Channel State Information Based LLR Clipping in List MIMO Detection

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***Abstract*—Suboptimal detection schemes, such as list MIMO detection, often face the challenge of having to "guess" at the decision reliability for some of the detected bits. A simple yet effective way of doing this is to set the maximum magnitudes of the associated log-likelihood-ratios (LLRs) to a certain predefined value: LLR clipping. However, the choice of the clipping level has a significant impact on the system performance. A majority of prior approaches attempted to determine appropriate clipping levels by manual optimization. In this work we propose to use an SNR-aware approach for calculating the LLR clipping levels in list MIMO detection. The proposed scheme exploits knowledge of the channel state information to determine the instantaneous bit error probability of the list detector, and from this an appropriate level for clipping of the LLRs. Simulation results show that this strategy outperforms schemes using a fixed clipping level.**

I. Introduction

In many communications problems, implementing the optimal detector is of prohibitive complexity. Multiple antenna (MIMO) systems are a prominent example where a significant amount of research has been performed in order to find low complexity detection schemes with near-optimal performance. List MIMO detectors [1]–[3] are attractive in this respect, as they enable a flexible trade-off between complexity and performance. These schemes often suffer from the inability to generate exact (or even any) reliability information for some of the detected bits. That is, while the sign of the log-likelihood ratio (LLR) can be determined with reasonably low probability of error, its magnitude is often unknown. Numerous approaches have been proposed for addressing this problem, e.g. bit flipping [4] and path augmentation [2], yet many of these schemes require high complexity.

A low complexity approach for addressing this problem is to set the maximum LLR magnitude for detected bits to a predefined value – LLR clipping [1], [5]. The selection of the clipping level has a significant impact on the achievable performance in coded communication systems [5]. Choosing the clipping level too high induces the decoder to assume overly high reliability for bits with missing counter-hypotheses, potentially preventing decision errors occurring at these bit positions from being corrected. Conversely, setting the clipping level too low substantially distorts the soft output for bits with counter-hypotheses, also leading to a performance loss.

Most previous approaches for computing the LLR clipping level have relied upon selecting a fixed clipping level, usually based on an attempt to maximize the mutual information at the detector output over the choice of the clipping level. Such fixed LLR clipping (FLC) level schemes possess the obvious advantage that, once the FLC level has been selected, they are easy to implement. Drawbacks of FLC schemes include an involved selection process for the clipping level [5], [6]. Finally, and perhaps most importantly, these approaches are, as we will show, limited in the error rate they can achieve.

In this work a low complexity approach for computing *SNR-aware LLR clipping (SLC)* levels, based on an estimate of the bit error probability at the detector output, is proposed. The K-best algorithm [7], [8], which is a breadth-first tree search detection scheme, is used as a representative suboptimal detection algorithm[1]. In [9] it was shown that LLR clipping has more effect on system performance when using suboptimal algorithms such as the K-best algorithm. For a given channel realization and list length of the algorithm, we calculate the instantaneous error probability on the different detected layers. This information is then used to predict the LLR clipping level. The additional complexity of the proposed approach can be considered almost negligible in coded systems. Results show that the proposed scheme can outperform even the best FLC schemes for coded MIMO communication systems.

The remainder of this paper is organized as follows: after discussing the employed system model in Section II, Section III details fundamentals related to LLR clipping. Section IV reviews selected prior work on LLR clipping and motivates and describes our approach for variable LLR clipping based on channel state information (CSI). In Section V we provide simulation results illustrating the effectiveness of our scheme. Finally, conclusions are drawn in Section VI.

II. System Model

Consider an $N_T \times N_R$ MIMO system based on a BICM transmit strategy as depicted in Fig. 1: the vector $\mathbf{u}$ of i.i.d. information bits is encoded and interleaved. The resulting

[1]The K-best algorithm is a specific realization of the more general M-algorithm [3].

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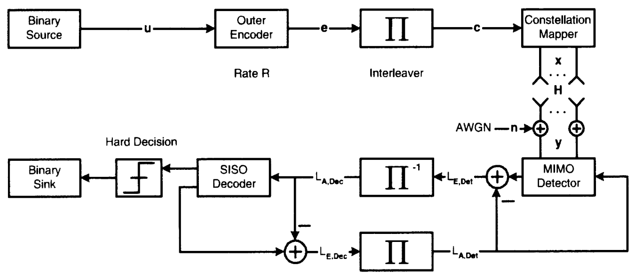

Figure 1. System model using a BICM transmit strategy.

coded bit stream is partitioned into blocks $\mathbf{c}$ of $N_T \cdot L$ bits and mapped onto a vector symbol $\mathbf{x} \in \mathcal{X}$ whose components are taken from some complex constellation $\mathcal{C}$ (e.g. Gray mapped 64-QAM). Here, L denotes the number of bits per complex QAM symbol, resulting in $q = |\mathcal{C}| = 2^L$ different constellation points (i.e. q is the square of the PAM alphabet size). We consider transmission over a flat fading channel.

In the equivalent discrete-time base-band model, the received signal $\mathbf{y}$ is thus given by:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}, \tag{1}$$

where $\mathbf{H} \in \mathbb{C}^{N_R \times N_T}$ is the channel transfer matrix which is assumed to be perfectly known at the receiver. The entries of $\mathbf{H}$ are realizations of zero mean i.i.d. complex Gaussian random processes of variance 1 (passive subchannels). The average transmit energy is normalized such that $\mathcal{E}\{\mathbf{x}\mathbf{x}^{\mathrm{H}}\} = E_s/N_T\ \mathbf{I}$. The vector $\mathbf{n} \in \mathbb{C}^{N_R \times 1}$ represents the receiver noise whose components are zero mean i.i.d. complex Gaussian random variables with variance $N_0/2$ per real dimension: $\mathcal{E}\{\mathbf{n}\mathbf{n}^{\mathrm{H}}\} = N_0\ \mathbf{I}$. The signal-to-noise ratio (SNR) at each receive antenna is hence given by $\mathrm{SNR} = E_s/N_0$.

III. Fundamentals

In a coded system, the detector in Fig. 1 has to generate reliability information, or "soft output", for each of the code bits $c_{m,l}$ in $\mathbf{x}$, where $m \in \{1, \ldots, N_T\}$ is the symbol index, and $l \in \{1, \ldots, L\}$ the bit index in the m-th symbol. Since we are dealing with binary numbers, this information is conveniently expressed in the form of an LLR:

$$L(c_{m,l}|\mathbf{y}, \mathbf{H}) := \ln \frac{P\left[c_{m,l} = +1|\mathbf{y}, \mathbf{H}\right]}{P\left[c_{m,l} = -1|\mathbf{y}, \mathbf{H}\right]} \tag{2}$$

$$= \ln \frac{\sum\limits_{\hat{\mathbf{x}} \in \mathcal{X}_{m,l}^{+1}} \exp\left(\frac{-\|\mathbf{y}-\mathbf{H}\hat{\mathbf{x}}\|^2}{N_0} + \sum\limits_{j=1}^{N_T}\sum\limits_{k=1}^{L} \ln P\left[c_{j,k} = \hat{c}_{j,k}\right]\right)}{\sum\limits_{\hat{\mathbf{x}} \in \mathcal{X}_{m,l}^{-1}} \exp\left(\frac{-\|\mathbf{y}-\mathbf{H}\hat{\mathbf{x}}\|^2}{N_0} + \sum\limits_{j=1}^{N_T}\sum\limits_{k=1}^{L} \ln P\left[c_{j,k} = \hat{c}_{j,k}\right]\right)} \tag{3}$$

where $\mathcal{X}_{m,l}^{\pm 1}$ is the set of $2^{N_T \cdot L - 1}$ symbols $\hat{\mathbf{x}} \in \mathcal{X}$ for which $\hat{c}_{m,l} = \pm 1$ and $\hat{\mathbf{x}}$ denotes a certain hypothesis on the transmit sequence, with $\hat{\mathbf{c}}$ as the corresponding vector of code bits.

Evaluating (3) exactly by a brute-force approach is well known to require an effort growing exponentially in the number of transmitted bits per vector symbol. Several detection strategies [1]–[3], therefore construct a subset list $\mathcal{L} \subset \mathcal{X}$ of size $\ell = |\mathcal{L}|$ from which the LLRs are determined.

After application of the max-log approximation, the list detection version of (3) can be expressed as:

$$\begin{aligned} L(c_{m,l}|\mathbf{y}, \mathbf{H}) &:= \ln \frac{P\left[c_{m,l} = +1|\mathbf{y}, \mathbf{H}\right]}{P\left[c_{m,l} = -1|\mathbf{y}, \mathbf{H}\right]} \\ &\approx \max_{\hat{\mathbf{x}} \in \mathcal{L}_{m,l}^{+1}} \left\{ \frac{-\|\mathbf{y}-\mathbf{H}\hat{\mathbf{x}}\|^2}{N_0} + \sum_{j=1}^{N_T}\sum_{k=1}^{L} \ln P\left[c_{j,k} = \hat{c}_{j,k}\right] \right\} \\ &- \max_{\hat{\mathbf{x}} \in \mathcal{L}_{m,l}^{-1}} \left\{ \frac{-\|\mathbf{y}-\mathbf{H}\hat{\mathbf{x}}\|^2}{N_0} + \sum_{j=1}^{N_T}\sum_{k=1}^{L} \ln P\left[c_{j,k} = \hat{c}_{j,k}\right] \right\} \end{aligned} \tag{4}$$

with $\mathcal{L}_{m,l}^{\pm 1} = \mathcal{L} \cap \mathcal{X}_{m,l}^{\pm 1}$. List detection schemes necessarily exclude many hypotheses on the transmit sequence from their search, to achieve low complexity. As a result, there is a non-zero probability that counter-hypotheses are missing for some of the detected bits: $\mathcal{L} \cap \mathcal{X}_{m,l}^{\pm 1} = \emptyset$. In fact, counter-hypotheses will typically be available for the least reliable bits while they will often be missing for the most reliable bits. This problem is a direct result of the fact that, by definition, hypotheses for unreliable bits have very similar Euclidean distances, while flipping the value of a reliably detected bit will result in a drastic change of Euclidean distance. Tree search schemes typically aim at constructing their list such that it contains the ℓ hypotheses with lowest Euclidean distance to the received signal. Following the above argument, it is easily seen that the aforementioned detectors typically have to invest much more effort in determining precise LLRs for the reliable bits than for the unreliable ones, see [6]. Since the overall goal is to minimize detection complexity (e.g. use small ℓ), reliable soft information is often unavailable for some of the received bits.

In such a situation, one has to resort to some estimation of the corresponding LLR value. Numerous techniques have been proposed to address this issue, e.g.:

- LLR clipping [1], [5], [6]: the maximum magnitudes for LLR information is assumed to be fixed or at a fixed worst-case distance from the MAP estimate. This strategy is extremely simple to implement, but the achievable performance depends crucially on an appropriate selection of the clipping level. An enhanced version of this approach is to add the fixed worst-case distance to an ML distance metric [10].
- Bit flipping/Chase decoding [4]: a counter-hypothesis is generated by taking the MAP estimate, flipping the bit of interest and calculating the Euclidean distance of the resulting hypothesis. This method is relatively complex to implement and due to the coupling between layers, the generated counter-hypotheses are often of low quality.
- Last list entry: one may use the last entry of the list (the one with largest Euclidean distance) as a lower bound on the Euclidean distance of the counter-hypothesis. However, this bound is rather loose such that using it will cause the LLRs to be "clipped" very aggressively,

causing a significant performance loss [11].[2].

- Path augmentation [2]: the dead-ends of the tree structure can be extended based on the output of a linear filter and/or available a priori information. This method is also rather complex and yields poor estimates if no a priori information is available [6].

Among the above mentioned approaches, LLR clipping offers excellent performance at negligible additional complexity. An open problem, however, is the appropriate selection of the clipping level without having to resort to time-consuming manual optimization. This is the motivation for our contribution.

IV. LLR CLIPPING

A. Fixed LLR Clipping Level (FLC)

This setup requires careful selection of the LLR clipping value. If the clipping level, denoted L_{clip}, is chosen too high, this prevents decision errors occurring at some bit positions from being corrected, resulting in poor performance. Conversely, setting the clipping level too low limits the mutual information at the detector output, also leading to decreased performance. Early FLC approaches relied on trial and error for choosing L_{clip}. An example of this is [1], which set $L_{\text{clip}} = 8$, after observing good performance results for this choice. In fact, this value can be considered to be a reasonable upper bound, as clipping the LLRs to this level has a negligible impact on the mutual information if the LLRs are exact i.e., obtained from an optimal detector [6].

More recent approaches for determining FLC values have relied upon maximizing the average mutual information at the detector output for specific system configurations [5], [6]. Assuming genie knowledge of the transmitted bits c_i in the code bit stream, this mutual information can be determined from the calculated LLRs $L_p(c_i)$ using [12]:

$$I(\mathbf{c}; L_p(\mathbf{c})) \approx 1 - \frac{1}{N_C}\sum_{n=1}^{N_C} \log_2\left(1 + \exp\left(-c_i \cdot L_p(c_i)\right)\right),$$

where $N_C \gg 1$ is the number of bits of the codeword. Using a mutual information based approach, it was shown in [5] that for many cases, $L_{\text{clip}} = 3$ is a reasonably good choice. This work was elaborated upon in [6] to show that the optimal clipping level depends on the list size and modulation alphabet. This is because the probability that counter-hypotheses are not available increases as the list size decreases.

Fig. 2, first presented in [6], plots the mutual information versus LLR clipping levels for the example case of 64-QAM transmission over a spatially and temporally i.i.d. Rayleigh fading 4×4 MIMO channel. A K-best algorithm [7], [8] with $K = 2, 4, 8, 16, 64$ is used for detection. The system parameters in [6] are the same as those used to obtain the results which will be presented in this work. We observe that

[2]The authors hypothesize that by using an intelligently selected metric corresponding to a distance larger than the last list entry would result in improved performance, although no analysis of this idea is provided here.

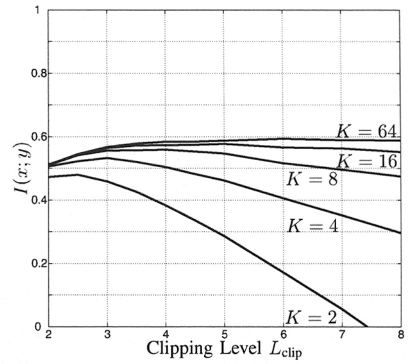

Figure 2. Mutual information as a function of the LLR clipping level, for different list sizes (4×4 MIMO, spatially uncorrelated Rayleigh fading, ergodic channel for a 64-QAM transmission alphabet where $E_b/N_0 = 13$dB).

K	2	4	8	16	64
4-QAM	3	3.5	4	5	–
64-QAM	2.5	3	4	5	6

Table I
FIXED LLR CLIPPING LEVELS L_{CLIP} DETERMINED IN [6] FOR A SPATIALLY AND TEMPORALLY I.I.D. FADING 4×4 MIMO CHANNEL.

for small list sizes the clipping level must be chosen relatively low, between 2 and 3, in order to maximize the mutual information. Conversely, for large list sizes, the clipping level must be chosen higher, between 6 and 8, to avoid the situation where mutual information is decreased by the LLR clipping.

Using an inappropriate LLR clipping level was found in [6] to result in a performance loss between 0.2 and 0.5dB for intermediate values of K, thus potentially offsetting any performance increase from using a larger list size. The clipping levels derived from the results presented in Fig. 2 are summarized in Table I for 4-QAM and 64-QAM modulation.

B. SNR-aware LLR Clipping (SLC)

We now extend the approach just described to one where L_{clip} is based on the available CSI. Consider again the definition of the LLRs from (2), which can be restated as:

$$\begin{aligned} L(c_{m,l}|\mathbf{y},\mathbf{H}) &:= \ln \frac{P\left[c_{m,l} = +1|\mathbf{y},\mathbf{H}\right]}{P\left[c_{m,l} = -1|\mathbf{y},\mathbf{H}\right]} \\ &= \hat{c}_{m,l} \ln \frac{P\left[c_{m,l} = \hat{c}_{m,l}|\mathbf{y},\mathbf{H}\right]}{P\left[c_{m,l} \neq \hat{c}_{m,l}|\mathbf{y},\mathbf{H}\right]} \\ &= \hat{c}_{m,l} \ln \frac{P\left[c_{m,l} = \hat{c}_{m,l}|\mathbf{y},\mathbf{H}\right]}{1 - P\left[c_{m,l} = \hat{c}_{m,l}|\mathbf{y},\mathbf{H}\right]} \end{aligned} \quad (5)$$

where $\hat{c}_{m,l}$ is the hard output estimate of the considered bit, as obtained from the detector. In the absence of a counter-hypothesis for this bit, the expression inside the logarithm of (5) is unknown. Furthermore, the knowledge of the Euclidean distance $d(\hat{\mathbf{x}}) = \|\mathbf{y} - \mathbf{H}\hat{\mathbf{x}}\|^2$ and associated $p(\mathbf{y}|\hat{\mathbf{x}})$ is not sufficient to establish a precise estimate for $P\left[c_{m,l} = \hat{c}_{m,l}|\mathbf{y},\mathbf{H}\right]$, as calculating the normalization factor $p(\mathbf{y})$ required to determine $P[\hat{\mathbf{x}}|\mathbf{y},\mathbf{H}]$ is computationally complex.

We therefore propose to resort to an approximation of (5), by averaging out the influence of $\mathbf{y}$. With this simplification, expression (5) becomes:

$$L(c_{m,l}|\mathbf{y},\mathbf{H}) \approx \hat{c}_{m,l} \ln \frac{P[c_{m,l}=\hat{c}_{m,l}|\mathbf{H}]}{1-P[c_{m,l}=\hat{c}_{m,l}|\mathbf{H}]} = \hat{c}_{m,l} \ln \frac{1-P_b(\mathbf{H})}{P_b(\mathbf{H})}, \quad (6)$$

where P_b is the bit error probability at the detector output, which needs to take into account the CSI, the current signal-to-noise ratio (SNR), the modulation format, and the configuration of the detector. This predicted bit error probability will be denoted as $P_b(\ell, \mathrm{SNR}_i)$ in the following, where SNR_i is the instantaneous SNR for the ith detection layer (ith component of the transmit signal) and given by:

$$\mathrm{SNR}_i = \frac{E_s}{N_T N_0} R_{i,i}^2, \quad (7)$$

where $R_{i,i}$ is the ith diagonal element of the upper triangular matrix $\mathbf{R}$ resulting from a QR-decomposition of the channel matrix, $\mathbf{H}=\mathbf{QR}$. To model the improvement in the quality of the detector output for larger list sizes, $P_b(\ell, SNR_i)$ is therefore given as a function of the list length ℓ.

This predicted error probability $P_b(\ell, \mathrm{SNR}_i)$ yields our SNR-aware L_{clip} value for bits in the ith detection layer:

$$L_{clip,i} := \ln \frac{1-P_b(\ell, \mathrm{SNR}_i)}{P_b(\ell, \mathrm{SNR}_i)} \approx -\ln P_b(\ell, \mathrm{SNR}_i). \quad (8)$$

We prove in a concurrent submission [13] that the optimal LLR clipping level, for BPSK transmission over the AWGN channel and repetition codes of arbitrary rate, is of the form of the definition (without approximation) in (8). In such a situation the list length is either one, in the absence of a counterhypothesis, or two, in which case the exact LLR is known.

To illustrate the capabilities for our SLC approach, we consider the error probabilities for PAM and QAM modulation. Specifically, the symbol error probability for the maximum-likelihood detector in the case of pulse amplitude modulation (PAM) transmission over an AWGN channel with effective signal-to-noise ratio $\sqrt{\ell}\mathrm{SNR}_i$ is given by:

$$P_{s,PAM} = 2\left(1-\frac{1}{\sqrt{q}}\right) Q\left(\sqrt{\frac{3}{q-1}\sqrt{\ell}\mathrm{SNR}_i}\right). \quad (9)$$

Note that since one-dimensional PAM modulation is considered, we elect to use $\sqrt{\ell}=\sqrt{K}$ to compute (9), where K is the parameter for the representative K-best tree search detection scheme [7], [8]. Performance results support such a selection[3]. Plugging $P_{b,QAM}$ into (8) yields improved LLR values in the absence of a counter-hypothesis, relative to FLC.

Following standard QAM extensions of the PAM SER expression yields the QAM symbol error rate:

$$P_{s,QAM} = 1-(1-P_{s,PAM})^2 \quad (10)$$

[3]Another option, which yielded good results was to use γ_ℓ, the square root of γ_ℓ^2 as detailed in [14]

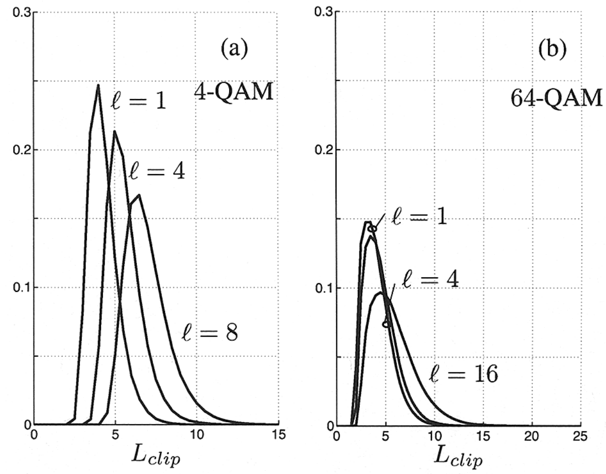

Figure 3. The pdf for L_{clip} for i.i.d Rayleigh fading for a 4×4 MIMO system using (a) 4-QAM and (b) 64-QAM transmission and $K=\ell=\{1,4,8\}$ and $K=\ell=\{1,4,16\}$, respectively. The maximum values for L_{clip} are not shown due to precision issues which force these values to be infinite.

from which the QAM bit error rate can be easily obtained as $P_{b,QAM} \approx P_{s,QAM}/L$ (and equivalent for the PAM BER). Plugging these error rate expressions back into (8) allows for an improved LLR clipping level, relative to fixed LLR clipping. Note that (8) is a general expression capable of working with a host of modulation and detection schemes, provided that proper treatment is given to the application of the effective SNR gain due to an increase in the list length.

The probability density function for L_{clip} is shown in Fig. 3 for a 4×4 MIMO system in i.i.d. Rayleigh fading using (a) 4-QAM and (b) 64-QAM transmission with $K=\ell=\{1,4,8\}$ and $K=\ell=\{1,4,16\}$ [7], [8], respectively. Results shown are for SNR values corresponding to a coded BER of 10^{-5} for the given list length using sorted MMSE preprocessing [15] (cf. Fig 4 and Fig 5). The plot was obtained using L_{clip} values found for over 400,000 distinct channel realizations and the approximation in (8) was used.

We conclude this section by providing a brief description of the complexity aspects for the SLC approach. Specifically, the complexity of the approach is the complexity required to compute (8) for each detection layer and channel realization. While the complexity associated with such a computation typically involves complicated calculations such as the $Q(\cdot)$ function and square root operation, as in (10), the practical complexity of SLC can be significantly reduced through the use of a table lookup, or approximation like the one in (8).

V. Results

We consider transmission over a spatially and temporally i.i.d. fading 4×4 MIMO channel, using 4- and 64-QAM alphabets. The information block size (including tail bits) is 9216 bits. Detection is performed based on the real-valued system model. Since tree search schemes with fixed (or tightly bounded) detection complexity benefit from the use of MMSE preprocessing, we employ unbiased MMSE detection [15]

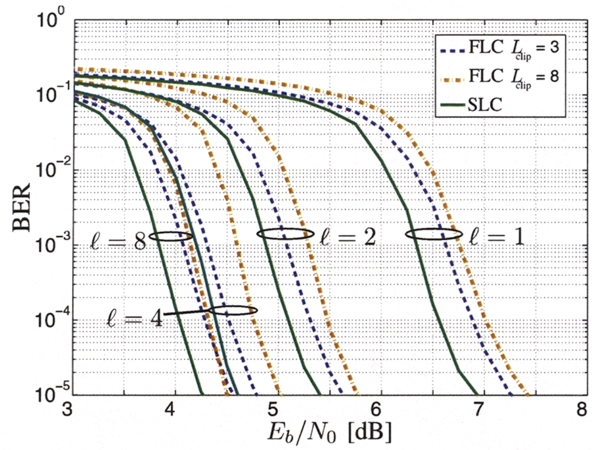

Figure 4. SNR-aware LLR clipping versus fixed LLR clipping for a 4-QAM coded non-iterative system for a 4×4 MIMO system under Rayleigh fading.

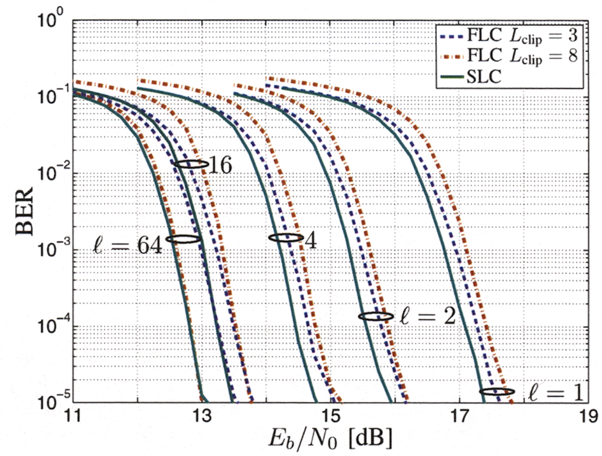

Figure 5. SNR-aware LLR clipping versus fixed LLR clipping for a 64-QAM coded non-iterative system for a 4×4 MIMO system under Rayleigh fading.

with all techniques. For coded transmission, we use a setup equivalent to the one in [1]: a rate 1/2 PCCC based on $(7_R, 5)$ convolutional codes using 8 internal iterations of logMAP decoding.

Fig. 4 provides a performance comparison of the proposed SLC approach and fixed-valued LLR clipping using the simulation setup just described in the case when there are no iterations between the detector and the decoder (i.e. only decoder iterations) and 4-QAM transmission. All results shown are obtained using the K-best Algorithm for the case when $K = \ell = \{1,\ 2,\ 4,\ 8\}$ and employ the approximation found in (8). In all cases, for the same detection algorithm (i.e. same value of ℓ), our SNR-aware approach outperforms the FLC approaches. As an example, when $\ell = 1$, the SLC approach outperforms the clipping of ± 8 proposed in [1] by 0.5 dB and the clipping of ± 3 prosed in [5] by 0.3 dB at a BER of 10^{-5}. Furthermore, the K-best for $\ell = 8$ shown in Fig. 4 is used to demonstrate that the SLC approach with $\ell = 4$ is roughly equivalent to the performance of the higher complexity $\ell = 8$ algorithm when using FLC.

Fig. 5 provides the same performance comparison between the SLC approach and fixed-valued LLR clipping, this time for the case of 64-QAM. Results are obtained for the K-best algorithm for $K = \ell = \{1,\ 2,\ 4,\ 16 \text{ and } 64\}$ and employ the approximation found in (8). Again, in all cases, for the same detection algorithm, the SLC approach outperforms the fixed LLR clipping approaches.

VI. Conclusions

In this contribution, we presented a scheme for computing the log-likelihood ratio clipping level for suboptimal MIMO list detection. This problem was framed as determining variable LLR clipping levels conditioned on both the CSI and the list length of the detector. QAM error rate performance over i.i.d. Rayleigh fading MIMO channels indicated that the proposed SNR-aware LLR clipping approach outperformed fixed LLR clipping schemes.

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