$) by 10% of the noise variance of AWGN part (σ_1^2). The proposed method was compared the result with the conventional method which performs noise variance of AWGN part (σ_1^2) estimation by M_2 SNR estimation based on hard-decisions of the previous codeword which they are feedback to multiply with that codeword to eliminate modulated information. Fig. 10 depicts that the proposed technique has only about 0.2 dB at BER = 10^{-3} different from the SNR know experiment. The efficiency of the proposed approach is close to the performance of ideal technique because the noise variance (σ_1^2) is re-calculated in every time of iterative LDPC decoder.

Algorithm 2 Joint iterative between channel estimation and decoding under $S\alpha S$ noise.

1: **procedure** SaSnoise (y_k)
2: **LLR computation**(α, γ, y_k)
3: **if** decoding iterative =1 **then**
$\alpha = 2$
$\gamma = LM(\alpha, S\alpha S \text{ sequence})$
4: **else**
$\alpha = LM(S\alpha S \text{ sequence})$
$\gamma = LM(\alpha, S\alpha S \text{ sequence})$
5: **end if**
6: **Once iterative time LDPC decoder**
7: **Output feedback**
$S\alpha S \text{ sequence} = y_k - hard\_decision$
Check Node (CN) update vale
Variable Node (VN) update value
8: **Check stopping criteria**
9: **if** iterative time = max iteration || Error=0 **then**
Go to 12
10: **else**
Back to 2
11: **end if**
12: **Output codeword**
13: **end procedure**

Therefore, the noise variance (σ_1^2) has always been changed and obtains more accurate value in each time of LDPC iteration.

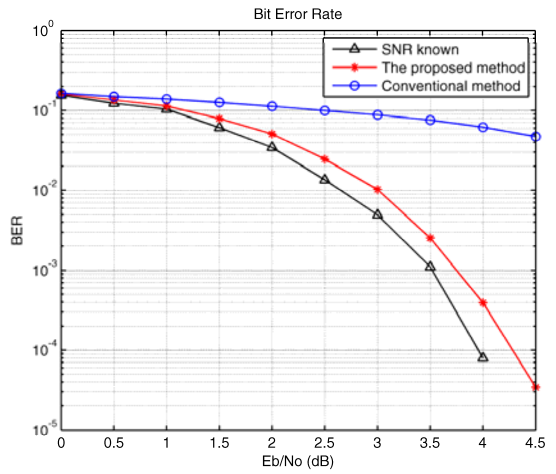

Fig. 10 Bit-Error-Rate performance in the jiggling AWGN channel simulated with the AR4JA (8192, 4096) code.

Table I. makes a comparison of the single block information for SNR estimation of those two techniques. Two kind of operation are taken into our consideration; addition, and multiplication.

Table I
Comparison of Single Block Information for SNR Estimation of the Two Method

	Conventional method	Proposed method
Addition	$M-1$	$(M-1)k$
Multiplication	$M+1$	$(M+1)k$

Let length of single block information which pulsed RFI is removed is M . k represents the number of LDPC iteration.

In the single block information, the conventional method requires the smallest operational numbers. The proposed approach has a little more computational complexity than the conventional technique depending on the number of LDPC iteration (k).

B. Symmetric Alpha-Stable noise Model

There are two test cases of α value in this simulation: $\alpha = 1.2$ and $\alpha = 1.7$ which is no closed form PDF for these α values. The results demonstrate the proposed method for the joint iterative method between $S\alpha S$ channel estimation and LDPC decoder in term of BER as depict in Fig. 11. BER curve of $\alpha = 1.2$ is more corrupt than BER curve of $\alpha = 1.7$. This is because $\alpha = 1.2$ generate stronger impulsiveness to the channel, thus, it needs high level of GSNR value to obtain a satisfied performance. In the curves, GSNR equals to 2.85 could give BER about 10^{-4} in case of $\alpha = 1.2$ and the GSNR equals to 2 could get BER about 10^{-4} when $\alpha = 1.7$, because $\alpha = 1.7$ is near to be Gaussian noise and has smaller impulsiveness on channel than $\alpha = 1.2$.

From Fig 6 and 11, it is clear that the proposed method has ability to mitigate the effect of $S\alpha S$ noise better than the conventional method which widely uses normal distribution to calculate an initial LLR for LDPC decoder.

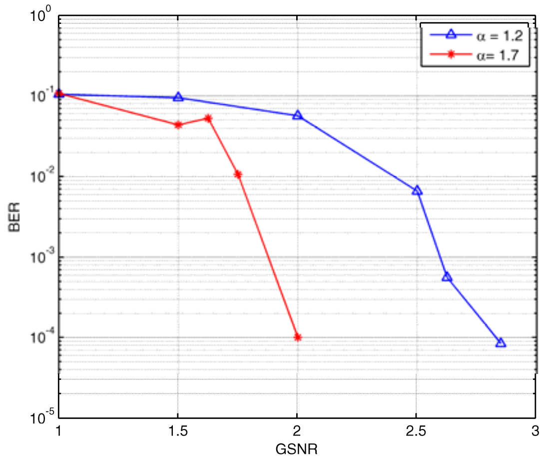

Fig. 11 Bit-Error-Rate performance in the $S\alpha S$ channel when $\alpha = 1.2$ and $\alpha = 1.7$ simulated with the AR4JA (8192, 4096) code.

TABLE II
THE ESTIMATED PARAMETER VALUES OF $S\alpha S$ CHANNEL IN EVERY LDPC ITERATIVE TIME

	Characteristic exponent (α)	Dispersion parameter (γ)
Iteration =1	2.0000	1.2377
Iteration =2	1.7153	0.3491
Iteration =3	1.7304	0.3289
Iteration =4	1.7551	0.3218
Iteration =5	1.7425	0.3205
.	.	.
.	.	.
.	.	.
Iteration =49	1.7099	0.3384
Iteration =50	1.7099	0.3384

Table II demonstrates an updating of two important parameters of SαS channel in the case of GSNR = 2 in fifty iterative times. Both α and γ tend to go to the true value in every iterative time, which established by α = 1.7 and γ = 0.5338. Although, the updating of γ parameter is unable to reach the true value, it is better than implement LDPC decoder with LLR from normal distribution, which both two parameters have the values same as setting in the first iteration time. The updated parameter reflects the enhancement of LLR calculation that mean LLR value can go to the appreciated values which has influence to decoder performance.

VI. CONCLUSION

Joint iterative between channel estimation and decoding have been presented in this work for pulsed RFI and $S\alpha S$ channel. The second order moment method have been implement for estimating the noise variance in pulsed RFI channel and logarithmic moment is used for estimating the important two parameter under $S\alpha S$ channel. In addition, a joint iterative technique also provides a chance to update channel parameters in every iterative time of the decoder which is useful for decoder performance. The results demonstrate that the proposed technique can mitigate the impact of the impulsive noise on LDPC decoder for both pulsed RFI and $S\alpha S$ channel. The proposed method is attractive to apply to various applications which pulsed RFI or $S\alpha S$ is involved with their requirements.

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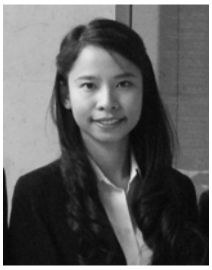

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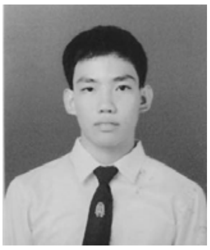

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