Multi-Level Log-Likelihood Ratio Clipping in a Soft-Decision Near-Maximum Likelihood Detector

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***Abstract*—Consider the MIMO detection background. While the hard-decision Sphere Decoder has been widely and recently considered as the most promising near-optimal detector, this perspective might fall down during the soft-decision extension through a List Sphere Decoder (LSD). Due to the finiteness of the LSD list output - that does not necessarily allow for generating explicitly the Log-Likelihood Ratios (LLRs), even through a max-log approximation - the issue of how to set the missing reliabilities has been addressed. This paper presents existing works concerning the main trend. In particular, it consists in setting the LLR to a pre-defined value, this operation being commonly referred as LLR Clipping. We discuss this choice that has a significant impact on the system performance, by providing a brief state of the art of the existing solutions. In addition in the presented work, a novel solution lies in the multi-level bit mapping. Despite of its simplicity, it allows for low distortion approximated LLR computation. By simulation, the superiority of our method over the existing solutions, is shown.**

***Index Terms*—Log-Likelihood Ratio; bit clipping; List Sphere Decoder;**

I. Introduction

In order to achieve the 3GPP Long-Term Evolution (LTE) and 3GPP LTE-Advanced (LTE-A) requirements, Spatial Multiplexing Multiple-Input Multiple-Output (SM-MIMO) communication schemes have been implemented. In such a configuration and from the receiver point of view, a linear superposition of separately transmitted information symbols is observed, due to multiple transmit antennas that simultaneously send independent data streams. The interest of the detectors consists in recovering the transmitted symbols while approaching the channel capacity [1], and corresponds to an inverse problem with a finite-alphabet constraint.

As treated in several publications, the optimal - while exponentially complex in the number of transmit antennas and constellation size- Maximum Likelihood (ML) detector [2] can be efficiently approximated although avoiding an exhaustive search. In particular, some well-established techniques such as Sphere Decoding (SD) [3], Lattice Reduction (LR) [4] or a combination of both [5], have been shown to offer near-optimal performance. In the practical case of coded systems and due to their performance-complexity flexibility, the aforementioned detectors are straightforwardly modified in order to provide Soft-Output (SO). In the particular case of the classical K-Best, strongly presented in [6], a list $\mathcal{L}$ - of size $|\mathcal{L}|$ - of candidate solutions is generated from a subset of lattice points. This detector is denoted as the List SD (LSD) and will be considered in the present article, unless otherwise specified.

In order to achieve the channel capacity, the acknowledged way of bit transmission lies in providing redundancy and interleaving, denoted as channel coding. By focusing on the performance-complexity optimization, modern capacity-approaching codes lie in probabilistic coding schemes [7]. In particular, convolutional codes [8] led to the widely employed turbo codes [1], that are considered in this paper. In such schemes, soft-decisions of coded symbols are typically produced from the detector output, plus any available side information, and passed on to the decoder in the form of bit-wise Log-Likelihood Ratios (LLR), with the sign representing the decision and the magnitude representing the reliability. In the present article, the logical $\{0, 1\}$ bit patterns are respectively mapped onto the - zero-mean - set of amplitude levels $\{+1, -1\}$. From the decoder point of view, a close-to-zero value corresponds to an unreliable bit.

By evoking first that the LLR approximation is expressed as a function of both the probability of any bit representation [9], given the data in reception, and since the LSD output contains at least one solution, it is clear that a correct LLR sign is robustly reached, leading to promising uncoded performance [3], [4], [5]. However, due to the finite nature of the output $\mathcal{L}$ of the LSD that only offers a reduced-size list of candidates, its magnitude is regularly unknown, namely when the list does not contain both the hypothesis and its counter-hypothesis for a given bit. As a consequence, the performance in coded communication systems may be dramatically impacted.

While the exact LLR calculation is processed when possible, the main issue in the soft decision extension consists in how to estimate the missing LLR magnitude. To the best of the authors' knowledge, two distinct trends have been explored in the digital communications literature. Bäro has early met this problematic aspect and proposed a path augmentation [10] that consists in considering a bit-wise granularity during the tree construction. However, such a scheme requires high computational complexity and poor performance is reached if no apriori information is available [11].

A simple yet efficient operation has been widely studied.

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It is commonly referred as LLR Clipping (LC) [9], [12] and lies on setting the missing LLR values to a predefined value. However, the choice of the clipping level has a strong impact on the system performance [12], as addressed in the following, and any optimization is valuable.

Contribution: Our contributions can be summarized as follows:

- A multi-level bit mapping in the LC is presented, which leads to a significant coded performance improvement and offers convenient result improvement;
- The introduced solution preserves a general approach that makes it applicable to any LC.

Outline of the paper: In Section II, the notations are given and the problem statement is presented. In Section III, the necessity of introducing an accurate LC is explained and a detailed description of existing solutions is presented. The proposed solution is defended and introduced in Section IV. Section V aims at providing simulation results that show the superiority of the proposed solution. Finally, concluding remarks and perspectives are given in Section VI.

II. System model and problem statement

Consider a n_T-transmit and n_R-receive $n_T \times n_R$ MIMO system model. Assuming narrow-band flat-fading, the receive symbols vector $\mathbf{y} \in \mathbb{C}^{n_R}$ typically reads

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}, \quad (1)$$

where $\mathbf{H} \in \mathbb{C}^{n_R \times n_T}$ is the complex channel matrix, assumed to be perfectly known at the receiver, and $\mathbf{n}$ is a complex additive white Gaussian noise of variance σ_n^2. The entries of the transmit symbol vector $\mathbf{x}$ are independently withdrawn from a constellation set ξ, containing $|\xi|$ symbols, and such that $\mathbf{x} \in \xi^{n_T}$. Also, a layer ν is defined as a spatial stream, the number of spatial multiplexing data streams being $\min\{n_R, n_T\}$. Each channel inputs $\mathbf{b}_k^n \in \{\pm 1\}^{\nu \log_2\{|\xi|\}}$ is assigned to a symbol according to any encoding scheme, where n is any given layer and k is any bit position within the corresponding symbol. The whole symbols vector is mapped from a block of bit stream $\mathbf{c}_l$, by denoting l as the number of bits of the codeword with $1 \leq l \leq \nu \log_2\{|\xi|\}$. At this step, the uncoded or coded case is not distinguished. The block code may consider channel coding through the addition of redundancy and correlation, by introducing the code rate $R \leq 1$ - where $R = 1$ makes $\mathbf{c}$ corresponding to uncoded bits -, and interleaving [13]. By applying an efficient modulation and code rate scheme, the channel capacity is almost achieved at any SNR point [1].

The extrinsic LLR of the bit $\mathbf{b}_k^n$ is conditioned on the receive signal and is denoted $\Lambda(\mathbf{b}_k^n \mid \mathbf{y})$. Through the Bayes' theory and by using the *high SNR approximation*, which makes the classical max-log approximation accurate, it becomes advantageously rewrites [9]:

$$\Lambda(\mathbf{b}_k^n \mid \mathbf{y}) \approx -\frac{1}{2\sigma_n^2}\left((d^2_{\mathbf{b}_k^n=+1})_{\min} - (d^2_{\mathbf{b}_k^n=-1})_{\min}\right), \quad (2)$$

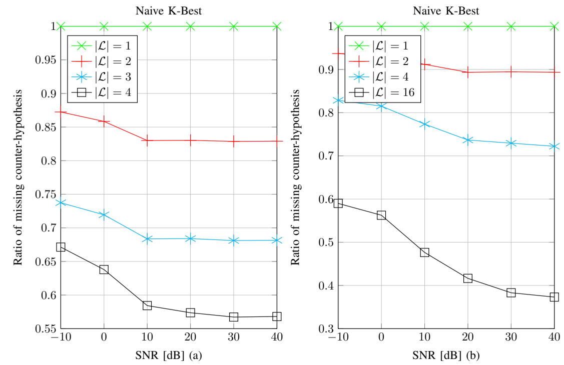

Fig. 1. Expected ratio of to-be-clipped values by considering the output of a naive K-Best, 4x4 complex MIMO system, QPSK (a) or 16QAM (b) modulation on each layer with $|\mathcal{L}| = \{1, 2, 3, 4\}$ and $|\mathcal{L}| = \{1, 2, 4, 16\}$, respectively, 10,000 simulations.

where $(d^2_{\mathbf{b}_k^n=+1})_{\min}$ and $(d^2_{\mathbf{b}_k^n=-1})_{\min}$ denote the minimal square euclidean distance between $\mathbf{y}$ and all the possibly transmit symbols vectors which are assigned to $\mathbf{b}_k^n = +1$ and to $\mathbf{b}_k^n = -1$, respectively. Also, the layer index n within the transmit data and the bit index k within any symbol inside are such that $1 \leq n \leq \nu$ and $1 \leq k \leq \log_2\{|\xi|\}$, respectively.

The importance of LC on both performance and complexity is highlighted. With this aim in view, the ratio of the occurrence of a missing counter-hypothesis as a function of $|\mathcal{L}|$ is depicted in Figure 1, for a 4×4 system, with QPSK or 16QAM modulations on each antenna.

It appears that the number of to-be-clipped LLR values depends on the SNR, on the list size $|\mathcal{L}|$ and on the modulation scheme. In particular in the 16QAM case and for a high range SNR, the ratio is 73% and 38% for $|\mathcal{L}|$ being 4 and 16, respectively.

Beyond this consideration, it is also clear that also the clipping value depends on the list size $|\mathcal{L}|$. As seen in Figure 1 and in the case of large $|\mathcal{L}|$, few missing counter-hypothesis remain. However, since a large neighbourhood study of lattice points has been processed, the corresponding bit sign is very improbable and should be clipped to a large value. In the case of a small $|\mathcal{L}|$, more missing counter-hypothesis are contained within the list output. Due to the partial nature of the study, the bit sign is not completely improbable, leading to choosing a small clipping value.

This idea has to be introduced in theory. It can be done through the introduction of the *mutual information* $\mathcal{I}\{\mathbf{b}_k^n, \Lambda_{clip}(\mathbf{b}_k^n)\}$ of any bit amplitude and any corresponding LC value, denoted as $\Lambda_{clip}(\mathbf{b}_k^n)$. By assuming that the bit sign in the LC is correct and $\Pr\{\mathbf{b}_k^n = +1\} = \Pr\{\mathbf{b}_k^n = -1\} = \frac{1}{2}$, the mutual information reads:

$$\mathcal{I}(\mathbf{b}_k^n, \Lambda_{clip}(\mathbf{b}_k^n)) = 1 - \frac{1}{2}\left(\log_2\left\{1 + e^{-\Lambda_{clip}(\mathbf{b}_k^n)}\right\} + \log_2\left\{1 + e^{\Lambda_{clip}(\mathbf{b}_k^n)}\right\}\right). \quad (3)$$

Subsequently, the impact of the clipping value choice can be observed. In particular, a solution for determining the

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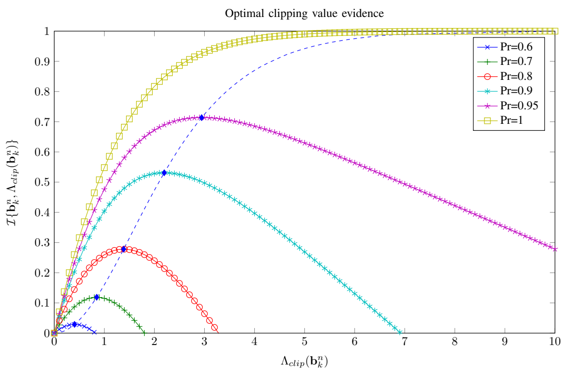

Fig. 2. Mutual information versus clipping value $\Lambda_{clip}(\mathbf{b}_k^n)$ for multiple $\Pr\{\Lambda_{clip}(\mathbf{b}_k^n)\}$ values.

LC setting relies on the mutual information maximization. Figure 2 plots the mutual information between $\mathbf{b}_k^n$ and the clipped detector output $\Lambda_{clip}(\mathbf{b}_k^n)$, versus the clipping value $\Lambda_{clip}(\mathbf{b}_k^n)$ and for different values of $\Pr\{\Lambda_{clip}(\mathbf{b}_k^n)\}$. Also, the maximal mutual information - given any $\Pr\{\Lambda_{clip}(\mathbf{b}_k^n)\}$ value - is pointed out by diamond marks and indicates the optimal clipping value.

Consequently, it is clear that there exists an optimal value that is not a constant. At least, it depends on the number of missing counter-hypothesis. Its evolution is plotted in Figure 2 with a dashed style. Also, while an efficient balance - namely $\Lambda_{clip}(\mathbf{b}_k^n) = 3$ [12] - can been found at intermediate clipping values, it has to be noticed that the LC choice has a significant impact on the coded performance. In particular, choosing the clipping level too high induces the decoder to assume a too high reliability for the bits with missing counter-hypothesis and consequently prevents error correction at these bit positions. Setting the clipping level too low limits the mutual information at the detector output and thus decreases its performance.

In order to address this problematic aspect, numerous techniques have been proposed and studied.

III. Summary of Existing Solutions

In case both the hypothesis and its counter-hypothesis given any bit arise in $\mathcal{L}$, the LLR is calculated according to Equation (2). Naively, we can consider that the soft information about any given bit $\mathbf{b}_k^n$ is essentially contained in $\mathcal{L}$. Indeed, if there are many entries in $\mathcal{L}$ with $\mathbf{b}_k^n = -1$, then it can be concluded that the likely value for $\mathbf{b}_k^n$ is indeed minus one, whereas if few entries occur, then the likely value is one. In the particular configuration of no arising counter-hypothesis within $\mathcal{L}$, Hochwald *et al.* proposed first a solution. It consists in setting its corresponding LLR value $\Lambda_{clip}(\mathbf{b}_k^n)$ to an extremely large value [9], namely ± 128, due to their high improbability. However, by proceeding this way, the importance given to unknown LLRs is too high and the channel coding gain is impaired.

A. Fixed LLR Clipping

Ideally, $\Lambda_{clip}(\mathbf{b}_k^n)$ should be different for each channel bit, such that the mutual information is maximized [12]. Nevertheless, a simple while robust proposal has been introduced [9], [5]. For a missing ± 1 bit value on any of the $\nu \log_2(|\xi|)$ locations, the corresponding LLR is set to ± 8, based on the argument that a missing bit value from the list of candidates makes it unlikely. As previously discussed, such a Fixed LC (FLC) value is efficient in the case of large $|\mathcal{L}|$ only. It can anyway be considered as a convenient upper bound [11], [14].

Another reasonably balanced FLC is commonly employed. The FLC value can be set to ± 3 and offers convenient performance [12]. Again, it has been previously discussed that such a FLC value is efficient in the case of small $|\mathcal{L}|$ only. In particular in [11], performance is depicted in the case of a naive LSD and $\Lambda_{clip}\mathbf{b}_k^n = 3$. Improved performance is offered compared to $\Lambda_{clip}\mathbf{b}_k^n = 8$, except with a very large study, namely with $|\mathcal{L}| = 64$. Consequently, this simple although efficient technique will be used as a reference in the simulation results.

By referring again to Figure 2, it must be highlighted that a constant clipping value that is much lower than the optimum value for a given bit position causes the channel decoder to largely ignore the clipped detector output values, which degrades its error-correction effectiveness. A clipping value that is much higher than its optimum value forces the channel decoder to assume that the clipped detector output values have the correct sign. In the case that this assumption is not correct, soft decisions on other bits must be compromised in order to meet the code constraints, leading to error propagation. Consequently, the optimal LC level strongly depends of the system configuration as well as the employed detector calibration.

B. Empirical LLR Clipping

Widdup *et al.* subsequently considered to use some information contained in the LSD output. In particular, since the costs of only the best $|\mathcal{L}|$ solutions are known, the others must be estimated from the knowledge that their cost is at least as high as that of the worst known point, namely the current radius [15]. This solution is also denoted as the last list entry [11] and offers a significant coded performance gain compared to the naive solution. However, this solution does not match with the widely used K-Best LSD nature, that does not take advantage (without early termination condition) of the radius constraint. Namely, contrary a depth-first search, the radius is not shrinked during the process. Even if the largest Euclidean distance of the counter-hypothesis is considered, such a clipping does not give importance enough to the fact that the counter-hypothesis does not appear in $\mathcal{L}$. Consequently, it leads to a significant performance loss [11].

That is why Kawamoto *et al.* introduced a likelihood function generation for selecting an appropriate clipping value [16], [17]. Briefly, this solution lies in a statistical study of the radius and an empirical result. In particular, the expectation of the minimum squared Euclidean distance among the bits within

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the list output that offers both the hypothesis and counter-hypothesis is calculated. This value is then grown up by 50% [16] in order to increase its weight. Thus, the LC value is obtained.

The presented solution lies in empirical results that offer neither strong theoretical results nor convincing performances.

C. SNR-aware LLR Clipping

As previously mentioned, the LC value is expected to depend on the Signal-to-Noise Ratio (SNR), on $|\mathcal{L}|$ and on the modulation scheme (not of the code rate, as clearly introduced in [18], Theorem 1). However, the main issue relies on how to take these factors into account.

Milliner *et al.* recently proposed an analytical expression, especially detailed in [18] in the particular case of BPSK, that is claimed to provide the optimal clipping value in the AWGN case for any arbitrary code rate. Starting from the exact LLR definition, an approximation is proposed through the introduction of the channel state information based bit error probabilities of a bit to be +1 and -1.

The Symbol Error Rate (SER) for the ML detector in the case of a Pulse Amplitude Modulation (PAM) transmission over an AWGN SISO channel with effective SNR [11] is:

$$P_{S,PAM} = 2\left(1-\frac{1}{\sqrt{|\xi|}}\right) Q\left(\sqrt{\frac{3}{|\xi|-1}\sqrt{|\mathcal{L}|}\left(\frac{E_s}{N_0}\right)_i}\right), \quad (4)$$

where $Q(\cdot)$ denotes the Q-function and $\left(\frac{E_s}{N_0}\right)_i$ is the instantaneous SNR for the i-th detection layer (i-th component of the transmit signal) [11]. Following classical QAM extensions of the PAM SER expression yields the QAM SER, from which the QAM BER can be easily obtained. The predicted error probability $P_b(|\mathcal{L}|, \left(\frac{E_s}{N_0}\right)_i)$ then yields the SNR-aware $\Lambda_{clip}(\mathbf{b}_k^n)$ value for bits in the i-th detection layer:

$$\Lambda_{clip,i}(\mathbf{b}_k^n) \approx -\ln P_b(|\mathcal{L}|, \left(\frac{E_s}{N_0}\right)_i). \quad (5)$$

This solution offers convenient results. In particular, it outperforms any FLC solution [11]. However, a drawback remain. Due to the AWGN assumption, the technique is optimal only in mean over multiple Rayleigh channel realizations. Also, the whole bit sequence is considered. In particular, the additional knowledge about the bit position is not taken into account.

By considering the aforementioned techniques, a simple while efficient optimization is introduced in the following.

IV. Proposed solution

An original approach lies in taking into account the multi-level bit mapping nature of Quadrature Amplitude Modulation (QAM) which is a multi-level bits-to-symbol mapping. Every symbol correspond to a codeword. They are each characterized by a different mean Euclidean distance, and hence a different level of protection against noise and amplitude impairments. This aspect depends on the bit position within the bit sequence. In Figure 3, the example of 16QAM case is shown. In particular for this case, two levels exist comprising Most Significant Bits (MSB), with a bold representation, and Least Significant Bits (LSB) for each signal point. For 64QAM, three levels of protection exist while only one exists for 4QAM. The latter being actually not strictly a multi-level mapping since the protection level is the same among the codeword. In the following, for the sake of simplicity and without loss of generality, only the 16QAM case will be considered.

In the LTE-A downlink case [19], QAM modulations with a multilevel Gray mapping can be partitioned into square subsets with minimum mean intrasubset Euclidean distance. In Figure 3, the MSBs of a signal point determine in which subset it is located.

This idea is novel in this context. In [12], the multilevel

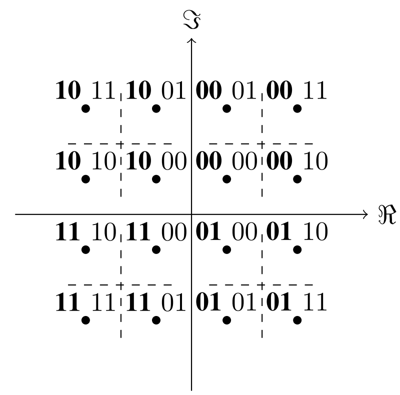

Fig. 3. 16QAM modulation constellation for LTE-A downlink [19]

bit mapping is considered in order to reduce the detector complexity, which becomes nearly independent of the signal constellation size. However, there is no link in the LLR calculation, and in particular in the LC.

Analytical expressions could be introduced from [20] and with required updates. However, Figure 4 exhibits more clearly the multilevel impact on LLRs in the particular case of a trivial 4×4 AWGN MIMO channel with 16QAM modulations on each layer. As it is shown, the LLRs are distributed differently depending on the bit index and the SNR. In particular, there is no maximal value for MSB, while there is a maximal value for positive LSB, from a Soft-Decision ML output. Also, consistently with previous discussions, this maximal value is different according to the SNR. The idea presented here consists in exploiting this additional knowledge in order to apply different clipping values to generate lower-distortion approximated LLR.

Through a SNR normalization, the positive LSB are shown in Figure 5 to be still upper bounded by a constant value in the Rayleigh channel case, this upper bound is marked by diamonds. For multiple SNR, the distribution of LLRs normalized by SNR is depicted in a 4×4 complex MIMO system. Similarly to [11], the proposed technique is SNR-aware and consider the modulation type as well. It does not consider the list size and offers an optimization by considering the additional knowledge of the bit position.

In particular, the positive LSB LLRs have been shown to

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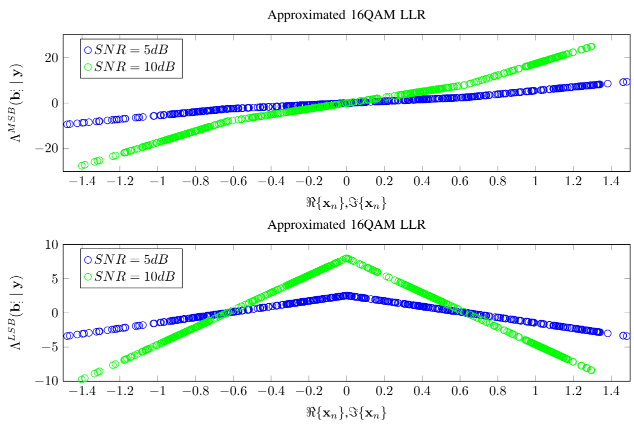

Fig. 4. Separated $\Lambda^{LSB}(\mathbf{b}_k^n \mid \mathbf{y})$ and $\Lambda^{MSB}(\mathbf{b}_k^n \mid \mathbf{y})$ representation of $\Lambda(\mathbf{b}_k^n \mid \mathbf{y})$ as a function of both the real and imaginary parts of any transmit symbol, for multiple SNR values, 4×4 complex MIMO system, 16QAM modulation on each layer, 800 distinct values per bit weight.

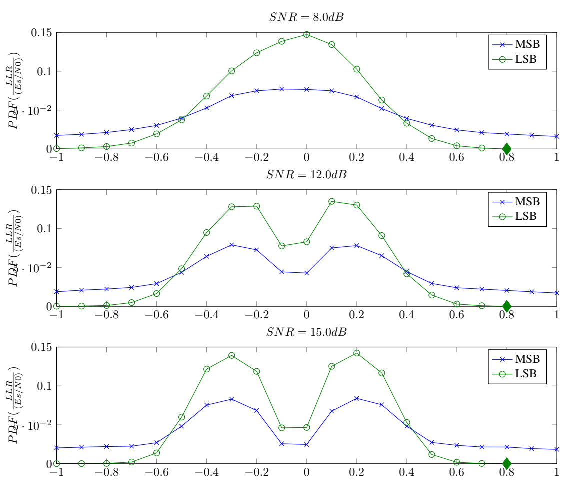

Fig. 5. Probability Density Function (PDF) of LLRs normalized soft ML output, 4×4 complex MIMO system, 800 distinct values per bit weight.

offer a constant maximal magnitude, through a SNR normalization step. Concerning the negative LSB and the MSB, the LLRs still depend on the SNR. This point is more clearly illustrated in Figure 6, where the maximum normalized LLRs from the soft-decision ML output are plotted as a function of SNR for different bit positions. From Figure 6, an efficient $\Lambda_{clip}^{LSB^+}(\mathbf{b}_k^n)$ is obtained. Also, the authors highlight that it is constant over the SNR range and independent of the number of clipped bits. This is not the case for the other bit positions which depend on the SNR and on the number of clipped bits. Nevertheless, further note that the absolute clipping values for MSB are the same:

$$\Lambda_{clip}^{MSB^-}(\mathbf{b}_k^n) = -\Lambda_{clip}^{MSB^+}(\mathbf{b}_k^n). \tag{6}$$

The efficiency of the proposed solution is presented through coded performance comparisons in the next section.

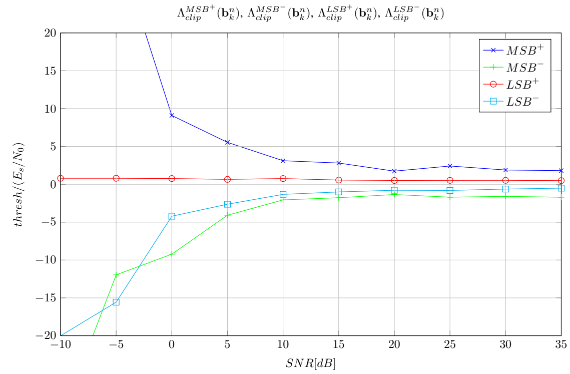

Fig. 6. Thresholds evolution as a function of SNR, 4×4 MIMO Rayleigh channel, $\frac{1000|\xi|\nu}{2}$ simulations per bit weight and per SNR value.

V. Simulation results

For the sake of comparability of the provided simulation results, the employed solution lies on considering a reference Soft-Decision ML detectors, which provides all the needed LLR values through an explicit computation. A certain percentage of these values are arbitrary discarded. Such a scheme is denoted as the *eclipsed* ML detector, and strictly corresponds to a LSD. The interest lies in keeping under control the percentage of missing LLR. The missing values are then clipped according to the proposed solution. The position of the discarded value is done randomly in order to make the ratio uniformly distributed between all the bit positions. Consequently, the real impact of the proposed solution is shown. In particular in the presented simulations, the ratio is defined as the proportion of forced clipped value from the soft-decision ML output. In the case of a ratio of p: $p\%$ of the provided LLRs are clipped according to any clipping technique. The $100 - p\%$ other LLR values remain explicitly computed. All the simulation results are compared to the ML detector with max-log approximation, used as a reference, and to the constant ± 3 clipping value, which has been previously shown to be a very efficient empirical result.

By considering the simulation conditions below, Figure 7 shows the BER and BLock Error Rate (BLER) error performance of the *eclipsed* ML detector in a 4×4 SM-MIMO system. At the transmitter, binary information data bits are first serial-to-parallel-converted into four data streams and are segmented into blocks containing a selected number of bits per packet frame according to the employed Modulation and Coding Scheme (MCS). The information data sequence is encoded by Turbo coding with memory 2 code and $1 + D + D^2$ and $1 + D^2$ feedback and feedforward polynomials, respectively. Also, the data is interleaved from a Look-Up Table. The original coding rate $R = \frac{1}{3}$ and then punctured according to the coding rate of $R = \frac{3}{4}$ when the resultant encoded sequence is data-modulated. The considered modulation format is 16QAM only with multilevel LTE-A bit mapping. Blocks of information bits are fed to the channel encoder, and subsequently transmitted over a Rayleigh fading

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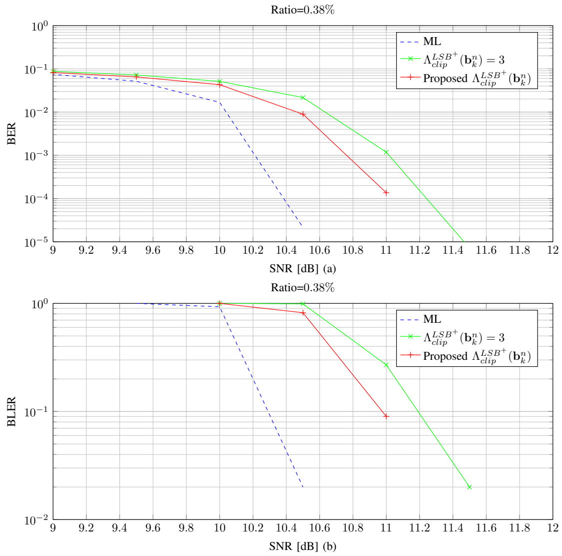

Fig. 7. BER (a) and BLER (b) performance plots, 4×4 complex MIMO system, 16QAM modulation and $R = \frac{3}{4}$ on each layer, 100 simulated blocks (a), 3×10^5 simulated bits (b).

MIMO channel, by forcing the 1,049 transmit symbols per block, independently of the employed modulation. It presently corresponds to 1008 transmit redundant bits per block, to within one rounding. For each block, 20 half iterations within the turbo decoder have been performed. We consider it is sufficient for achieving the convergence of the turbo decoder and consequently for reaching the maximal performance.

Both the BER and BLER performances confirm the efficiency of our proposed technique for 38% of clipping values. This proportion matches with the ratio obtained through a naive LSD with $|\mathcal{L}| = 16$ in a 4×4 MIMO system with 16QAM modulation on each layer, as depicted in Figure 1. Such a LSD calibration may be shown to allow near-optimal performance in the hard-decision case [6]. As it is shown in Figure 7, the proposed technique outperforms the clipping of ± 3 for every bit position proposed in [12] by $0.26\ dB$ for a BER of 10^{-3} and by $0.21\ dB$ for a BLER of 10^{-1}.

VI. Conclusion

This paper proposed a novel technique that allows for low distortion approximated LLR computation at the output of a soft-decision near-ML detector. Moreover, such approximation is judiciously applied depending on the actual bit-mapping indexing used by QAM constellations. To the best of the authors' knowledge, no method presented so far exploited this information to solve the problem of LC for the class of receivers considered and such a technical solution offers a significant performance gain. Also, the presented technique is general and may be applied to any more advanced technique.

Acknowledgment

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References

[1] C. Berrou, A. Glavieux and P. Thitimajshima. Near Shannon limit error-correcting coding and decoding: Turbo-codes (1). *Communications, IEEE International Conference on*, 2:1064–1070, May 1993.

[2] W. Van Etten. Maximum likelihood receiver for multiple channel transmission systems. *Communications, IEEE Transactions on*, pages 276–283, Feb. 1976.

[3] E. Agrell, T. Eriksson, A. Vardy and K. Zeger. Closest point search in lattices. *Information Theory, IEEE Transactions on*, 48(8):2201–2214, Aug. 2002.

[4] D. Wübben, R. Böhnke, V. Kühn and K.-D. Kammeyer. Near-maximum-likelihood detection of MIMO systems using MMSE-based lattice reduction. *Communications, IEEE International Conference on*, 2:798–802, June 2004.

[5] X.F. Qi and K. Holt. A Lattice-Reduction-Aided Soft Demapper for High-Rate Coded MIMO-OFDM Systems. *Signal Processing Letters, IEEE*, 14(5):305–308, May 2007.

[6] K.-W. Wong, C.-Y. Tsui, S.-K. Cheng and W.-H. Mow. A VLSI Architecture of a K-Best Lattice Decoding Algorithm For MIMO Channels. *Circuits and Systems, IEEE International Symposium on*, 3:273–276, May 2002.

[7] D.J. Jr Costello and G.D. Jr Forney. Channel Coding: The Road to Channel Capacity. volume 95, pages 1150–1177, Feb. 2007.

[8] P. Elias. Channel Coding: The Road to Channel Capacity. pages 37–46, March 1955.

[9] B.M. Hochwald and S. ten Brink. Achieving Near-Capacity on a Multiple-Antenna Channel. *Communications, IEEE Transactions on*, 51(3):389–399, Mar. 2003.

[10] S. Bäro, J. Hagenauer and M. Witzke. Iterative detection of MIMO transmission using a list-sequential (LISS) detector. *Communications, IEEE International Conference on*, 4:2653–2657, May 2003.

[11] D.L. Milliner, E. Zimmermann, J.R. Barry and G. Fettweis. Channel State Information Based LLR Clipping in List MIMO Detection. *Personal, Indoor and Mobile Radio Communications International, IEEE International Symposium on*, pages 1–5, Sept. 2008.

[12] Y.L.C. de Jong and T.J. Willink. Iterative Tree Search Detection for MIMO Wireless Systems. *Communications, IEEE Transactions on*, 53(6):930–935, June 2005.

[13] S. Benedetto, D. Divsalar, G. Montorsi and F. Pollara. Serial Concatenation of Interleaved Codes: Performance Analysis, Design, and Iterative Decoding. *Information Theory, IEEE Transactions on*, 44(3):909–926, May 1998.

[14] M. Myllyla, J. Antikainen, M. Juntti and J.R. Cavallaro. The effect of LLR clipping to the complexity of list sphere detector algorithms. *Signals, Systems and Computers, Asilomar, Conference on*, pages 1559–1563, Nov. 2007.

[15] B. Widdup, G. Woodward and G. Knagge. A Highly-Parallel VLSI Architecture for a List Sphere Detector. *Communications, IEEE International Conference on*, 5:2720–2725, June 2004.

[16] J. Kawamoto, H. Kawai, N. Maeda, K. Higuchi and M. Sawahashi. Investigations on likelihood function for QRM-MLD combined with MMSE-based multipath interference canceller suitable to soft-decision turbo decoding in broadband CDMA MIMO multiplexing. *Spread Spectrum Techniques and Applications, IEEE International Symposium on*, pages 628–633, Aug. 2004.

[17] K. Higuchi, H. Kawai, N. Maeda, M. Sawahashi, T. Itoh, Y. Kakura, A. Ushirokawa and H. Seki. Likelihood function for QRM-MLD suitable for soft-decision turbo decoding and its performance for OFCDM MIMO multiplexing in multipath fading channel. *Personal, Indoor and Mobile Radio Communications, IEEE International Symposium on*, 2:1142–1148, Sep. 2004.

[18] E. Zimmermann, D.L. Milliner, J.R. Barry and G. Fettweis. Optimal LLR Clipping Levels for Mixed Hard/Soft Output Detection. *Global Telecommunications Conference, IEEE*, pages 1–5, Dec. 2008.

[19] 3GPP. Technical Specification Group Radio Access Network; Spreading and modulation (FDD) v9.1.0. TR 25.213, 3rd Generation Partnership Project (3GPP), 2009.

[20] C.C. Wang. A Bandwidth-Efficient Binary Turbo Coded Waveform Using QAM Signaling. *Communications, Circuits and Systems and West Sino Expositions, IEEE International Conference on*, 1:37–41, July 2002.