International Journal For Technological Research In Engineering
Volume 2, Issue 6, February-2015 *ISSN (Online): 2347 - 4718*

COMPRESSIVE SENSING FOR SPREAD SPECTRUM RECEIVERS: A NEW APPROACH

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ABSTRACT: Power efficiency and Production cost are two important designing parameters for the wireless communication devices. In the Compressive sensing, the receiver is enabled in such devices to sample below the Shannon-Nyquist sampling rate, which may lead to a decrease in the two design parameters. Receiver samples the signal below the Shannon-Nyquist sampling rate when it provided with the compressive sensing. But as a drawback the above two parameters degrades. In this paper we have studied the use of Compressive Sensing (CS) in a general Code Division Multiple Access (CDMA) receiver. We found that when using spread spectrum codes in the signal domain, the CS measurement matrix may be simplified. This measurement scheme, named Compressive Spread Spectrum (CSS), allows for a simple, effective receiver design. In the observation, we also found that Bit error rate performance is degraded by the sub-sampling in the CS-enabled receivers; this may be remedied by including quantization in the receiver model. Bit error performance can be further improvised by the use of Hybrid Interference Cancellation in the received signal. In this paper we have proposed a new scheme, which can provide much better error performance for the CDMA receivers.

I. INTRODUCTION

The concept of compressive sensing [1, 2] is attracting high attention in the field of signal processing. The classical Shannon-Nyquist sampling theorem requires a signal to be sampled at twice its signal bandwidth; But the compressive sensing samples the signal at its information rate, which could be much lower. Compressive sensing is used to reconstruct a signal to a full Nyquist rate representation, but if only inference about information in the signal is desired, compressive signal processing is better suited [3]. Compressive signal processing is used when inference about information in a signal is the subject of interest, rather than the reconstruction of the signal. Compressive sensing and compressive signal processing, samples the signal using a typical sampling scheme with a randomized structure and then it exploits sparsity in the signal to enable sub-sampling.

In DSSS systems, the sparsity is in the selection of a code that used for transmission of a given data sequence. In the work we show how compressive signal processing can be applied to a spread spectrum receiver to lower the sampling rate at the receiver. This may lower the overall energy consumption of the device and/or lower the price of the Analog to Digital Converter (ADC). To exemplify this consider the following:- This work is based on a signal model used in the IEEE 802.15.4 standard [4], in which a baseband signal with a Nyquist frequency of 200kHz must be sampled. To show the benefit of lowering the sampling rate, we compare two ADCs from Analog Devices1: The AD7819 and the AD7813. The AD7819 is an 8-bit ADC with a maximum throughput of 200 kilosamples per second, whereas the AD7813 is an 8- or 10-bit ADC with a maximum throughput of 400 kilosamples per second. We are aware that 400 kilosamples per second is the Nyquist rate of the system and the sampling rate should be higher than this to comply with the Shannon-Nyquist sampling theorem. However, we use these two ADCs as they are almost identical in every aspect except for the sampling rate, making them perfect for comparison. In present IEEE 802.15.4 compliant receivers, an ADC similar to the AD7813 must be used to comply with Shannon-Nyquist, but if compressive signal processing is applied to lower the sampling rate by a factor of two, the AD7819 may be used instead. In the previous works, the CS is applied to the CDMA system. The reduced complexity RD implementation is also provided for the CDMA system, which performs equally well for CDMA signals but is simpler and cheaper to implement. The main idea behind is that to take fewer samples and to conserve power in the receiver. We have seen the performance of the CS CDMA receiver structure for the simple discrete case, when compared to a classic receiver structure and an RD receiver structure. Due to noise folding the CS approach suffers a penalty for down-sampling, but we show that if quantization is taken into account CS outperforms the classic receiver in some cases. Further we have shown the results for the modified CS-CDMA, which is more improvised as compare to the normal CS-CDMA. Modification in the CS-CDMA is provided by the Hybrid Interference Canceller, which is additionally introduced in the receiver structure. In a CDMA system, the performance of channel matched filter receivers are limited by the presence of multi-user interference. Hence receivers must be capable of suppressing these interferences. Interference suppression or interference cancellation works by subtracting the interfering users from the received signal, thus allowing an improved error rate for the user of interest.

II. TRANSMITTER STRUCTURE

In both the transmitter and the receiver structure we treat the signal symbol-by-symbol, where each symbol may be a single bit of information or a block of bits. Let

$\mathbf{b}_k \in \{\pm 1\}^{N\times 1}$ be a binary vector, signifying the *k*th symbol to be transmitted and consisting of N information bits. Now define a binary pseudo-random noise (PRN) sequence as $\mathbf{c}_k \in \{\pm 1\}^{C\times 1}$. These two binary vectors are the discrete equivalents of an information signal and a PRN signal, $b_k(t)$ and $c_k(t)$, respectively as shown in Fig. 1 and are defined as:

$$b_k(t) = \sum_{n=0}^{N-1} \mathrm{b}_k[n]\mathrm{rect}\left(\frac{t-nT_b}{T_b}\right),\ 0 \le t < NT_b$$

$$c_k(t) = \sum_{c=0}^{C-1} \mathrm{c}_k[c]\mathrm{rect}\left(\frac{t-cT_c}{T_c}\right),\ 0 \le t < CT_c$$

where T_b and T_c are the bit and chip duration, respectively, and $NT_b = CT_c$.
We define:

$$\mathrm{rect}(t) = \begin{cases} 1 & \text{if } 0 \le t < 1 \\ 0 & \text{otherwise.} \end{cases}$$

When multiplied, they form the spread spectrum data signal. $d_k(t) = b_k(t)c_k(t),\ 0 < t < NT_b$.
The notation used in the above may in some cases be simplified, as the choice of a PRN sequence might be implemented as a mapping from one bit or a block of bits directly to a given sequence of chips, as done in e.g. IEEE 802.15.4. In the following the signal model we define is based on the IEEE 802.15.4 standard's 2.4 GHz band specification. This means the encoding using DSSS may be written as a matrix-vector product, with $M = 2^N$ possible data signals:

$$d_k(t) = \mathbf{\Psi}(t)\boldsymbol{\alpha}_k, \text{ where}$$

$$\mathbf{\Psi}(t) = \begin{bmatrix} d_1(t) \\ d_2(t) \\ \vdots \\ d_M(t) \end{bmatrix}^{\mathrm{T}},\ 0 \le t < NT_b,$$

where $\psi(t)$ is a dictionary of possible data signals and α_k is a sparse vector with only one non-zero entry, namely the entry that selects a given PRN sequence from the dictionary. It may also be considered a symbol vector as it corresponds to the kth symbol being transmitted. The sparsity of k is what enables us to use compressive sensing for demodulation. The sparsity of the signal lies in which PRN sequence is chosen for transmission. So that even-indexed chips in $d_k(t)$ are transmitted in the in phase path and odd-indexed chips in the quadrature-phase path. In the following we only state the equations for the in phase path, but similar expressions may be derived for the quadrature-phase part. The resulting data signals are then used to modulate some pulse shape function, g(t):

$$s_k^I(t) = \mathbf{\Psi}^I(t)\boldsymbol{\alpha}_k$$

$$\mathbf{\Psi}^I(t) = \begin{bmatrix} \sum_{c\in S} d_1(t)g(t-cT_c) \\ \sum_{c\in S} d_2(t)g(t-cT_c) \\ \vdots \\ \sum_{c\in S} d_M(t)g(t-cT_c) \end{bmatrix}^{\mathrm{T}},\ S = \{0, 2, \ldots, C\}$$

Here the dictionary matrix has been recast into an in-phase version, with pulse shape function included.

III. SPREAD SPECTRUM DICTIONARY OF GOLD SEQUENCES

In spread spectrum signals, a possible dictionary is a set of Gold sequences, as used in e.g. GPS technology [20]. A set of Gold sequences is a special dictionary of binary sequences with very low auto and cross-correlation properties [21]. To generate a Gold dictionary, two maximum length sequences must be generated by two linear feedback shift registers (LFSR). A maximum length sequence is often denoted an m sequence (it has m elements), and is a special kind of pseudorandom noise sequence generated by a LFSR, such that it is periodic and produces a sequence of length 2m − 1. It is called a maximum length sequence as its period is at maximum length. The reason for the length being 2m −1 rather than 2m is that the state where all cells are zero must be avoided. To obtain an m-sequence, the LFSR must be carefully chosen as there is no algorithm for ensuring maximum length. However, there are many known LFSR setups for varying choices of m. Furthermore, the two m sequences must be chosen so that their periodic cross-correlation is three-valued and takes on only the values {−1,−t, t − 2}. When using such a CDMA dictionary, the received signalmust be sampled at a rate corresponding to the chip rate, where a chip is one entry in the received Gold sequences. If is sparse the information rate of the signal is much lower and it may be possible to decrease the sampling rate by using CS. In this paper, we use three Gold dictionary sizes: m = 5,m = 7 and m = 10.

IV. COMPRESSIVE SENSING RECEIVER STRUCTURE

In hardware compressive sensing sampling structures, such as the Random Demodulator [10, 11], a PRN sequence is mixed with the received signal followed by low-pass filtering. Due to the presence of a PRN sequence in a spread spectrum transmitter, which spreads the data signal, a compressive sensing-enabled receiver may merely use a repeated version of its matched filter, subsample the received signal and still demodulate the information. Before sampling the matched filter must be modified to contain not only a single chip pulse shape but as many chip pulse shapes as shall be contained per sample. This received signal vector may then be written as:

$$\mathbf{y}_k^I[\ell] = \int_{\ell T_c/\kappa}^{(\ell+1)T_c/\kappa} \theta_\ell(t) r_k^I(t)\mathrm{d}t, \text{ where}$$

$$\mathbf{\Theta}_{1/\kappa}(t) = \begin{bmatrix} \theta_0(t) \\ \theta_1(t) \\ \vdots \\ \theta_{L-1}(t) \end{bmatrix},\ \ \theta_j(t) = \sum_{c=j/\kappa}^{(j+1)/\kappa} g(t-cT_c),\ \ 0 \le t \le CT_c$$

Here each value of $\ell = 0, 1, \ldots, L$ signifies a collection of chips due to the subsampling where $L = C\kappa$ is the number of samples taken per symbol. $\kappa = \frac{k}{c} \in]0,1]$ is the undersampling ratio in the compressive sensing system and signifies the ratio between taken samples and Nyquist samples. In this work we limit ourselves to scenarios where $1/\kappa$ is an integer number, i.e. only an integer number of Nyquist samples are compressed together into one sample. To verify that the use of an additional PRN sequence at the receiver is unnecessary, we may look at the outcome of the subsampling ADC. Assuming a noise-free setting (n(t) = 0), the outcome becomes:

$$y_k^I[\ell] = \sum_{c=\ell/\kappa}^{(\ell+1)/\kappa} \int_{cT_c}^{(c+1)T_c} r_k^I(t) p_{\mathrm{PRN}}(t)\mathrm{d}t$$

$$= \sum_{c=\ell/\kappa}^{(\ell+1)/\kappa} \int_{cT_c}^{(c+1)T_c} \sum_{c'=0}^{C/2-1} b_k(t+c'T_c)c_k(t+c'T_c) \cdot g(t-nT_c)p_{\mathrm{PRN}}(t)\mathrm{d}t$$

Notice that the up and down-conversions have been assumed perfect and PPRN(t) is a new PRN sequence, added at the receiver as is done in the Random Demodulator receiver structure [10]. The symbol c′ denotes a chip picked out in dk(t) at the transmitter and used to avoid confusion with c, the chips added together into a sample at the receiver. The special indexing with Tc in connection with bk(t) and ck(t) is to pick out the chips in the in-phase path only, similar to what was done in (7). Because everything is multiplicative, it can be seen that ck(t + nTc) and PPRN(t) are synchronized and have the same chip rate, i.e. they may be viewed as a single PRN sequence. It follows that the multiplication of a PRN sequence at the receiver is unnecessary here. Because we wish to demodulate a signal, which is equivalent to a classification problem, it is not necessary for us to reconstruct the full original signal as is done in compressive sensing. Instead we use the recently introduced concept of compressive signal processing [3] to perform classification in the compressed domain. By classification, we mean to classify which of the signal candidates in the dictionary I and Q has been transmitted. This does not require reconstruction of the signal itself and may therefore be done with less computational complexity by using compressive signal processing, rather than classic compressive sensing algorithms, that reconstruct the full signal.

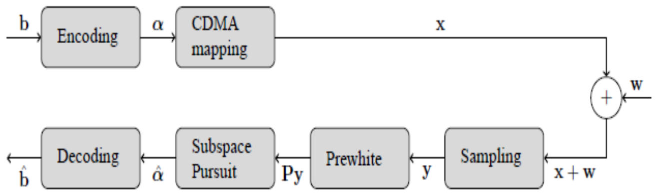


Fig. 1 Flow Diagram of discrete numerical experiment

V. MODIFICATION IN RECEIVER WITH HYBRID INTERFERENCE CANCELLER

The HIC combines both SIC and PIC to the correct proportion so that the receiver performance is enhanced to reach the near optimal level. The successive interference cancellation receiver is the simple, but requires high computational time, whereas the parallel interference cancellation receiver is more complex, but computational time is less. The best solution to get the desired improvement in performance is to have a perfect trade off between the computational time and receiver complexity, so that with minimum number of iterations, with minimum complexity, a better performance can be achieved. The proposed receiver has the PIC as the first stage of cancellation, where a few number of users who have the decision statistic above a certain threshold, are cancelled in parallel. The remaining users are given to the second stage of cancellation i.e SIC. The SIC cancel users, one at a time, until the desired user's signal is separated. Finally, the subtracted signal is given to a conventional receiver to get the users' information. The performance of the proposed HIC receiver solely depends on the threshold or decision static.

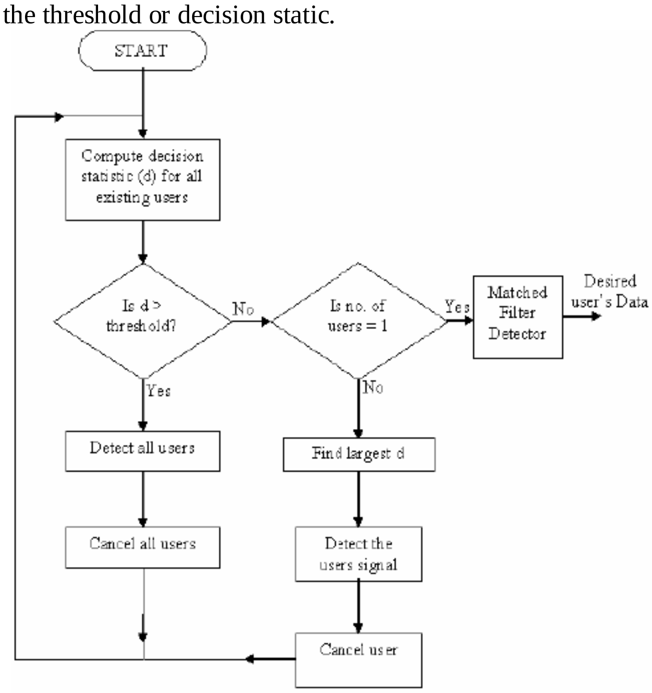


Fig. 2 Flow Diagram of hybrid interference Cancellation receiver

VI. RF NUMERICAL EXPERIMENT

To obtain more realistic communication-relevant results, we have extended the above discrete numerical experiment to a full transmitter/receiver simulation with RF up and down conversion and with root raised cosine pulse shaping and matched filter. This we have done to demonstrate that the results from Fig. translate to a realistic transmitter/receiver system. We have performed a numerical experiment in which we compare the Bit Error Rate (BER) of a classical receiver with that of a compressive sensing-enabled receiver. This is done for a range of Signal-to-Noise-Ratio (SNR) levels. The system used for this experiment is our MATLAB implementation of the physical layer of the IEEE 802.15.4 2450 MHz OQPSK radio band specification [4]. Each block of four bits is mapped into one of 32 binary chip sequences, according to the mapping in [4]. The chip sequence is then modulated using Offset Quadrature Phase Shift Keying (OQPSK). This standard has been chosen due to its widespread use, having been deployed already in many applications around the world and because it is a known standard to many scientists and engineers.The experiment is repeated for a range of SNRs or more specifically energy per bit per noise spectral density (Eb/N0). The noise is added in a bandwidth corresponding to that of the baseband signal, i.e. 2 MHz [4]. Our experiment is conducted by transmitting randomly generated data packets of length 127 × 8 = 1016 bits each (the maximum size of an IEEE 802.15.4 data packet). For each of the two tested methods and for each Eb/N0 level, bits are transmitted until the desired number of bits has been received in error.

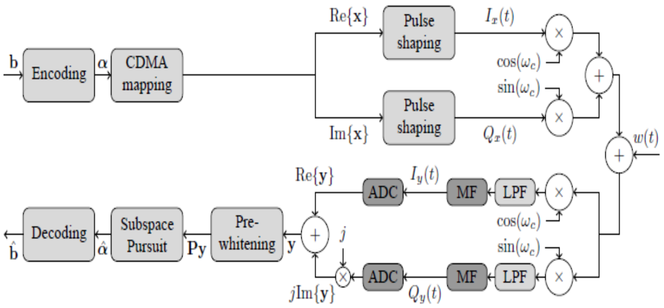
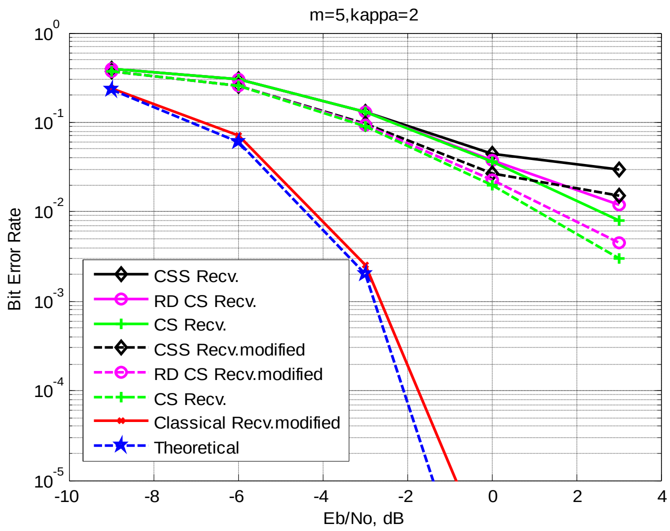


Fig. 3 Conceptual flow chart of the RF numerical experiment.

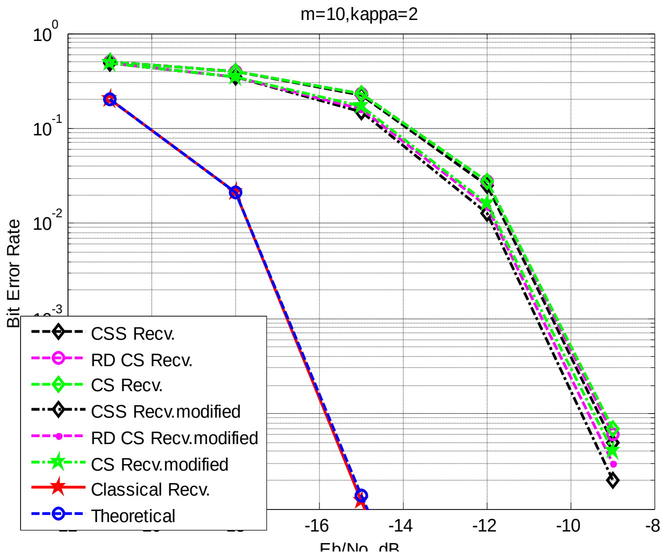
The results of the experiment are shown in Fig. 5. The theoretical curve is calculated using a modified version of the non-coherent MFSK equation used before:

$$P_b = \frac{M}{2(M-1)}\frac{1}{M}\sum_{k=2}^{M}(-1)^k\binom{M}{k}\cdot$$
$$\exp\left(\log_2(4)\frac{E_b}{N_0}\left(\frac{1}{k}-1\right)\right),$$

, where Eb/N0 is the energy per bit per noise spectral density and we multiply Eb/N0 with log2(4) because there are 4 constellation points in QPSK.



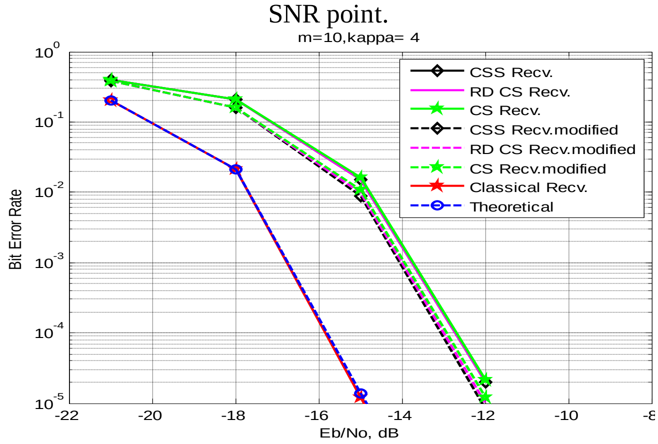
(a)



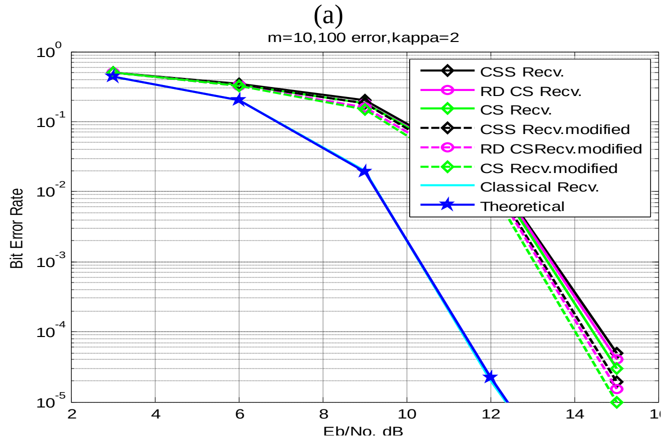
(b)

Fig. 4 BER versus SNR for different dictionary sizes and choices of κ . CS here is the Rademacher measurement scheme.

Simulations were run until 100 bit errors were found for each SNR point.



(a)

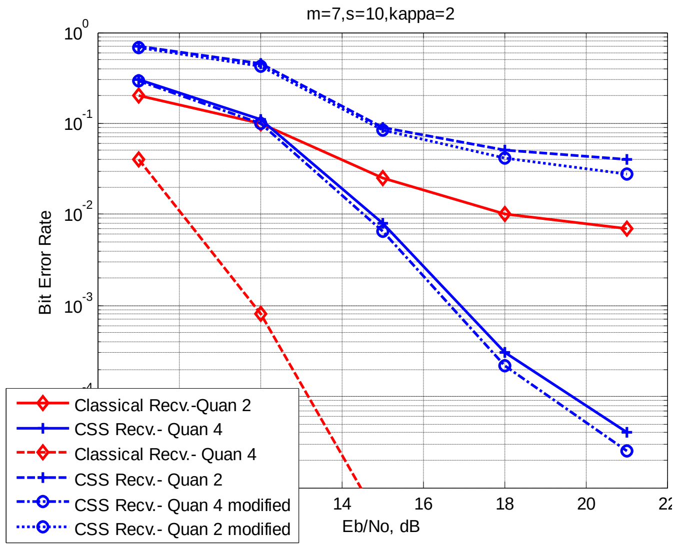


(b)

Fig.5: BER versus Eb/N0 for different dictionary sizes. CS here is the Rademacher measurement scheme. Simulations were run until 100 bit errors were found for each Eb/N0 point

A. RF Numerical Experiment with Quantization

It have been proposed to combat noise folding with quantization as a CS receiver is able to quantize the sampled signal better, since it takes fewer measurements. By better quantization we mean that if the CS receiver takes half as many samples, it may quantize twice as well at no additional cost. We have investigated this by applying uniform quantization to the RF experiment performed in the previous section. However, as simple QPSK modulation is used, only the sign matters for demodulation and therefore quantization has no effect in the simple case of S = 1 used so far. Therefore, we investigate S = 10 instead and used 2 bits of quantization per sample (i.e. 4 bits of quantization for CSS as κ = 2). This is merely intended as an example study to show that when taking into account quantization, CS may perform better than a classical receiver. The result of the numerical experiment is shown in Fig.



VII. CONCLUSION

In this work we apply HIC to CS-CDMA system and we show that it is possible to use a very simple measurement scheme at the receiver side to enable sub-sampling of the CDMA signal with the improvised error performance. We have shown that the performance of the proposed modified receivers is performing sufficiently better in BER performance for CS, CSS and also for RD scheme. However, we also show that when taking quantization into account, the proposed receiver model performs better in our example than a classical receiver with the same quantized bit rate. In the work we have shown that HIC used in spread spectrum receivers allows for a simplified architecture. Furthermore, we have shown that the problem of noise folding may be remedied in some cases by using quantization. The use of the HIC is also possible at lesser sampling rate, which saves the power. This can be verified by the experimentation also. The main result of this paper is the observation that in a spread spectrum receiver it is possible to use compressive sensing without generating a new PRN sequence and mixing it with the received signal and in an association with the HIC to get the optimized performance.

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