Joint Packet Detection and Decoding for Maritime Data Exchange Systems

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***Abstract*—A random access technique based on spatially coupling information transmission modulation is proposed. We consider an application of the method to the maritime VHF data exchange system (VDES) in which a large pool of vessels attempt uplink ship-to-satellite communication in an uncoordinated fashion using short data packets. The proposed receiver performs joint packet detection and data decoding of the stream of interfering (coupled) incoming packets. An efficient receiver architecture is designed using a low-complexity approximate message passing (AMP) algorithm. It is demonstrated that the proposed random access technique can successfully accommodate traffic loads that are considerably higher than loads achievable with competing random access mechanisms.**

I. Introduction

Maritime data exchange systems have received significant research attention in recent years due to the rapidly growing number of vessels and information services required for their operation including positioning, tracking, and navigation. The automatic identification system (AIS) has proven to be a very successful means of tracking ships. It is expected, however, that the capacity of AIS will soon be exceeded. In some ocean areas, such as the Gulf of Mexico, the current AIS load is already above 50% of its capacity [1]. Moreover, new applications such as e-navigation require additional capabilities to exchange data. The new standard of VHF data exchange (VDES) which is currently under development by the International Telecommunications Union (ITU) will include satellite ship tracking, ship-to-shore, and ship-to-ship communication and extend the services provided by AIS [2].

Random access communications is an important mechanism of accessing the communication channel resources in VDES. In satellite uplink communications a large varying user (vessel) population is accessing the channel and coordinated channel access is hard to accomplish due to the large feedback delay. At the same time is it important to allow as many vessels as possible to communicate and share channel resources efficiently. Hence, the present time-division based random access mechanism may not be able to cater to the growing traffic demand. In addition to ship-to-satellite requests for downlink data transfer, tracking, and emergency messages, short message exchange between ships and ship-to-shore can also be efficiently handled with random access. Often, vessels do not need to establish a link but just need to send a single packet.

We consider an approach to random access at the physical layer that is based on techniques for simultaneously receiving several packets originating from distinct transmitters. In a random access setting such receiver needs to accomplish two tasks: detect the start of each incoming packet (timing acquisition) and decode the packets. The latter typically also involves cancellation of mutual interference between the packets. It has been observed already in the 90s for the case of code-division multiple access that the detection task limits the multi-packet reception (MPR) capabilities of a receiver [3] [4]. Later it has been shown [5] [6] that joint detection and decoding can improve the MPR capability. It has also been proved in [7] that for packets modulated using generalized modulation format and for specific packet power distribution the detection results in a loss that amounts to a small fraction of the channel capacity asymptotically.

In this work we focus on a spread spectrum transmission system that uses a constant envelop modulation and a spread preamble as in VDES uplink option [2], suitable for a simple implementation in a vessel transmitter. We propose an algorithm for packet encoding at the transmitter and joint packet detection and decoding at the receiver. Due to the specific encoding format, from a communications theoretic perspective we can represent each message as a bi-partite graph relating the data bits and the transmitted modulated symbols. At the receiver consecutively received interfering packet graphs couple into a chain of graphs. Our proposed joint iterative preamble detection, interference cancellation and decoding algorithm is based on the AMP technique [8] and operates on the received graph chain. Numerical results demonstrate that incorporation of the preamble detection into the iterative decoding loop leads to a relatively minor performance loss compared to the throughput of the system with an a priori timing knowledge assumption. The total throughput follows the capacity of the multiple access channel. We also demonstrate that packet loads as high as 2.95 packets per packet length per spreading factor are achievable and that the presented random access mechanism outperforms the existing random access techniques.

II. System Model

We consider a scenario when a number of users send their data packets to the common receiver. The users send packets on demand at random times and create random traffic at the

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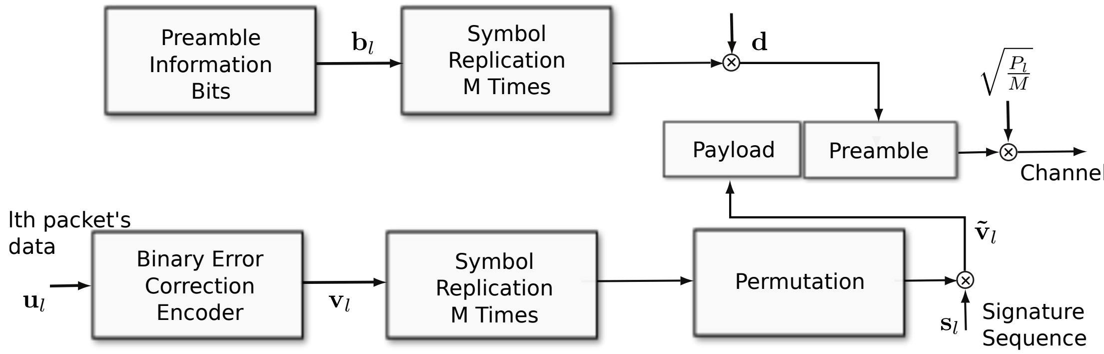

Fig. 1. Encoding and modulation of one data packet.

receiver. We assume that all packets are of the same length, the total number of packets transmitted within a frame is L_{pack}, and the ordered set of packet indices is $\mathcal{L}_{\text{pack}} = (1, 2, \cdots, L_{\text{pack}})$.

Each packet begins with a preamble that will be used for the packet's timing acquisition. The preamble is followed by a payload consisting of the modulated original data to be sent. To produce the payload of packet $l \in \mathcal{L}_{\text{pack}}$ a transmitter starts with encoding the binary data vector $\mathbf{u}_l = u_{1,l}, u_{2,l}, \cdots, u_{K,l}$ with a binary error correction code of rate $R = K/N$. After the encoding each bit of the encoded sequence $\mathbf{v}_l = v_{1,l}, v_{2,l}, \cdots, v_{N,l}$, $v_{j,l} \in \{1, -1\}$, is repeated M times and the resulting sequence is permuted using a permutation described by a binary permutation matrix $\mathbf{P}_l$. Finally, the permuted sequence is multiplied bit-by-bit by the binary pseudo-random signature sequence $\mathbf{s}_l = s_{1,l}, s_{2,l} \cdots, s_{MN,l}$, $s_{j,l} \in \{1, -1\}$, to produce $\tilde{\mathbf{v}}_l = \tilde{v}_{1,l}, \tilde{v}_{2,l}, \cdots, \tilde{v}_{MN,l}$. The resulting binary phase shift keying signal is multiplied by a factor $\sqrt{P_l/M}$, where P_l is the transmit power, and is transmitted over the channel (see Fig. 1) [1].

Hence, each data bit is spread with a spreading factor M and the resulting spread symbols (chips) are permuted. The modulation format follows the coupling information transmission modulation principle [9], [10], [11], [19] and also resembles interleave division multiple access (IDMA) [13], partition-spreading code-division multiple access (PS-CDMA) [15], [16], generalized modulation [12], and superposition modulation [14]. The modulated signal can be represented as a sparse graph connecting variable nodes corresponding to the data bits $v_{j,l}$ to the channel access nodes corresponding to the times at which M replicas of $v_{j,l}$ are transmitted over the channel. By $\mathcal{T}(j, l)$ we denote the set of times $\tau_1, \tau_2, \cdots, \tau_M$ at which the modulated replicas of bit $v_{j,l}$ are transmitted over the channel.

To help the receiver find the starting time of the packet each transmitter starts its packet with a preamble. Besides its purpose for timing acquisition the preamble also carries information. The information may include the seed of the spreading code and permutation used for modulating the packet's payload. Typically this information is encoded by a strong error correction code. Assume that the preamble of the lth packet contains an information bearing bit sequence $\mathbf{b}_l = b_{1,l}, b_{2,l}, \cdots, b_{N_\text{p},l}$ where N_p is the length of the preamble in bits. To modulate the preamble each bit of the sequence $\mathbf{b}_l$ is repeated M times and the resulting sequence is multiplied bit-by-bit by a binary signature sequence $\mathbf{d} = d_1, d_2, \cdots, d_{MN_\text{p}}$ to produce the preamble sequence $\tilde{\mathbf{b}}_l = \tilde{b}_{1,l}, \tilde{b}_{2,l}, \cdots, \tilde{b}_{MN_\text{p}}$. The signature sequence $\mathbf{d}$ is common to all preambles of all packets. As with the payload, the sequence $\tilde{\mathbf{b}}_l$ is finally multiplied by the amplitude factor $\sqrt{P_l/M}$ prior to transmission. The pseudo-random signature sequence $\mathbf{d}$ is constructed from sequences with low autocorrelation [20] to reduce the rates of false alarms.

Thus, one data packet consists of an $N_\text{p}M$-chip preamble followed by an NM-chip payload. The length parameter N_p determines the transmission rate-loss and complexity of the acquisition device, and it is desirable that this preamble length is minimized, especially for the case of transmitting short packets as in VDES.

Combining the payload and preamble data into one vector $\mathbf{x}_l = (\mathbf{b}_l, \mathbf{v}_l)^\text{T}$ we can represent the lth modulated packet, $l \in \mathcal{L}_{\text{pack}}$ as a product

$$\tilde{\mathbf{x}}_l = \mathbf{B}_l \mathbf{x}_l \ .$$

The matrix $\mathbf{B}_l$ is given by

$$\mathbf{B}_l = \sqrt{\frac{P_l}{M}} \operatorname{diag}(\mathbf{d}, \mathbf{s}_l) \begin{pmatrix} \mathbf{I}_{MN_\text{p}} & \mathbf{0} \\ \mathbf{0} & \mathbf{P}_l \end{pmatrix} \mathbf{R}_{N+N_\text{p}} \quad (1)$$

$$= \tilde{\mathbf{B}}_l \mathbf{R}_{N+N_\text{p}} \ , \quad (2)$$

where $\operatorname{diag}(\mathbf{d}, \mathbf{s}_l)$ is an $(N + N_\text{p})M \times (N + N_\text{p})M$ diagonal matrix with sequence $(\mathbf{d}, \mathbf{s}_l)$ on the main diagonal, $\mathbf{I}_{MN_\text{p}}$ is an identity matrix of size MN_p, and $\mathbf{R}_{N+N_\text{p}}$ is an $(N+N_\text{p})M \times (N+N_\text{p})$ matrix which performs M-times replication via right multiplication.

We assume that throughout the frame the users transmit their packets with the overall traffic intensity αM packets/packet length. Hence, at the beginning of the frame the packet traffic increases linearly from an intensity of 0 packets/packet length to αM packets/packet length and stays at that level (see Fig. 2). The graphs representing the packets couple in the channel randomly into a chain of graphs. Hence, we obtain

[1]To simplify the presentation of the material we consider BPSK modulated packets. The format can be extended to QPSK modulation in a straightforward manner (see [7]).

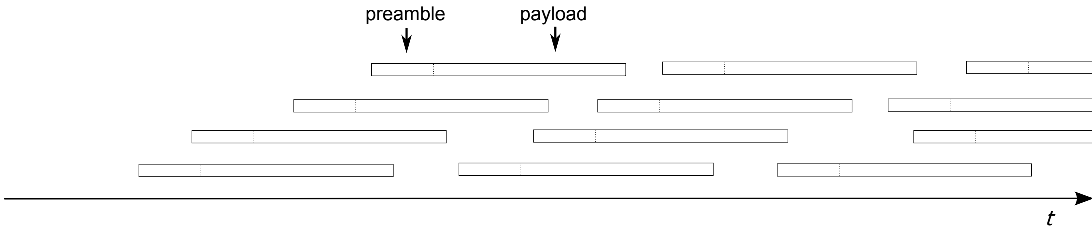

Fig. 2. A stream of coupled packets with random arrivals of approximately 4 packets/packet length.

a spatially coupled system on which we will try to exhibit the threshold saturation phenomenon [17] [18] via an iterative message-passing-like receiver processing. The fact that the inter-packet interference gradually increases at the beginning of the frame will help to start the detection and decoding process which will then continue successfully for the duration of the entire frame.

We consider communication over the additive white Gaussian noise (AWGN) channel, and so, the received signal $\mathbf{y} = (y_1, y_2, y_3, ...)$ is computed as a superposition:

$$y_\tau = \sum_{l=1}^{L(\tau)} \tilde{x}_{\tau,l} + n_\tau \ , \qquad \tau = 0, 1, 2, \cdots$$

where $L(\tau)$ denotes the number of interfering packets at time τ and $n_\tau \sim \mathcal{N}(0, \sigma^2)$ are the iid noise samples.[2] The received signal also allows for representation

$$\mathbf{y} = \mathbf{A}_{\mathcal{L}_{\text{pack}}} \mathbf{x} + \mathbf{n} \ , \tag{3}$$

where $\mathbf{x} = (\mathbf{x}_1^{\mathrm{T}}, \mathbf{x}_2^{\mathrm{T}}, \cdots, \mathbf{x}_{L_{\text{pack}}}^{\mathrm{T}})^{\mathrm{T}}$ is the vector of stacked data and preamble information of all packets, $\mathbf{n}$ is the vector of noise samples and

$$\mathbf{A}_{\mathcal{L}_{\text{pack}}} = \mathbf{C}_{\mathcal{L}_{\text{pack}}} \begin{pmatrix} \mathbf{B}_1 \\ \mathbf{B}_2 \\ \vdots \\ \mathbf{B}_{L_{\text{pack}}} \end{pmatrix} \tag{4}$$

where $\mathbf{C}_{\mathcal{L}_{\text{pack}}}$ is the coupling matrix representing the addition of all the transmitted packets in the channel. The starting times of the packets $1, 2, \cdots, L_{\text{pack}}$ are denoted by $\theta_1, \theta_2, \cdots, \theta_{L_{\text{pack}}}$. The coupling matrix $\mathbf{C}_{\mathcal{L}_{\text{pack}}}$ is depicted in Fig. 3. Generalizing this definition, given a set of indices $\mathcal{L} \subset \mathcal{L}_{\text{pack}}$ by $C_{\mathcal{L}}$ we denote a coupling matrix coupling only packets $l \in \mathcal{L}$, by $\mathbf{B}_{\mathcal{L}}$ we denote the matrix produced by vertical stacking of matrices $\mathbf{B}_l$, $l \in \mathcal{L}$ and by $\mathbf{x}_{\mathcal{L}}$ a vector of corresponding packet data $\mathbf{x}_l$, $l \in \mathcal{L}$, also stacked vertically. Finally, $\mathbf{A}_{\mathcal{L}}$ is computed analogously to (4).

A. Preamble Detection

The first iteration starts with the packet detection performed via correlating the preamble's signature sequence $\mathbf{d}$ with the incoming signal. The preamble detector slides through the incoming sequence $\mathbf{y}$ chip-by-chip and evaluates the correlation detection statistics

$$\eta_\tau = \sum_{j=1}^{N_{\text{p}}} \left| \frac{1}{\sqrt{M}} \sum_{m=1}^{M} y_{\tau+(j-1)M+m-1} d_{(j-1)M+m-1} \right|^2 . \tag{5}$$

Since the preamble contains unknown bits we have the inner summation over the chips spreading each bit and compute the energy using the outer summation in (5). The statistics η_τ is compared to the decision threshold G and in the case $\eta_\tau \geq G$ a decision about the presence of a packet starting at time τ in the received signal is made.

After the initial detection, some packets, mainly those located at the beginning of the frame, are detected, and will be processed by the data estimation and the interference cancellation block (multiple user detection block) discussed in Section II-B. At the end of the estimation and interference cancellation stage the residual vector $\mathbf{z}^{(1)}$ containing the residual interference, undetected packets, and noise will be incoming for the preamble detection at the second iteration. The statistics η_τ will be computed based on $\mathbf{z}^{(1)}$ and since $\mathbf{z}^{(1)}$ has reduced level of interference compared to $\mathbf{y}$ more packets may be detected. The newly detected packets join the estimation and interference cancellation process at the second iteration. The iterations are repeated until all or almost all packets are detected and the interference level is reduced to a level at which the error correction decoders can decode the packets. A block diagram of the receiver is depicted in Fig. 4.

The thresholding parameter G may be determined assuming that the noise and interference experienced by the packets in the multi-user environment can be approximated by a Gaussian random variable. [3] In this case the statistics η_τ follows a Chi-squared distribution with N_{p} degrees of freedom and non-centrality parameter $N_{\text{p}}(P + \sigma_{\text{I}}^2)$ where σ_{I}^2 is the noise and interference power and P is the power of the packet. If the packet is absent, η_τ follows a central Chi-squared distribution with N_{p} degrees of freedom. Supposing N_{p} is large enough, the Central Limit Theorem implies that η_τ may be treated as a Gaussian random variable with expectation and variance

$$\mathbb{E}\eta_\tau = N_{\text{p}}(P + \sigma_{\text{I}}^2), \quad \text{var}(\eta_\tau) = N_{\text{p}}(P + \sigma_{\text{I}}^2)\sigma_{\text{I}}^2$$

for the case of a packet being present at time τ and

$$\mathbb{E}\eta_\tau = N_{\text{p}}\sigma_{\text{I}}^2, \quad \text{var}(\eta_\tau) = N_{\text{p}}\sigma_{\text{I}}^4$$

[2]Without loss of generality we assume that the received packets are chip-synchronized, otherwise we would consider sample-based detection and interference cancellation processing with the sampling time the receiver would use to oversample the incoming signal.

[3]This assumption relies on the Central Limit Theorem as we have a collection of a large number of independent interfering packets. The packets' power distribution needs to satisfy Lindeberg Condition.

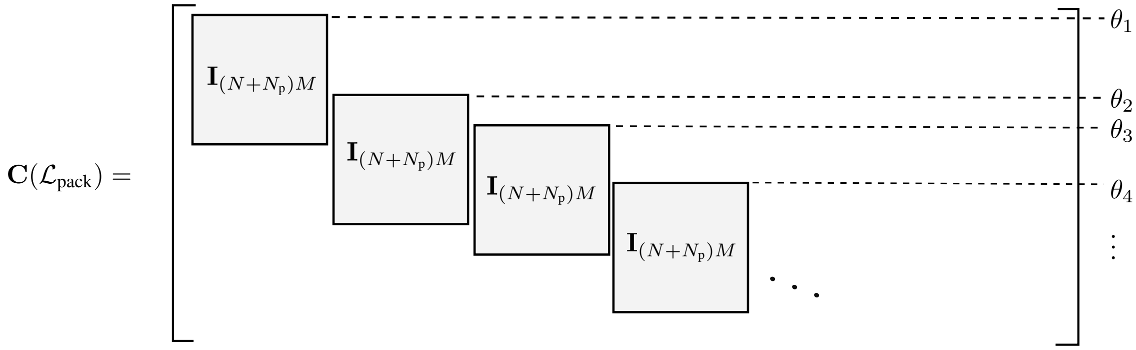

Fig. 3. The coupling matrix $C(\mathcal{L}_{\text{pack}})$ consists of identity matrices surrounded by zeros. Each identity matrix corresponds to an incoming packet and its row offset corresponds to the packet's starting time.

in the case it is absent. With this we may compute the probability of a false alarm as

$$p_{\text{fa}} = Q\left(\frac{G - N_\text{p}\sigma_\text{I}^2}{\sqrt{N_\text{p}\sigma_\text{I}^4}}\right) ,$$

where

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

Fixing the false alarm probability to some value p_0, and denoting $q_0 = Q^{-1}(p_0)$ allows us to solve for the threshold

$$G = \sigma_\text{I}^2\sqrt{N_\text{p}}(\sqrt{N_\text{p}} + q_0) .$$

At iteration i the variance σ_I^2 is estimated directly from the residual vector $\mathbf{z}^{(i-1)}$ which is processed for detection. To compute the decision threshold G for time τ the variance σ_I^2 needs to be computed for the part of the vector $\mathbf{z}^{(i-1)}$ around the time index τ. The length of this window may be chosen to be $3-5\%$ of the packet length since σ_I^2 varies significantly among different sections of the vector $\mathbf{z}^{(i-1)}$. Note also that in practice G has to be adjusted according to the observed false alarm probability, see Section IV. Let us denote the set of indices of packets detected by iteration i by $\mathcal{L}_{\text{pack}}(i)$.

B. Payload Decoding

One iteration of the payload decoding starts with a data estimation step performed for each bit of the detected packets $l \in \mathcal{L}_{\text{pack}}(1)$. We start with considering estimation of bit $v_{j,l}$ of packet l at the first iteration. At the beginning we form a vector $\mathbf{y}_{j,l}$ that consists of M received chips $\{y_\tau\}_{\tau\in\mathcal{T}(j,l)}$, $\mathcal{T}(j,l) = \{\tau_1, \tau_2, \cdots, \tau_M\}$ that contain $v_{j,l}$

$$\mathbf{y}_{j,l} = \mathbf{h}v_{j,l} + \zeta_{j,l}$$

where $\mathbf{h} = \sqrt{P_l/M}(s_{l,\tau_1-\theta_l+1}, s_{l,\tau_2-\theta_l+1}, ..., s_{l,\tau_M-\theta_l+1})^\text{T}$ and $\zeta_{j,l}$ is the vector of noise and interference from the other packets. Computing the minimum mean-squared error (MMSE) filtering of $\mathbf{y}_{j,l}$ to form an SNR-optimal linear of estimate of $v_{j,l}$ we find

$$w_{j,l} = (\mathbf{I} + \mathbf{h}^*\mathbf{F}_{j,l}^{-1}\mathbf{h})^{-1}\mathbf{h}^*\mathbf{F}_{j,l}^{-1}\mathbf{y}_{j,l}$$

where $\mathbf{F}_{j,l}$ denotes the covariance matrix of $\zeta_{j,l}$. One may compute the SNR of the signal $w_{j,l}$ as

$$\gamma_{j,l} = \mathbf{h}^*\mathbf{F}_{j,l}^{-1}\mathbf{h}.$$

In other words $w_{j,l}$ is a result of maximum ratio combining (MRC) of M received replicas of $v_{j,l}$. Since each bit $v_{j,l}$ can be 1 or -1, and takes each of the values with probability $1/2$, one can form a conditional expectation estimate $\hat{v}_{j,l}$ of $v_{j,l}$ as

$$\hat{v}_{j,l}^{(1)} = \mathbb{E}(v_{j,l}|w_{j,l}) = \tanh(w_{j,l}\gamma_{j,l}).$$

The estimates $\hat{v}_{j,l}^{(1)}$ of the data bits may then be used instead of the true bits $v_{j,l}$ unknown to the receiver to remodulate the packets. Since our estimation and interference cancellation process is performing message passing on a graph for each bit $v_{j,l}$ we compute M estimates $v_{j,l,m}^{(1)}$, $m = 1, 2, \cdots, M$. An estimate $\hat{v}_{j,l,m}^{(1)}$ is obtained exactly as $\hat{v}_{j,l}^{(1)}$ discussed above but without involving the observation y_{τ_m} in the estimation process (i.e. we start with vector $\mathbf{y}_{j,l,m}$ where the y_{τ_m} component is not present). We then produce a vector $\hat{\mathbf{v}}_l^{(1)}$ by stacking all the estimates $v_{j,l,m}^{(1)}$, re-modulate the packet and produce

$$\hat{\mathbf{x}}_l^{(1)} = \sqrt{\frac{P_l}{M}}\,\text{diag}(\mathbf{d}, \mathbf{s}_l)\begin{pmatrix} \mathbf{I}_{MN_\text{p}} & \mathbf{0} \\ \mathbf{0} & \mathbf{P}_l \end{pmatrix}\left[\mathbf{b}_l\mathbf{R}_{N_\text{p}}^\text{T}\ \hat{\mathbf{v}}_l^{(1)}\right]^\text{T} .$$

At the next step we perform full interference cancellation to compute the residual vector of noise and interference

$$\mathbf{z}^{(1)} = \mathbf{y} - \mathbf{A}_{\mathcal{L}_{\text{pack}}(1)}\hat{\mathbf{x}}_{\mathcal{L}_{\text{pack}}(1)}^{(1)}$$

which is forwarded to the preamble detector at the beginning of the second iteration. We also compute vectors

$$\mathbf{y}_l^{(1)} = \mathbf{y} - \mathbf{A}_{\mathcal{L}_{\text{pack}}(1)\setminus l}\hat{\mathbf{x}}_{\mathcal{L}_{\text{pack}}(1)\setminus l}^{(1)} , \qquad l \in \mathcal{L}_{\text{pack}}(1)$$

that will be used for estimation of data of packets $l \in \mathcal{L}_{\text{pack}}(1)$ at the second iteration. At the beginning of the second iteration the packet detector produces an updated set of detected packet indices $\mathcal{L}_{\text{pack}}(2)$. For the packets newly detected at the second iteration, i.e. $l \in \mathcal{L}_{\text{pack}}(2) \setminus \mathcal{L}_{\text{pack}}(1)$ the vector $\mathbf{z}^{(1)}$ will be used for data estimation. The data estimation stage of the second iteration computes $\hat{\mathbf{x}}_l^{(2)}$ which is then passed to the interference canceller. The entire process repeats for I

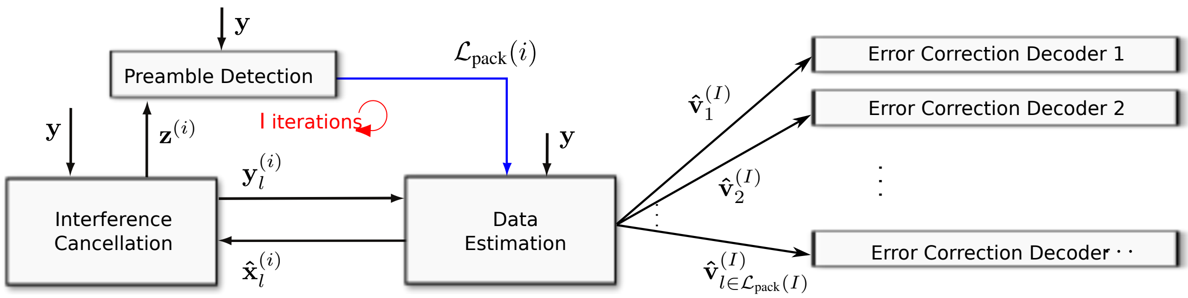

Fig. 4. Diagram of the iterative receiver that includes joint preamble detection and payload decoding followed by individual error correction decoding.

iterations until all or almost all packets are detected and mutual inter-packet interference is reduced to (approximately) the noise power level. After that the packets are individually decoded by the error correction decoders. A detailed analysis of convergence for such a system is given in [10] where it is proved that as preamble length increases the system's performance approaches the capacity of the channel.

III. AMP Algorithm for Joint Decoding and Detection

We notice that at each iteration we need to compute and store M copies of each estimated packet bit and up to $|\mathcal{L}_{\text{pack}}(i)|$ different cancelled vectors. To reduce this computation burden we will use the AMP framework [8] [21]. At each iteration we compute only one estimate for each data bit and each vector, and use correction factors to carefully approximate the original detection and decoding process.

The description of our proposed AMP algorithm based on series expansion and approximation of the passed messages is given in Fig. 5. The AMP version has a simple description and saves orders of magnitude in complexity and storage for algorithm implementation. Instead of storing the entire sparse matrix $\mathbf{A}_i$ matrix multiplications may be done via explicit modulation as in Fig. 1. In practical conditions with no chip synchronicity between the packets steps (6) and (7) can by done on sample basis.

Algorithm:

$$\text{Initialization:} \quad \mathbf{z}^{(0)} = \mathbf{y}, \ \hat{\mathbf{x}}^{(1)} = \mathbf{0}$$

For $i = 1, 2, \cdots, I$

$$\mathcal{L}_{\text{pack}}(i) = \text{CorrelationDetect}(\mathbf{z}^{(i-1)}) \tag{6}$$

$$\mathbf{A}_i = \mathbf{A}_{\mathcal{L}_{\text{pack}}(i)}$$

$$\mathbf{z}^{(i)} = \mathbf{y} - \mathbf{A}_i \hat{\mathbf{x}}^{(i)} + (\mathbf{A}_i * \mathbf{A}_i)\, \mathbf{v}_i * \frac{\mathbf{z}^{(i-1)}}{\text{var}\left(\mathbf{z}^{(i-1)}\right)} \tag{7}$$

$$\mathbf{w}_{i+1} = \mathbf{A}_i^{\text{T}} \mathbf{z}^{(i)} + \hat{\mathbf{x}}^{(i)} * \left(\left(\mathbf{A}_i^{\text{T}} * \mathbf{A}_i^{\text{T}}\right) \frac{1}{\text{var}\left(\mathbf{z}^{(i)}\right)} \right)$$

$$\hat{\mathbf{x}}^{(i+1)} = \tanh(\mathbf{w}_{i+1})$$

$$\mathbf{v}_{i+1} = \text{sech}(\mathbf{w}_{i+1})$$

$*$ denotes component-wise multiplication.

Fig. 5. AMP based joint decoding and detection algorithm. The above iterative process is followed by the individual error correction decoding of the packets.

IV. Numerical Results

Fig. 6 demonstrates the simulated performance of the proposed random access system with a joint detector and decoder. We consider a spreading factor $M = 64$, $N_{\text{p}} = 15$ bit preamble, $N = 390$ bit payload, and set the average rate of the new false alarms per iteration to be $< 10^{-5}$. Equal power distribution of the packets' powers is considered [4] and the transmission is over the AWGN channel with signal-to-noise ratio (SNR) equal to $P\alpha/\sigma^2$. Binary error-correction codes optimal for the SNR observed at the output of the joint decoder after the final iteration are assumed [5]. The rate R of the error-correction codes approaches 1 at high SNRs. The normalized traffic intensity is in the range $\alpha \in [0.84, 2.95]$ and the achievable throughput is computed for the maximum traffic intensity achievable for a given SNR as αR.

The achievable throughput is given in blue in Fig. 6, the theoretic throughput with perfect timing knowledge by the magenta stars curve, its asymptotic limit (as in [9]) by the magenta curve, the channel capacity by the red curve, and the throughput for separate detection and decoding (as in [10]) by the black curve. We notice that packet detection results in a moderate loss (about 2.2dB or smaller) compared to the results with perfect timing knowledge assumption.

The threshold G given by formula (6) needs to be adjusted to practical conditions to give specific false alarm probability. For this purpose we have multiplied G computed for the entire vector $\mathbf{z}^{(i)}$ by a single scaling factor ranging between 1.4 and 2.6 depending on the load. In addition, the preambles of the detected packets are removed upon detection to reduce the inter-packet interference. The rate of the false alarms that will inevitably occur during the detection process is kept low (suggested rate for VDES is 10^{-4} [2]). We do not consider explicit processing of the false alarms in this work, however, the false alarms may be detected by the decoding the preamble codes or noticing no improvement of the iterative

[4]Unequal power distribution typically results in better iterative multi-user detector performances since power differences facilitate the convergence of the interference cancellation process. However, a modified sliding correlation detector needs to be considered in this case. For each time position detection at a number of threshold levels needs to be attempted before sliding forward.

[5]We have not considered the impact of finite-length capacity.

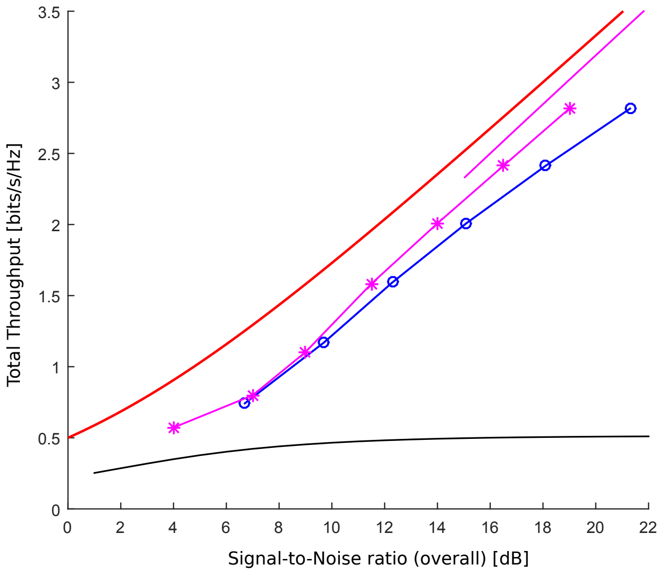

Fig. 6. The throughput achievable by the proposed random access system in bit/s/Hz as a function of the total system's SNR.

data estimation of a non-existing packet.

Comparing the results with other random access techniques we notice that the achieved traffic load of 2.95 of packets per packet length per spreading factor (bandwidth) is much higher than 1 packets per slot or (packet length for unslotted systems) which is a typical maximum for classic random access systems such as ALOHA [22], carrier-sense multiple access (CSMA), self-organized TDMA used in AIS, contention resolution diversity slotted ALOHA (CRDSA) [23], irregular resolution slotted ALOHA (IRSA) [24], coupled random access [25], and even recently proposed systems such as coded slotted ALOHA and enhanced contention resolution ALOHA (ECRA) [26] that can achieve loads up to 2 packets/packet length.

V. Conclusions

We have proposed a random access technique for maritime data exchange systems. At the receiver the detection and decoding of a stream of interfering randomly incoming packets is performed via an iterative process which includes correlation detection, data estimation, and interference cancellation. The technique exploits the principle of spatial graph coupling in order to reliably operate in high packet traffic conditions. A simplified receiver of lower complexity, based on approximate message passing technique is proposed. Simulation results demonstrate that the resulting performance suffers a relatively small loss compared to the system with perfect timing knowledge assumption and the achievable traffic loads are higher than the peak loads sustained by competing random access techniques.

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