Improved Reconstruction of Channel State Information in 3GPP

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***Abstract*— The closed-loop transmit diversity technique is used to increase the capacity of downlink channel in multiple-input-multiple-output (MIMO) communications systems. The WCDMA standard endorsed by 3GPP adopts two modes of downlink closed-loop schemes based on partial channel information that is fed back from mobile to base station through a low-rate uncoded feedback bit stream. In this article, soft reconstruction techniques are utilized to improve the performance of Mode 1 of 3GPP, by taking advantage of the redundancy available in the channel state information. We propose several algorithms for efficient reconstruction of beamforming weights, which could be used instead of Antenna Weight Verification algorithms in the closed-loop system. The performance is examined within a simulated 3GPP framework in the presence of feedback error at different mobile speeds. It is demonstrated that the proposed algorithms have substantial gain over the conventional approach, for low to high mobile speeds.**

I. Introduction

[1] The increasing demand for internet and wireless services highlights the need for an increase of the capacity of the communication systems. Third generation of mobile communication, namely 3GPP [1] and 3GPP2 [2] have developed the WCDMA [3] and CDMA2000, respectively, to address the trend. The improvement of the downlink capacity is one of the main challenges of the 3G systems, and closed-loop techniques are known to have the potential to solve the problem. Transmit Adaptive Array [4] is a part of the 3GPP standard with two transmit antennas at the base station and one receive antenna at the mobile unit, which uses some Channel State Information (CSI) to beamform the transmit signal.

The feedback data is a low-rate stream of quantized CSI, which is uncoded. Hence, the scheme is sensitive to feedback error. Furthermore, the reconstruction scheme suggested by the standard is not efficient. Our focus here is on Mode 1 of 3GPP [5] which only feedbacks the phase information of the channel with a special quantization scheme. Mode 1 has a good performance at low mobile speeds, but it fails at higher speeds. Here joint source-channel techniques are used to improve the performance of Mode 1 of 3GPP, by taking advantage of the redundancy available in the CSI stream. We propose several algorithms for reconstruction of beamforming weights that can successfully deal with the feedback error. For the channel model, we consider independent Rayleigh fading channels and for simulating the mobile fading channel, Jakes Model is assumed and a modified Jakes fading generator [6] is used to produce stationary fading signals.

A. Effect of Feedback Error on the Performance

When an error occurs in the feedback channel, an incorrect antenna weight (i.e., beamforming weight) vector is applied at the transmitter, which leads to two consequences. First, the received signal power is smaller [7], because a non-optimum weight is applied. However, our simulations show that the performance degradation due to this effect is rather small. The second consequence is much more serious: Each time a feedback error occurs, the mobile station does not know the actual antenna weight vector that is applied at the base station. Since the mobile station obtains the dedicated (i.e., user-specific) channel estimate by combining the estimates for individual antennas from common pilots with the assumed weight vector used at the base station. This causes serious dedicated channel estimation error, which results in some error floor in the performance. To minimize the effect of the problem, a technique called Antenna Weight Verification (AV) [8] has been suggested. But this technique has some drawbacks. Applying an AV algorithm requires extra calculations at the mobile unit, and also

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requires certain dedicated preamble bits to be transmitted to all users. This complexity at the mobile unit could be limiting for complicated adaptive beamforming schemes. Therefore, there has been a tendency to find a substitute technique for AV, for example [9].

We introduce a new approach to improve the performance of the beamforming scheme, especially in presence of feedback error. Our approach help to solve both problems, and the performance is usually good enough that an AV algorithm is no longer needed, although, an AV could be used along with our algorithms if higher performances are required. Furthermore our proposed algorithms are accomplished mostly at the base station.

II. Efficient Reconstruction of Beamforming Weights

We preserve the framing structure and the quantization scheme of mode 1 of the 3GPP standard. Hence, $w_n^{(1)} = \left|w_n^{(2)}\right| = \frac{1}{\sqrt{2}}$, which keeps the transmit power constant ($w_n^{(m)}$, $m = 1, 2$, is the complex beamforming weight of the m-th transmit antenna at time n). However, we improve the reconstruction algorithm of $w_n^{(2)}$. I_n is the transmit symbol index, where $I_n \in \{0, 1, 2, 3\}$ corresponding to the quantized phase $\tilde{\phi}_n \in \{-\pi/2, 0, \pi/2, \pi\}$, respectively. Similarly, J_n is the received symbol. For an error-free feedback channel, $I_n = J_n$.

It is shown that following a minimum mean-squared error (MMSE) approach [10], we may write

$$\check{w}_n^{(2)} = \sum E\left[w_n^{(2)} \middle| \underline{I}_n^{n-\mu+1}\right] P\left(\underline{I}_n^{n-\mu+1} \,\middle|\, \underline{J}_n\right), \tag{1}$$

where $\underline{J}_n = [J_n \cdots J_0]$, and the summation is over all the possible μ-fold sequences of $\underline{I}_n^{n-\mu+1} = [I_n \cdots I_{n-\mu+1}]$ for a sufficiently large value of μ. As opposed to our assumption, there is no control on the amplitude of $\check{w}_n^{(2)}$ in (1). We introduce a lemma which gives us a tool to control the amplitude of the solution (See Appendix I). According to the lemma, we can calculate the needed antenna weight using the MMSE solution as

$$\hat{w}_n^{(2)} = \frac{1}{\sqrt{2}}\, \overline{E}\left[w_n^{(2)} \middle| \underline{J}_n\right] = \frac{1}{\sqrt{2}} \frac{\check{w}_n^{(2)}}{\left|\check{w}_n^{(2)}\right|}. \tag{2}$$

where $\overline{E}[w] = \frac{E[w]}{|E[w]|}$.

A. Weight-Codebook

Assuming the error-free case of equation (1), the codeword associated with the specified $\underline{i}_n^{n-\mu+1}$ is constructed as

$$w_{CB}^{(2)}\left(\underline{i}_n^{n-\mu+1}\right) = \frac{1}{\sqrt{2}}\, \overline{E}\left[w_n^{(2)} \middle| \underline{I}_n^{n-\mu+1} = \underline{i}_n^{n-\mu+1}\right], \tag{3}$$

where $w_n^{(2)}$ is the weight calculated from Co-phase feedback algorithm.

γ	0	1	2	3	4	5
$v = 1$	0.00	0.72	1.56	1.57	1.59	1.59
$v = 10$	0.00	0.72	1.32	1.33	1.35	1.35
$v = 25$	0.00	0.72	1.08	1.08	1.11	1.11
$v = 100$	0.00	0.71	0.73	0.73	0.75	0.76

TABLE I

Redundancies $R(\gamma)$ for different memory depths γ, and different mobile speeds

B. Trellis Structure

For capturing the redundancies in the feedback bitstream, we assume a Markov source of order γ. A trellis structure is set up based on the Markov model to exploit the redundancies. The states of the trellis are defined as

$$S_n = \underline{I}_n^{n-\gamma+1}. \tag{4}$$

It has been shown in Appendix II that there are $N_{states} = (\gamma + 1)\, 2^\gamma$ possible states. The trellis is specified by the probabilities of the state transitions, $P(S_n|S_{n-1})$, or equivalently by conditional probabilities $P(I_n|S_{n-1})$. The probabilities constitute the a priori information of the Markov model and can be calculated and stored as *probability-codebook*.

To find a proper value for γ, we examine the redundancies defined as $R(\gamma) = 2 - H(I_n|S_{n-1})$ [10], [11], where $H(I_n|S_{n-1})$ is the average conditional entropy. Table I shows typical redundancies for values of $\gamma = 1, 2, 3, 4$, for different mobile speeds in the range of our interest. As expected, it is observed that redundancies are higher for larger memory depths. However, for memory depths $\gamma \geq 2$, redundancies do not increase significantly, which means there is no significant residual redundancy left in the sequence after that depth. Hence, we use $\gamma = 3$ in our simulations, which is a small value of memory depth but can capture most of the redundancies. The resulting trellis, which has 32 states, is used by some of our algorithms.

By selection of γ, the trellis is specified. To implement (1), a proper value of μ is also required. Considering (3), the value of μ should be chosen to include the number of feedback symbols that carry some useful information about the weight-codeword. The redundancies are

Coding Rate	1/3
No. of Memory Elements	3
Constituent Encoders (Convolutional)	$G_{bwd} = 1 + D^2 + D^3$ $G_{fwd} = 1 + D + D^3$
Output Pattern (P/S)	$\cdots, X_k, Z_k, Z'_k, \cdots$

TABLE II

TURBO CODE PARAMETERS

limited to a certain depth similar to the selection of γ, hence, a larger μ does not improve the weight-codebook. Therefore, the proper value of μ should be close to the selected value of γ.

C. Proposed Algorithms

Its notable that our algorithms decide on the needed antenna weight *without any delay* as required.

1) Non-Linear Weight Algorithm (NLW): This algorithm only uses the weight-codebook as $\hat{w}_n^{(2)} = w_{CB}^{(2)}\left(\hat{\underline{i}}_n^{n-\mu+1}\right)$, where $\hat{\underline{i}}_n = \underline{j}_n$.

2) Normalized-MMSE Algorithm (NMMSE): Equation (2) is implemented using the trellis. A BCJR algorithm [12] is utilized to find the probabilities of each state recursively [10], to calculate (1).

3) Soft-NMMSE: In earlier algorithms, it is assumed that the received feedback information is quantized. However, using the soft feedback data can potentially improve the performance of the aforementioned algorithms [13]. In these methods, it is assumed that instead of hard-decided bits of feedback data, soft-output (noisy) feedback symbols are available. In other words, instead of a BSC channel, an AWGN channel is assumed for the feedback data.

III. RESULTS

A. Simulation Parameters

For testing and comparison of the algorithms, we have simulated a communication system similar to the downlink of FDD WCDMA [3]. Fig. 1 represents our complete system, including the channel coding and the details of the feedback system. For channel coding, we have used Turbo coding [14]. The parameters of the Turbo code structure suggested by 3GPP are shown in Table II.

For the channel interleaving, we apply a randomly-generated interleaver with the same length as that of the frame, suggested by 3GPP. Because of this interleaver, the coding scheme faces a channel which is closer to an i.i.d. channel with respect to the case without any channel interleaving.

Carrier Frequency	2.15 GHz
Modulation	QPSK
Transmitter	2 Antennas
Receiver	1 Antenna
Data Rate	15000 bps
Feedback Rate	1500 bps
Channel Model	Modified Jakes
Channel Coding	Turbo Code
Code Rate	$1/3$
Frame Length	300 (20 mSec)
Bit Interleaving	One Frame
Channel Estimation	Ideal
Power Control	No

TABLE III

SIMULATION PARAMETERS

In 3GPP systems, there is a power control scheme intended to eliminate the slow fading. Therefore, we assume that slow fading is compensated by power control. Table III is a summary of the parameters of our simulations. The other parameters, that may change in different simulations, are specified in each case. It is also assumed that no antenna verification is used with the closed-lop algorithms, unless stated otherwise.

IV. SIMULATION RESULTS

For the feedback error rate, a 5 percent error is usually used as a typical case in a 3GPP framework. As a sample, Fig. 2 shows the FER performance of the algorithms versus transmit SNR (E_b/N_0), for mobile speeds of $v = 1, 5, 25$ and 100 kmph, and 5 percent feedback error. The performances are examined at low to high mobile speeds. The gains at different mobile speeds and different SNR's ar observable. Regarding the relative gains, Fig. 3 shows the required SNR for a target FER of 0.005, versus the mobile speed. The plot includes the standard algorithm with an ideal AV (standard-IAV), i.e., assuming that standard receiver knows the exact beamforming weights applied at the transmitter. It can be observed that our approach decreases the required transmit power by about 2 dB for low to high mobile speeds.

A. Conclusion

The closed-loop transmit diversity, which uses a combination of transmit diversity and channel feedback, is recognized as a promising scheme to achieve high data rates in mobile communications. In 3GPP systems, however, the performance of the closed-loop scheme is limited, which is the result of the rate limit and error in the feedback channel and also the non-optimum

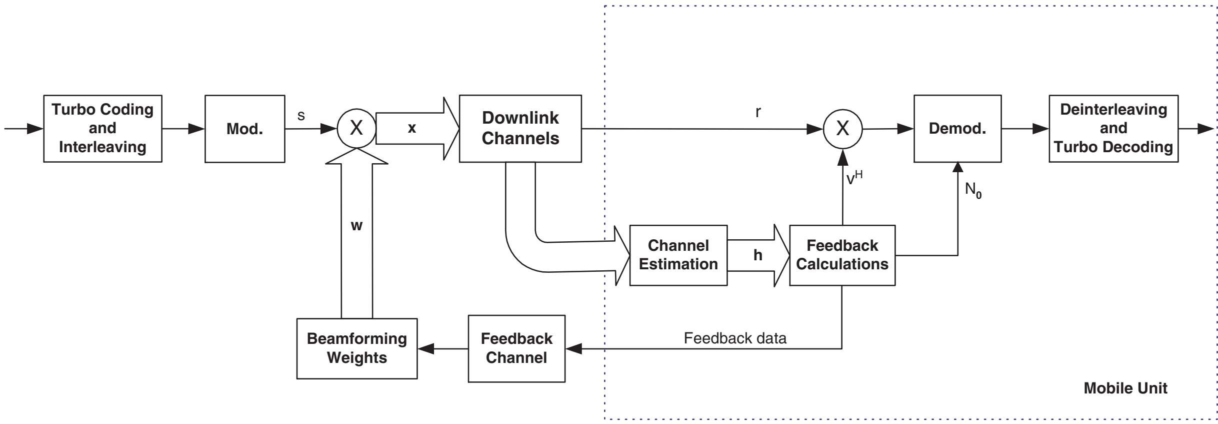

Fig. 1. Block diagram of our feedback system

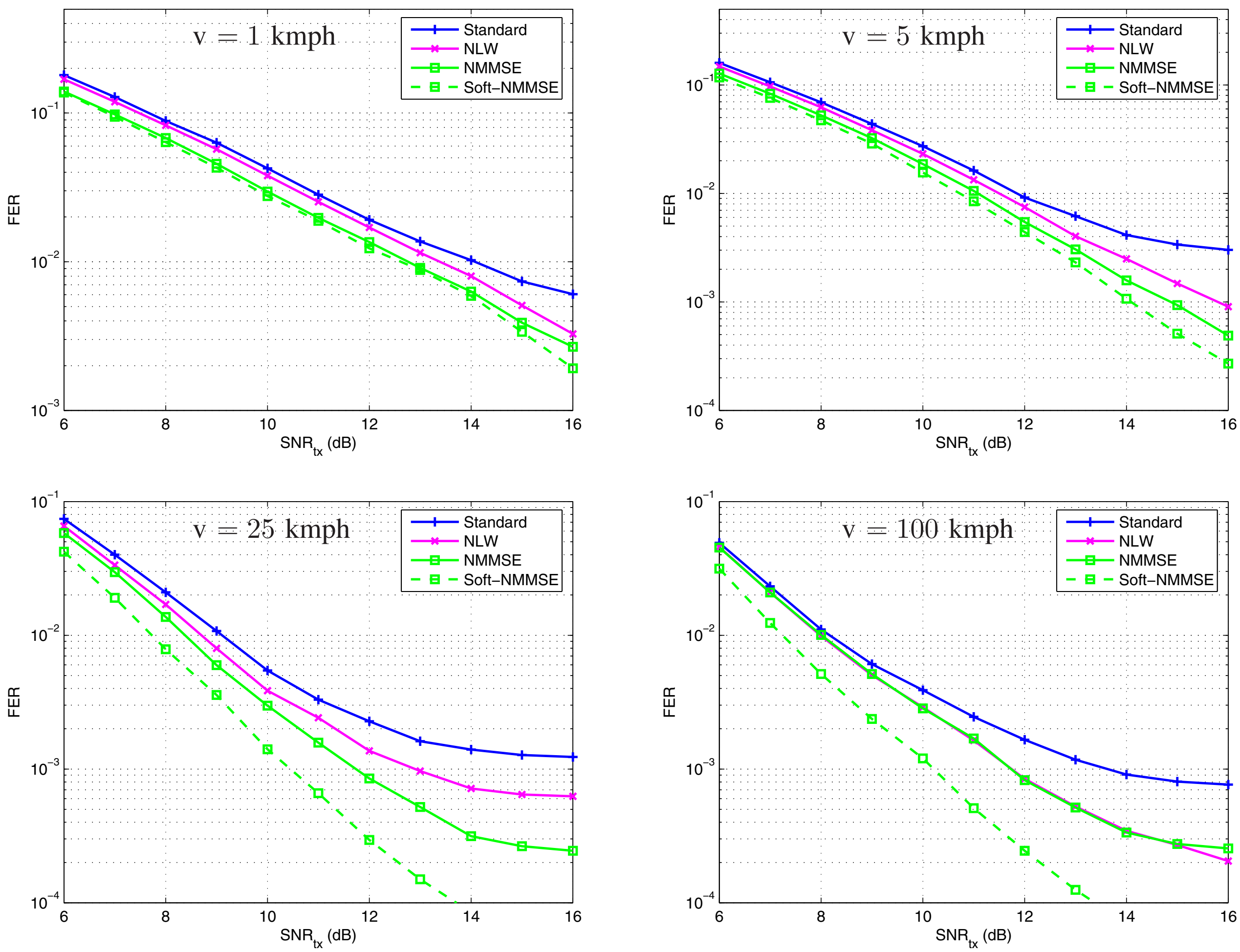

Fig. 2. FER in 5 percent feedback error

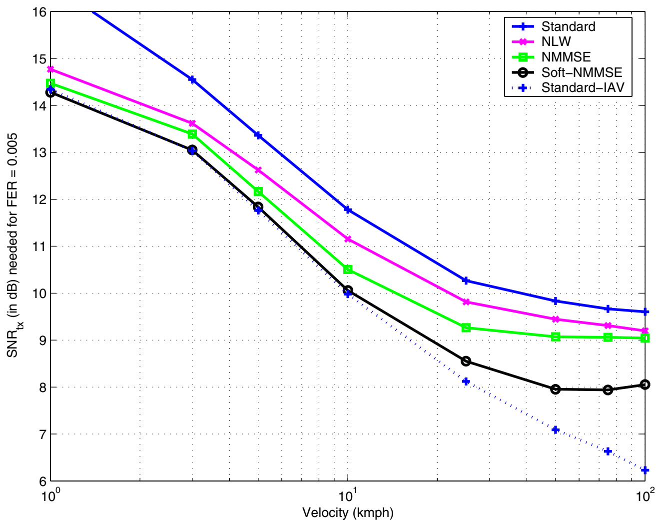

Fig. 3. Required SNR for FER=5e-3

weight reconstruction algorithm used in the standard. In this article, we propose an approach to improve the performance of the closed-loop system in the presence of feedback error, without changing the structure of the standard. We have introduced a number of algorithms to improve the performance of mode 1 of 3GPP. It has been shown that our algorithms can provide significant gains over the 3GPP standard algorithm at all mobile speeds. The performance of our algorithms, which do not need any preamble, are usually good enough such that an AV algorithm is no longer required. Though, an AV could be used along with our algorithms for further improvement.

Appendix I: Lemma

Lemma 1: If $\hat{w}$ is the MMSE estimation of a complex variable w, given $|w| = \beta$ and $|\hat{w}| = \alpha$, then

$$\hat{w} = \alpha \, \overline{E}[w], \tag{5}$$

where $\overline{E}[w] = \frac{E[w]}{|E[w]|}$ is the normalized MMSE solution.

Proof: We should estimate the $\hat{w} = \alpha e^{j\hat{\phi}}$ using the random variable $w = \beta e^{j\phi}$, where α and β are real positive constants. In the MMSE sense, it turns out to minimizing the following criterion

$$\begin{aligned} & E\left[|w - \hat{w}|^2\right] & (6) \\ = \; & \alpha^2 + \beta^2 - 2\alpha\beta \, E\left[\cos(\phi - \hat{\phi})\right] & (7) \end{aligned}$$

or maximizing

$$\begin{aligned} & E\left[\cos(\phi - \hat{\phi})\right] & (8) \\ = \; & E[\cos\phi]\cos\hat{\phi} + E[\sin\phi]\sin\hat{\phi} & (9) \end{aligned}$$

which results in

$$\tan\hat{\phi} = \frac{E[\sin\phi]}{E[\cos\phi]}, \tag{10}$$

where

$$\begin{cases} E[\cos\phi] \geq 0 & \Rightarrow \; -\frac{\pi}{2} < \hat{\phi} \leq \frac{\pi}{2} \\ E[\cos\phi] < 0 & \Rightarrow \; \frac{\pi}{2} < \hat{\phi} \leq 3\frac{\pi}{2} \end{cases} \tag{11}$$

Finding the phase information of $\hat{w}$ from (10), it is easy to show that in both cases of (11),

$$e^{j\hat{\phi}} = \frac{E[e^{j\phi}]}{|E[e^{j\phi}]|}. \tag{12}$$

Therefore,

$$\hat{w} = \alpha \, \overline{E}[w]. \tag{13}$$

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