Uplink Single-User MIMO for 3GPP LTE

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Abstract – **In this paper a multiple-input multiple-output (MIMO) scheme proposed for a single carrier frequency division multiple access (SC-FDMA) system is described. Transmit and receive algorithms are developed for SC-FDMA MIMO systems. A spatial precoding closed-loop transmit beamforming (TxBF) algorithm for SC-FDMA MIMO system which uses singular value decomposition (SVD) at the transmitter and linear minimum mean square error (LMMSE) equalization at the receiver is described. Quantization and feedback schemes to support closed-loop operation are discussed. The PAPR effects and a mitigation approach are presented. Simulation results are shown for throughput vs. signal-to-noise ratio (SNR). The performance for the new TxBF algorithm is compared to Single-Input-Multiple Output (SIMO). Also, single and dual-codeword performance is compared.**

Index Terms— **MIMO, beamforming, pre-coding, SC-FDMA, diversity, space time coding, space frequency coding, feedback, PAPR.**

I. INTRODUCTION

MIMO is considered essential for 3GPP E-UTRA Long Term Evolution (LTE) to provide high data rate and increased system capacity for the OFDMA downlink. It is also desirable to use MIMO for the SC-FDMA uplink for the same reasons. As stated in [1], the E-UTRA should support an instantaneous uplink peak data rate of 50Mb/s within a 20MHz uplink spectrum allocation (2.5 bps/Hz). Although theoretically a UE with a single transmitter can almost achieve 50 Mbps using 16QAM, it has been concluded that at least 2x2 MIMO is necessary to realistically achieve the required uplink throughput [2], [3]. It has also been shown that to achieve the highest throughput in uplink transmission the use of precoding is a necessity [4], [5]. An additional advantage to having two transmitters on the UE is the possibility to use beamforming to enhance Multi-User MIMO [6] and also better transmit diversity schemes such as Space Time/Frequency Decoding (ST/FD).

We envision two transmitters to be used more likely in a laptop, data card or possibly PDA-type device rather than in a handset. The user benefit would be substantially higher uplink data rate for up-loading pictures, music or video, interactive gaming and other large file transfers. The impact on the device in terms of baseband complexity and RF size, power drain and cost will be discussed later.

This paper discusses these aspects of a proposed method for uplink single-user MIMO:

- Architecture
- Precoding
- Feedback
- PAPR mitigation

Additionally, selected simulation results are provided.

II. SYSTEM OPERATION AND DESCRIPTION

A. Architecture

Figure 1 shows a transmitter block diagrams for a double codeword (DCW) configuration of uplink MIMO using precoding with dual transmit chains. In the case of a single codeword (SCW) the coded data is split into parallel streams, each with a (possibly) different modulation. Figure 2 shows a dual-antenna receiver block diagram for the DCW case. A single decoder would be used in the SCW case. The precoder matrix codeword index is assumed to be fed back from the base station (i.e., the eNodeB in LTE terminology) to the UE.

B. Precoding

The precoding is based on transmit beamforming (TxBF) using, for example, eigen-beamforming based on single-value decomposition (SVD). While SVD is optimal, other algorithms may be used by the eNodeB.

For precoding using eigen-beamforming the channel matrix is decomposed using a SVD or equivalent operation as

$$H = UDV^H \tag{1}$$

where H is the channel matrix. The 2-D transform for spatial multiplexing, beamforming, etc. can be expressed as

$$x = Ts \tag{2}$$

where *s* is the data vector and *T* is a generalized transform matrix. In the case when transmit eigen-beamforming is used, the transform matrix T is chosen to be a beamforming matrix V which is obtained from the SVD operation above, i.e., T = V. This is similar to eigen-beamforming for OFDMA, modified to apply to SC-FDMA.

Because the SVD operation results in orthogonal streams, the eNodeB can use a simple LMMSE receiver. It can be expressed as

$$R = R_{ss}\tilde{H}^H(\tilde{H}R_{ss}\tilde{H}^H + R_{vv})^{-1} \tag{3}$$

where R is the receive processing matrix, R_{ss} and R_{vv} are the correlation matrices for the signal and nosie, respectively, and $\tilde{H}$ is the effective channel matrix which includes the effect of the V matrix on the estimated channel response. In figure 2 the precoder block at the eNodeB produces the effective channel matrix at the UE using the last quantized precoder matrix sent from the eNodeB to the UE.

C. Feedback

One approach to feeding back the pre-coding matrix employs a codebook-based MIMO precoding scheme using combined differential and non-differential feedback [7].

1) Non-differential feedback

Jacobi rotation is used to perform the matrix diagonalization. The channel correlation matrix can be decomposed into

$$R \equiv H^H H = VD^2V^H \quad (4)$$

Diagonalizing the channel response matrix H to find the eigen-matrix V is equivalent to diagonalizing the channel correlation matrix R. Jacobi rotation is used to perform the matrix diagonalization of the channel correlation matrix such that

$$D^2 = J^H RJ \quad (5)$$

The Jacobi rotation matrix J can be used as a precoding matrix. The Jacobi rotation or transform matrix for 2 x 2 MIMO configuration is represented as

$$J(\hat{\theta},\hat{\phi}) = \begin{bmatrix} \cos(\hat{\theta})e^{j\hat{\phi}} & \sin(\hat{\theta})e^{j\hat{\phi}} \\ -\sin(\hat{\theta}) & \cos(\hat{\theta}) \end{bmatrix}$$

2) Differential feedback

Differential feedback uses iterative Jacobi transforms. For feedback instance n, the Jacobi rotation is represented by

$$J(n)^H R(n)J(n) = D^2 \quad (6)$$

For the next feedback instance n+1, the Jacobi rotation is

$$J(n)^H R(n+1)J(n) = \tilde{D}^2 \quad (7)$$

When the channel changes, $\tilde{D}^2$ is not diagonal. The precoding matrix or Jacobi rotation matrix needs to be updated for correct diagonalization. Call J the differential precoding matrix that represents the delta of the feedback update which is sent back to the transmitter from the receiver. The previous precoding matrix J(n) is updated to obtain the next precoding matrix J(n+1) by multiplying the previous precoding matrix with the differential precoding matrix,

$$J(n+1) = J(n)\cdot J \quad (8)$$

The differential feedback J can be computed at the receiver from the previous precoding matrix J(n) and the current precoding matrix J(n+1) by

$$J = J(n)^H J(n+1) \quad (9)$$

3) Combined differential and non-differential feedback

Combining differential and non-differential feedback can reduce feedback overhead and improve performance. Differential feedback can be reset every N TTIs, where a TTI is defined as 0.5 msec, or every certain period of time to avoid error accumulation due to feedback erasures or error propagation due to differential processing. At each reset non-differential feedback is used.

Two codebooks are used for the combined differential and non-differential feedback. The codebook used for differential feedback concentrates on the origin of the (θ,ϕ) plane while the codebook for non-differential feedback is uniform with codewords evenly distributed. A codebook consisting of eight codewords which requires three (3) feedback bits for quantization is used for non-differential feedback, while four codewords is used for differential feedback which requires less feedback bits (2 bits). For an SCME-C [1] channel the codebook can be based on averages over 6 resource blocks (RBs) where a RB is defined as a block with 25 subcarriers.

Figure 3 shows the performance of MIMO precoding using differential feedback and feedback delay for an SCME-C channel and vehicle speed of 3 km/h. The simulation parameters assumed are given in Table 1. It is shown that about 0.2 dB degradation is observed for feedback delay of 2 TTIs and 1 dB degradation for feedback delay of 6 TTIs with respect to the performance of no feedback delay. The combined performance degradation for 2-bit quantization and feedback delay is about 1.2 dB and 2 dB for feedback delay of 2 and 6 TTIs, correspondingly, with respect to ideal precoding with no quantization and no feedback delay.

D. PAPR mitigation

SC-FDMA with precoding such as TxBF will increase the peak-to-average power ratio (PAPR) because each transmit signal becomes a composite signal due to spatial processing. Expressed in terms of a transmit signal in the time domain, applying precoding in the frequency domain is equivalent to a convolution and summation of the data symbols in the time domain. Thus precoding will increase the PAPR of the composite transmitted signal.

1) PAPR of Uplink MIMO Precoding Signal

Figure 4 illustrates the PAPR characteristics of a MIMO precoding signal with 16QAM and QPSK in each stream. Without averaging of the channel estimates and quantization of the precoding matrix, the MIMO precoding signal has 1.5 ~2 dB higher PAPR with respect to single antenna transmission. Also plotted are cases where the channel estimates are averaged over 50 continuous subcarriers (750 kHz), and where direct-quantization of the precoder matrix V using 3 bits (1 bit for amplitude and 2 bits for phase information) was performed. Each of these is shown to reduce the PAPR. When both averaging of the channel estimates and quantization are considered, the PAPR for MIMO precoding decreases by 0.8 ~ 0.9 dB, and has about 0.5 dB higher PAPR for 16QAM with respect to single antenna transmission.

2) PAPR Reduction by Symbol Amplitude Clipping

One common way to reduce PAPR is to limit or clip the peak amplitude of the transmitted symbols. The problems associated with clipping are in-band signal distortion and generation of out-of-band signals. Because SC-FDMA modulation spreads the information data across all the modulated symbols, in-band signal distortion is mitigated when an SC-FDMA symbol is clipped.

Figure 5 shows the CCDF of symbol PAPR with clipping at various levels. With 7 db clipping less than 1 % of the symbols are clipped. Note that even with as much as 3 dB PAPR clipping, only about 10% of the modulated symbols are clipped.

Clipping will generate both in-band and out-of-band frequency components. Figure 6 and 7 shows the FER performance and frequency spectrum of the clipped signals by power spectral density (PSD). For the PSD calculation, a Hanning window was used with 1/4 of window overlapping. For 7 dB PAPR clipping, the spectrum is almost the same as that of the original signal. More pronounced out-of-band components arise when 5 or 3 dB PAPR clipping is used. It is clear that 7 dB clipping has minimal impact on the power spectrum.

III. Simulation Results

This section presents selected simulation results for SU-MIMO. A comparison between SU-MIMO and SIMO is discussed first, followed by a comparison of the performance for single and double codeword SU-MIMO.

A. *Simulation assumptions*

The simulation parameters assumed are given in Table 1. In Table 2 we list the achievable throughputs for various selections of the MCS for each spatial stream. It is worth noting that the maximum achievable throughput using a double codeword and practical code rates in 5 MHz is 19.968 Mbps, which scales to 79.87 Mbps in a 20 MHz bandwidth, and has a spectral efficiency of 4 bps/Hz. SIMO, on the other hand, is limited to 10.75 Mbps in 5 MHz, a spectral efficiency of 2.15. Therefore, SU-MIMO can almost double the uplink data rate compared with SIMO.

B. *Comparison of SU- MIMO to SIMO*

Figure 8 shows a comparison of double codeword performance for SU-MIMO to SIMO for the high data throughput SNR regions. When the SNR is 24 dB the maximum achievable throughput is approximately 19 Mbps, and when the SNR is greater than 26 dB the achievable throughput is approximately 19.97 Mbps. From this comparison it is worth noting that using SIMO the maximum achievable throughput is 10.5 Mbps at an SNR of 20 dB.

C. *Comparison of SU- MIMO with single and double codeword*

This section presents a comparison of the performance for single and double codeword using uplink Precoding MIMO for 2 antennas at the UE and eNodeB with the SCME-C channel. Because HARQ was not simulated, the same code rate was used for both SCW and DCW in order to compare them fairly. Also, it is impractical to use the same modulation for SCW for both streams when using Precoding, so only combinations of QPSK and 16QAM are shown. Therefore, the higher throughput achievable with DCW is not shown.

Figure 9 shows a comparison of the performance for single and double codeword using uplink Precoding MIMO for 2 antennas at the UE and eNodeB with SCME-C channel. Notice that the DCW achieves a higher throughput at lower SNRs, while the opposite is true at higher SNRs where the SCW performs better than DCW. The difference is more pronounced at the highest data rates where a 3 dB difference can be seen. Eventually, since equal modulation and coding was used, both schemes reach the same maximum throughput, almost 14 Mbps in 5 MHz for the highest MCS simulated.

The reason that DCW performs better at lower SNR is because the upper eigenmode has higher SNR than the total system SNR. Therefore at low SNR that stream contributes some successful transmissions while the lower stream generally does not. However, at higher SNR the lower stream still has relatively high BLER which tends to reduce the total throughput for DCW. But, in the case of SCW, the upper stream protects the lower stream because the coding covers both streams. This results in an overall lower BLER for SCW at higher SNRs.

From these results it may be concluded that very high uplink spectral efficiency, about 2.8 bps/Hz, can be achieved using either method. However, DCW can achieve a higher spectral efficiency, about 4 bps/Hz, because it can use 16QAM with different code rates on each stream, whereas SCW must use a single code rate and different modulations.

IV. Conclusions

These conclusions on uplink SU-MIMO for SC-FDMA can be reached:

- Precoding at the UE can be based on SVD or a comparable algorithm performed at the eNodeB. For an SCME-C channel the codebook can be based on channel averages taken over several, e.g. six adjacent RBs.
- Feedback of the precoding matrix index can be performed efficiently using combined differential and non-differential feedback. Representative feedback parameters are 2 bits every 6 RBs sent every 6 TTIs, or a maximum of 1333 bps for 24 RBs in 5 MHz. Since the equivalent maximum data rate is 19.968 Mbps, the feedback efficiency is very high.
- PAPR due to transmit beamforming can be effectively mitigated using amplitude clipping. It was shown that clipping at 7 dB above the average power caused negligible increased BLER and out-of-band spectral growth.
- Simulations showed that SU-MIMO can almost double (186 %) the uplink data rate compared with SIMO.

V. References

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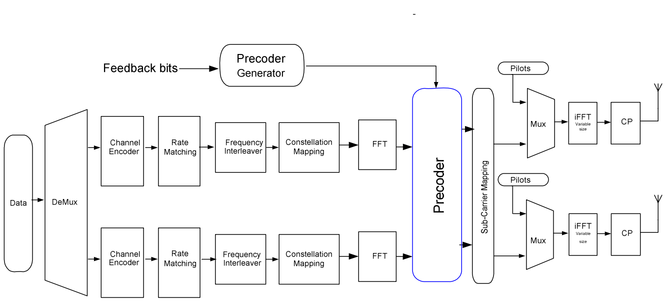

Figure 1. Double Codeword Transmitter Block Diagram

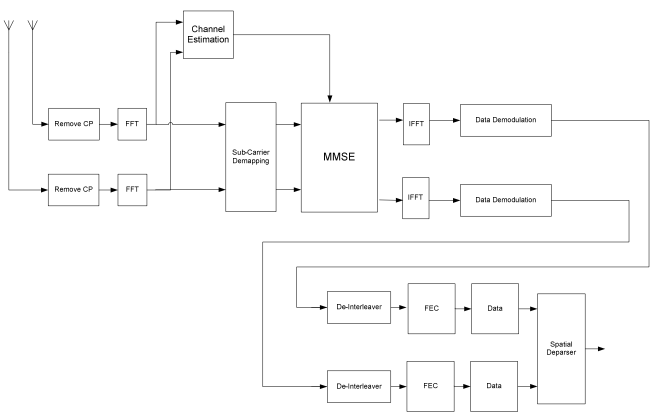

Figure 2. Double Codeword Receiver Block Diagram

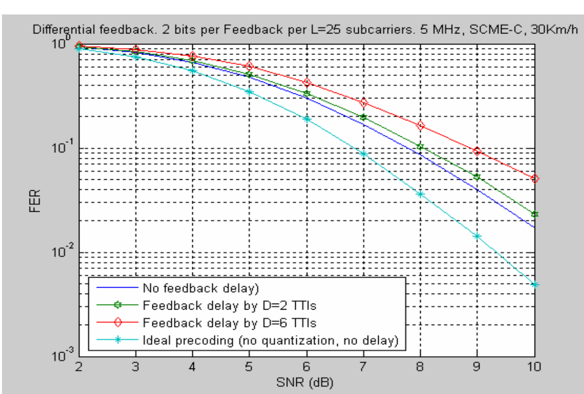

Figure 3. Performance of MIMO precoding using differential feedback and feedback delay

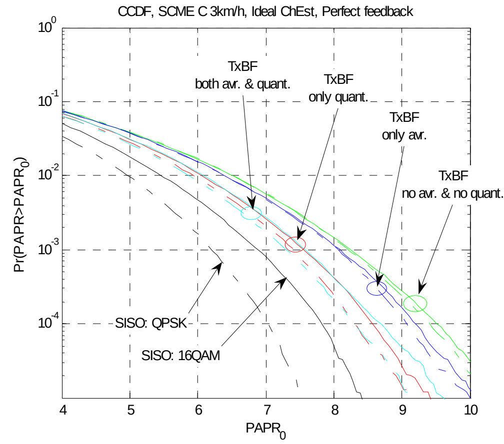

Figure 4. PAPR of 2x2 MIMO precoding.

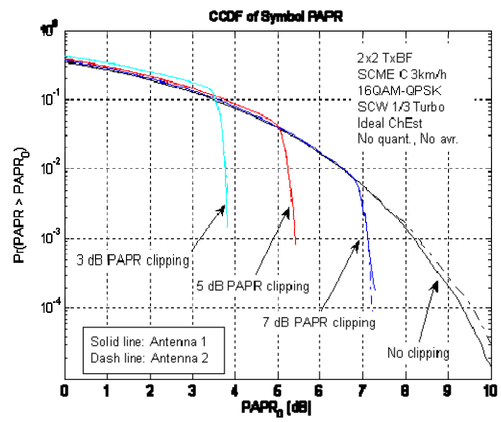

Figure 5. CCDF of symbol PAPR.

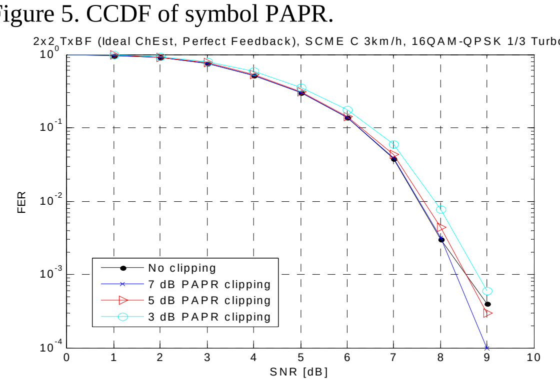

Figure 6. FER performance of clipping.

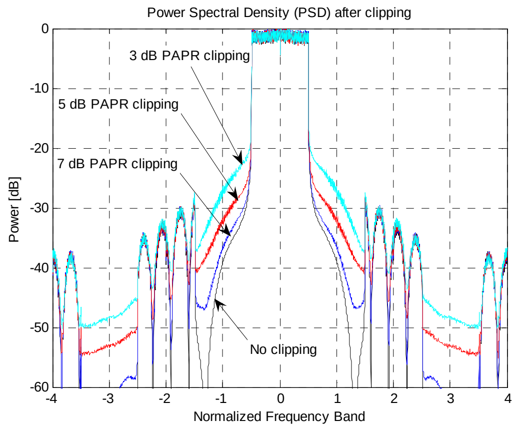

Figure 7. PSD after clipping.

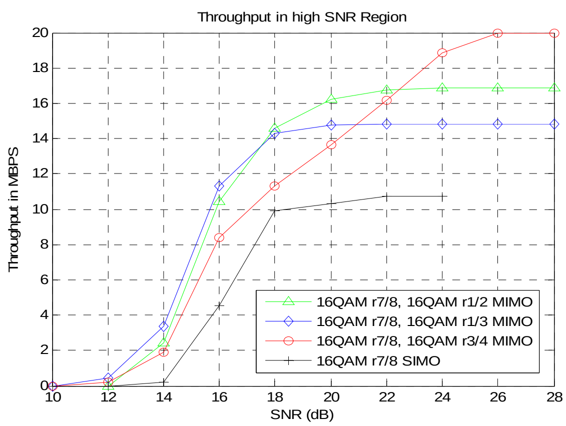

Figure 8. Throughput comparison of precoding and SIMO at higher SNR regions (TU6)

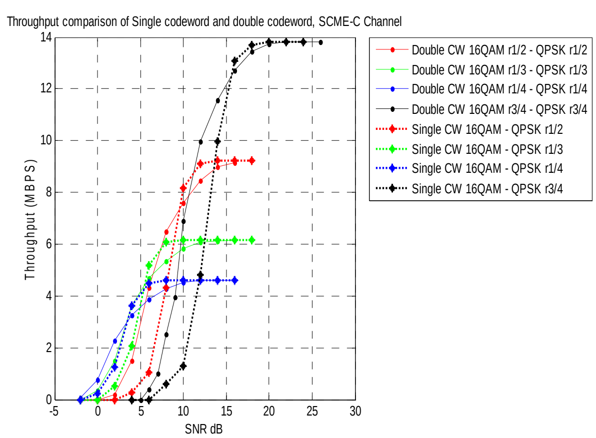

Figure 9. Throughput Comparison of Single and Double Codeword for UL precoding MIMO for SCME-C channel

Table 1. Simulation parameters

Parameter	Assumption
Carrier frequency	2.0 GHz
Sampling frequency	7.68 MHz
Transmission bandwidth	5 MHz
TTI length	0.5 ms
Number of long/short blocks per 0.5ms TTI	6/2
Number of occupied subcarriers	300
FFT block size	512
Number of used subcarriers for data	256
Cyclic Prefix (CP) length	5.078 us (39 samples)
Channel models	Typical Urban (TU6) and SCME-C
Antenna configurations	2 x 2 (MIMO)
Moving speed	3 km/h
Data modulation	QPSK and 16QAM
Channel coding	Turbo code with soft-decision decoding
Equalizer	LMMSE
Feedback error	None (Assumed Ideal)
Channel Estimation	Ideal channel estimation

Table 2. Achievable data rates and spectral efficiency for double codeword in a 5 MHz bandwidth

MCS	Achievable data rate (Mbps)	Spectral efficiency (bps/Hz)
16QAM r7/8– 16QAM r3/4	19.9680	3.99
16QAM r7/8– 16QAM r1/2	16.8960	3.38
16QAM r7/8– 16QAM r1/3	14.8480	2.97
16QAM r5/6 – QPSK r1/8	11.08	2.22
16QAM r5/6 – QPSK r1/2	10.752	2.15
16QAM r3/4 – QPSK r1/6	10.24	2.05
16QAM r1/2 – QPSK r1/3	8.192	1.64
16QAM r1/2 – QPSK r1/6	7.168	1.43
16QAM r1/3 – QPSK r1/8	4.864	0.97
16QAM r1/4 – QPSK r1/8	3.840	0.77