Performance of Non-uniform Modulations on the Nonlinear Satellite Channels

Nghia H. Ngo, Steven S. Pietrobon and S. Adrian Barbulescu

A technique of using a non-uniform 16QAM modulation combined with bit interleaved coded modulation and a new iterative decoding (BICM-ID) scheme is investigated . Simulation results of a parallel concatenated convolutional code (PCCC) shows a gain of 0.95 dB at a BER of 10^{-5} compared to conventional uniform 16QAM modulation in the linear channel. For a nonlinear satellite channel, this scheme offers a gain of 1.2 dB at the same BER compared to an uniform constellation.

Introduction: A non-uniform distribution of a uniformly spaced signal constellation is one option to shaping. In this technique, low energy transmitted signal points are used more frequently than high energy signal points, which results in the reduction of the average transmitted power and thus a shaping gain is obtained. When a powerful channel coding scheme such as turbo code are combined with a non-uniform constellation, a significant gain can be achieved [1]. In a conventional turbo decoder [2], extrinsic information of the data bits is iterated between the two soft input soft output (SISO) decoders. After a few iterations extrinsic information, which includes both data and parity bits, is fed back to the soft demapper in a BICM-ID system [3]. In this letter, we proposed a new iterative decoding scheme in which an extra soft-demapper is used to improve the channel estimation coming into the second SISO decoder. When combined with the non-uniform signal shaping technique for 16QAM modulation, it offers an excellent performance compared to a conventional BICM system with uniform constellation.

The ultimate goal of shaping is to generate a discrete Gaussian like distribution of the signal constellation. For example, to achieve a spectral efficiency of 2 bit/s/Hz using uniform 16QAM modulation, we need to map 2 information bits into a 4 bit 16QAM symbol. In this letter, we propose a new non-uniform 16QAM mapping scheme where one 16QAM symbol is represented by 6 bits with some overlapping. This results in a discrete Gaussian like distribution shown in Fig.1. We simulate this scheme with a PCCC [2] and the new iterative receiver scheme described in [4].

Design of Non-Uniform 16QAM Constellation: The system in this letter is modelled similar to [1], but the mapper uses 16QAM with a non-uniform distribution. Unlike in [1] where puncturing is used to shape the signal set, we map several three bit encoder binary symbols into one signal point according to the irregular mapping shown in Fig.1.

In one dimension, we map three bits $\underline{b}_t = (b_{t,0}, b_{t,1}, b_{t,2})$ to four unique PAM $c_t = g(\underline{b}_t)$ signal points. We also arrange these bits to ensure that the data bit is in the most protected position. In this case, the data bit is mapped onto the most significant bit (MSB) position of the signal constellation. Careful attention needs to be made in designing the mapping rule since the system performance is very sensitive to the arrangement of these bits. Our technique of selecting the best mapping was based on an extensive search for all possible mappings with a symmetric pattern in each dimension. In our actual 16QAM scheme on the nonlinear channel we map this 4PAM signal set to both in-phase and quadrature dimensions.

At the receiver, the soft output for each bit in the soft-demapper is calculated as (using -ln as our log-likelihood ratio (LLR) function)

$$\mathrm{L}(b_{t,i}) = -\ln\Big(\frac{P(b_{t,i}=1|r_t)}{P(b_{t,i}=0|r_t)}\Big) + \ln\Big(\frac{P(b_{t,i}=1)}{P(b_{t,i}=0)}\Big) \qquad (1)$$

Where r_t is the received signal and the second term "subtracts" the *a priori* information from the soft-output to ensure that it is not reused by the turbo decoder. Using Bayes' rule, we can derive the probability values as

$$\begin{aligned} P(b_{t,i}|r_t) &= \frac{P(b_{t,i}, r_t)}{P(r_t)} = \frac{\sum_{c_t:b_{t,i}} P(b_{t,i}, c_t, r_t)}{P(r_t)} \\ &= \frac{\sum_{c_t:b_{t,i}} P(r_t|c_t)P(c_t|b_{t,i})P(b_{t,i})}{P(r_t)} \end{aligned} \qquad (2)$$

where c_t is one of 2^m signal points and the summation is over all points c_t conditioned on bit $b_{t,i}$ being either 0 or 1. Thus, the LLRs become

$$\begin{aligned} \mathrm{L}(b_{t,i}) &= -\ln\Big(\frac{\sum_{c_t:b_{t,i}=1} P(r_t|c_t)P(c_t|b_{t,i}=1)}{\sum_{c_t:b_{t,i}=0} P(r_t|c_t)P(c_t|b_{t,i}=0)}\Big) \\ &= -\ln\Big(\frac{\sum_{c_t:b_{t,i}=1} \mathrm{f_t}(c_t)P(c_t|b_{t,i}=1)}{\sum_{c_t:b_{t,i}=0} \mathrm{f_t}(c_t)P(c_t|b_{t,i}=0)}\Big) \end{aligned} \qquad (3)$$

where $f_t(c_t) = \exp[\frac{-|r_t - c_t|^2}{N_0}]$. For a nonlinear channel, we replace c_t with $\tilde{c}_t$, the centroid estimation point. The *a priori* values $P(c_t|b_{t,i}=b)$ in the soft-demapper are calculated as

$$P(c_t|b_{t,i}=b) = \sum_{\substack{\underline{b}_t:b_{t,i}=b \\ g(\underline{b}_t)=c_t}} \prod_{\substack{k=0 \\ k\neq i}}^{n-1} P(b_{t,k}) \qquad (4)$$

where n is the number of the encoded bits per symbol, $P(b_{t,k})$ is the *a priori* probability of bit position k in a symbol at time t. This is equal to 0.5 in the first iteration and is updated through extrinsic values output from the turbo decoder for the next iteration.

New Iterative Decoding Scheme: The new iterative receiver structure is described in [4]. Initially, the soft-demapper M_1^{-1} calculates the extrinsic log likelihood ratio (LLR) for the data and parity bit sequences based on the received symbol r_t. *A priori* information in the -ln domain for the data and parity bits into M_1^{-1} are assumed to be zero in the first iteration. These soft output estimation sequences are multiplexed, de-interleaved using S-random interleavers and fed into the first decoder, $SISO_1$. The difference between a conventional PCCC turbo decoder and the decoder in this letter is that an extra soft-demapper, M_2^{-1}, is placed between the two SISO decoders. The soft-demapper M_2^{-1} is identical to M_1^{-1}. The updated LLR extrinsic information for the data and parity bits are estimated from the second soft-demapper M_2^{-1} in a similar fashion to that for M_1^{-1} and fed into $SISO_2$. The outputs from $SISO_2$ are interleaved and fed back to M_1^{-1} in the next iteration.

We denote iterations between the $SISO_2$ and the soft demapper M_1 and iterations between $SISO_1$ and $SISO_2$ as outer and inner iterations, respectively. It is desirable to pass the most updated extrinsic information between the code components and thus a number of iteration scheduling between these code components were investigated. We observed that for an uniform Gray mapped constellation and conventional turbo decoding, there is no error performance gain by performing outer iterations for the same decoder complexity. On the other hand, for our new iterative decoding scheme, an appropriate iteration scheduling between these code components showed some large gain. We therefore selected the best iteration schedule scheme in our Monte Carlo simulations.

Simulation Results: On both channels, the non-uniform constellation techniques follows the mapping rule presents in Fig.1. Simulation results for these channel are compared to an uniform constellation with the same spectral efficiency. For this purpose, we used a rate 1/3 turbo code for the non-uniform constellation and puncture it to rate 1/2 for the uniform constellation. This is to ensure that both schemes operate at the spectral efficiency of 2 bit/s/Hz.

The channel encoder uses two identical recursive convolutional codes in a parallel concatenated scheme with the feed forward and feedback polynomials of 37/23 in octal. The interleavers are of S-random types. The encoder memory is four, the information block length is 8192 and a complexity of 24 SISO decoder is used. The trellis is terminated by adding four bits which are equal to the feedback bits coming from each encoder. In conventional turbo decoding, we perform 1 outer 12 inner and 4 outer 3 inner for uniform and non-uniform constellations, respectively. In our new iterative decoding scheme, the best iteration scheduling schemes are 3 outer 4 inner and 12 outer 1 inner for the uniform and non-uniform constellation respectively.

A gain of about 0.95 dB for the linear AWGN channel is shown in Fig.2 at a BER of 10^{-5} and this is 1.05 dB away from the Shannon capacity. For a nonlinear channel, a slightly greater gain of 1.2 dB can be obtained and this is 1.43 dB away from the capacity of 16QAM square constellation in the nonlinear memoryless channel.

Conclusion: In this letter, we present the advantages of using a non-uniform constellation over a linear and nonlinear satellite channel through our simulations. An additional gain of 1.2 dB was obtained in the iterative demapping and decoding of the 16QAM system. The simulation results lead us to believe that potentially higher coding gains can be obtained for higher spectral efficiency systems using non-uniform constellations. This is especially useful since the power constraint is one of the major issues for satellite communications. The current approach to finding the best mapping based on an extensive search and is thus probably not suitable for very high order modulation. An algorithm to optimise the signal labelling is a subject of future research.

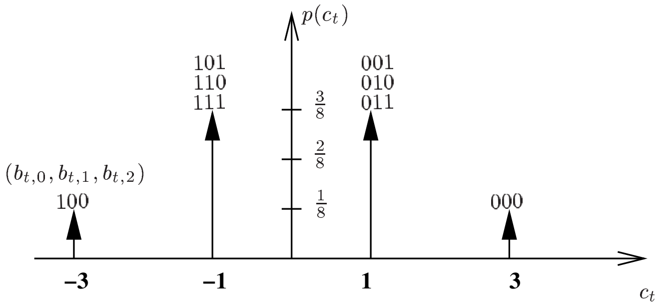

Fig. 1 Probability distribution and mapping rule for non-uniform 4PAM.

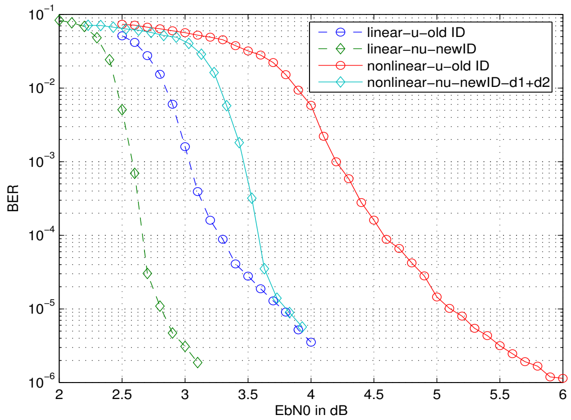

Fig. 2 Simulation Results.

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Nghia H. Ngo, Steven S. Pietrobon and S. Adrian Barbulescu(*University of South Australia, Mawson Lakes Boulevard, Mawson Lakes, SA 5095, Australia*)

E-mail: Nghia.Ngo@postgrads.unisa.edu.au

Steven S. Pietrobon (*Also with Small World Communication, 6 First Avenue, Payneham South SA 5070, Australia*)

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